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AUDIO CYCLOPEDIA

by
HOWARD M. TREMAINE

the most comprehensive
and authoritative reference volume
on audio ever published

covers every phase of the subject, including
the latest solid-state and integrated circuits

AUDIO

CYCLOPEDIA

This book is dedicated to my wife, Sandy, for her patience, understanding, and help through two editions.

AUDIO CYCLOPEDIA

by

Howard M. Tremaine, D.Sc., FAES

Howard W. Sams & Co., Inc.
4300 WEST 62ND ST. INDIANAPOLIS, INDIANA 46268 USA

SECOND EDITION
EIGHTH PRINTING—1979

International Standard Book Number: 0-672-20675-7
Library of Congress Catalog Card Number: 77-82885

Printed in the United States of America.

Preface

In compiling the second edition of the *Audio Cyclopedia*, the author is reminded of the statement of one critic, that it was too great an undertaking for one man. However, no group has come forward to compile a revision. The first edition was an attempt to compile an accumulation of more than 30 years of technical data into one book, which the author had entitled "The Audio Engineer's Handbook." The publisher suggested that it had exceeded the handbook category, and recommended the present title, *Audio Cyclopedia.*

Since the first edition, the growth in audio engineering has made the revision a mountainous undertaking, requiring almost two years of unrelenting endeavor. New standards have been established, and the development of solid-state and integrated-circuit devices has opened a wide vista not conceived ten years ago.

The pattern of this edition closely follows the scheme of the first edition. However, the book has been shortened to 25 sections. Section 26 of the first edition (Stereophonic Recording and Reproduction of Disc Records) has been absorbed in other sections, as stereo is a general part of the overall picture and need not be treated as a separate and distinct subject. The first edition received wide acceptance and a large market for it developed outside the United States. Therefore the author has included in this edition many items developed and manufactured in countries other than the United States.

The question was once asked, "Who remembers to bless the hands that milked the cow, when he eats the cheese?" As we enjoy the fruits of today's progress, we would do well to remember men like Leon Scott, who in 1857 first scribed a sound wave, using a stiff bristle attached to a diaphragm actuated by a horn. Between the years of 1858 to 1862, Koenig devoted a great part of his time to improving on Scott's invention and presented the results of his labors in London in 1862, which he termed "Phonograms." However, it remained for men like Edison, Berliner, and Johnson to invent and develop the methods culminating in the cylindrical and disc records so well known today. Valdemar Poulson, the Danish Edison, might be called the father of present-day magnetic recording, and certainly Lord Rayleigh is the father of acoustics and its many phases, with Professor Wallace C. Sabine of Harvard University responsible for modern acoustics. It is of interest to note that in 1711 John Shore discovered the tuning fork, and in 1908 G. W. Pierce described in a paper, "A Simple Method of Measuring the Intensity of Sound," based on his work in measuring sound intensities in auditoriums and of train whistles; thus was born the first sound-level meter. All these men and many more, leave us forever in their debt.

Those of us who have been closely associated with the industry some 40 years more or less, view with pride the progress that has been made. It is gratifying to observe some of our original thinking and experiments becoming a reality with modern materials, components, and facilities. With this in mind, the author has included some obsolete devices, which may still be in daily use, but mainly to offer information to the experimenter. Examples of this are the electret, which has been around for years but is now being utilized as a microphone, and the fuel cell, discovered by Sir William Grove in 1839, which is finding its greatest application in the space age.

The author is deeply indebted to the many engineers who have supplied information for compiling this book, and to the many manufacturers who have freely furnished photographs and technical data required to make this work a reality. Selection of illustrations resolved into the problem of choosing items that would be typical examples of a device, or in some instances where a manufacturer had included features in an instrument that made it unique or of particular interest. The final selection was often difficult; today's market offers a wide variety of good equipment.

The author wishes to pause and personally thank the following: Mr. Robert O. Cook, of Walt Disney Productions; Mr. Arthur C. Davis, of Altec-Lansing; Mr. Wallace Hamilton, of Trans-Canada Laboratories, Vancouver, B.C.; Mr. Jack V. Leahy, of Radio Corp. of America; Mr. Frank E. Pontius, of Westrex Corp.; Mr. Michael Rettinger, acoustical engineer; Mr. Loren L. Ryder, of Ryder Sound Services; Mr. Ross H. Snyder, of Hewlett-Packard Co.; and Mr. Waldon O. Watson, of University City Studios; to mention but a few who made generous contributions. Also, the author is indebted to the following societies for their journals and standards: Acoustical Society of America; Audio Engineering Society; Electronic Industries Association; Institute of Electrical and Electronic Engineers; Institute of High Fidelity; National Association of Broadcasting; Society of Motion Picture and Television Engineers; and the United States of America Standards Institute for their much-used reference sources. In addition, thanks to the publishers of the following: *Audio; Broadcast Engineering; db, the Sound Engineering Magazine; Electronic Instrument Digest; Electronics; Electronics and Communications* (Canada); *Electronics World; Radio Electronics;* and *Wireless World* (England) for their cooperation. Special thanks are offered to my friend and personal physician, Dr. Ronald McAdams, who indirectly contributed to this work.

The first edition of the *Audio Cyclopedia* has often been referred to as "The Sound Man's Bible." If this name fits, the author offers this edition with the conviction that it will serve even better.

HOWARD M. TREMAINE

Preface to First Edition

This book was prompted by the response to a series of lectures presented to naval personnel by the author, in 1945, on "High Fidelity Sound Systems." It is intended as a practical engineering guide for the individual who has an understanding of electronics and desires to apply that knowledge to the recording and reproducing of sound.

In the pursuit of fidelity in sound recording and reproduction, numerous problems have been encountered. While many of the problems have been overcome to a marked degree and much credit is due the research workers whose efforts have thus far advanced the art, a large number of problems still remain to be solved.

The term "high fidelity" is often abused and loosely used, and equipment bearing such a title assumes a great responsibility. Webster defines fidelity as "exactness, as in a copy." High quality sound systems require careful design and adequate test equipment for their proper installation, adjustment, and maintenance.

The ultimate reproduction of sound waves is affected by many factors, each contributing its own particular type of distortion, which may be acoustical, electrical, or mechanical. This distortion may manifest itself as frequency or pitch variations, harmonic distortion, hum, or noise, and any or all of these factors may be interposed between the original recording and the listener.

An attempt has been made to compile sufficient data into one book, in the form of questions and answers, to synthesize sound systems generally, together with their components and allied equipment, and the accepted methods of installation, testing, and operation. Section 22, pertaining to test equipment, describes many pieces of apparatus of different manufacturers. However, no evaluation of their individual merits has been made, but their characteristics and operation are so described that the audio engineer may readily determine which equipment will best meet his needs.

The various sections of this book cover the fundamental concepts of sound waves and treat of their complexity, their behavior, and the equipment associated with their recording and reproduction. The illustrations offer workable circuits with practical valnes. Section 3 discusses the devices used in the motion picture and recording industries for maintaining a constant speed and the control apparatus necessary to achieve this end. Sections 5, 6, and 7, "Attenuators," "Equalizers," and "Wave Filters," include numerous devices and circuits used for motion picture sound recording and reproduction, which are equally adaptable to other types of recording and reproduction.

The improvement of sound mixers, from a simple two-position to an elaborate hybrid-coil design for motion picture rerecording, is described in Section 9. Audio amplifiers are dealt with extensively in Section 12.

Motion picture projection equipment is discussed in Section 19 from the standpoint of the audio engineer.

Magnetic and optical film recording techniques are covered extensively in Sections 17 and 18 and will be of particular interest to those employed in both the motion picture and television industries.

Section 23 discusses the techniques of audio frequency measurements, and Section 24, tbe techniques of installation. These two sections should appeal to the installation and maintenance engineer because of their unusual type of information.

Finally, Section 25, "General Information, Charts, and Tables," contains much useful data for ready reference, including many charts for simplifying mathematical design procedures. Also, references will be found at the end of each section.

The data within this book is the result of over 30 years of experience in audio engineering, particularly in the motion picture industry. Tbe author sincerely hopes the material presented in this work will appeal to both the technician and the engineer.

I wish to extend my appreciation to Dr. George K. Teffeau for his valued suggestions and assistance in editing this book, to Messrs. Jack Laing and Glenn Osborn, as well as to Miss Janice Snyder, for their assistance in making tests for the sections on "Magnetic Recording," and "Optical Film Recording." I also wish to thank the following individuals, manufacturers, and publishers for making much of the data and photographs available for publication.

Lt. Col. James L. Gaylord, former Commander of the USAF Lookout Mt. Laboratory; Lt. Col. James P. Warndorf, present Commander; Mr. Watson Jones, RCA Victor Division, Hollywood, Calif.; Mr. Ralph Wight, Westrex Corp., Hollywood, Calif.; Mr. J. N. A. Hawkins, Motion Picture Research Council; Dr. Oliver Read, publisher of *Radio and Television News;* Mr. C. G. McProud, Editor and Publisher of *Audio* magazine; Mr. Hugo Gernsback, Editor and Publisher of *Radio Electronics* magazine; *Journal of the Society of Motion Picture and Television Engineers; Journal of the Institute of Radio Engineers; Journal of the Audio Engineering Society;* and *Journal of the Acoustical Society of America.* My appreciation also to the many other contributors of data and information.

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Contents

Note

In an effort to reach worldwide understanding in published scientific and technical work, standardization of terms is essential. The Institute of Electrical and Electronic Engineers (IEEE) standards committee has adopted a number of new standards of electrical nnits and symbols. These were established in close cooperation with many international organizations.

Where standards are quoted, the information has been extrapolated for the essential details given. The term recommended is used to signify accepted industry practice and is not necessarily a standard.

Section 1

Basic Principles of Sound

To the audio engineer, an understanding of the basic principles and behavior of sound generation as well as the response of the human ear, frequency ranges and the levels that are encountered in daily life, especially when recording, are equally as important as the understanding of electronic circuitry. New aspects of sound may be observed with the application of white- and black-sound techniques. Shown are the variations of the Fletcher-Munson curves developed by Churcher and King and Robinson and Dadson of England, and by Pollock of the United States. General phenomena associated with the generation and propagation of sound waves such as the Doppler effect, units of measurement, terminology, equations, ranges of musical instruments, and peak powers are included in this section.

1.1 What is the definition of sound? —It is a wave motion propagated in an elastic medium, traveling in both transverse and longitudinal directions, producing an auditory sensation in the ear by the change of pressure at the ear.

1.2 How are sound waves produced? —By a vibrating body in contact with the air.

1.3 What is the peak-to-peak amplitude of a waveform?—It is the value measured from the positive peak to the negative peak. (See Question 25.149.)

1.4 What is a compressional wave? —A wave in an elastic medium such as air, which causes an element of the medium to change its volume without undergoing rotation.

1.5 What is a longitudinal wave?—A wave in which the direction of displacement at each point of the medium is normal to the wave front.

1.6 What is a shear wave?—A wave in an elastic medium which causes an element of the medium to change its shape without changing its volume.

1.7 What is a transverse wave?—A wave in which the direction of displacement at each point of the medium is parallel to the wave front.

1.8 What is a plane wave?—A wave in which the wave fronts are parallel planes normal to the direction of propagation.

1.9 What is a spherical wave?—A wave in which the wave fronts are concentric spheres.

1.10 What is a cylindrical wave?—A wave in which the wave fronts are coaxial cylinders.

1.11 What is a beat?—A periodic variation resulting from the superpositioning of waves having different frequencies.

1.12 What is a simple sound source? —It is a source which radiates sound uniformly in all directions under free-field conditions.

1.13 What is direct-sound radiation?—Sound emitted directly from the sound source without reflections or echoes.

1.14 What is the effective acoustic center of an acoustic generator?—It is the point at which the spherically divergent sound waves, observed at a remote point, appear to diverge.

1.15 What is considered to be the audible frequency range? — 15 Hz to 20,000 Hz.

1.16 What is noise? — A random sound composed of many different frequencies not harmonically related. If noise is of too great an intensity, it will impair the intelligibility of speech and music, reducing the listening pleasure. The average dwelling has a noise level of about 40 dB above the threshold of hearing. (See Question 1.117.) The noise level of a business office will rise to around 55 dB, while an average factory will indicate about 80 dB. Noise meters (see Question 22.94) do not measure the true noise present, because of the complexity of the waveforms. As a rule, noise meters are equipped with filters and weighting networks, to simulate the human ear hearing characteristics at different levels above the threshold of hearing. Weighting networks are discussed in Question 2.93.

1.17 What are infrasonic frequencies?—Frequencies below 15 Hz.

1.18 What are ultrasonic frequencies?—Frequencies above the audible range of 20,000 Hz.

1.19 What are supersonics?—A term used in aerodynamics to denote a velocity greater than the velocity of sound.

The term supersonics was formerly used in acoustical engineering to designate frequencies above audibility. The term ultrasonics has now replaced the former term supersonics.

The installations are generally of high power and operate above 20,000 Hz, although at times they may be operated at a lower frequency.

1.20 What are macrosonics?—Macrosonics are the utilization of high-amplitude sound waves for cleaning small metal parts, drilling, emulsification, soldering, plating, and the aging of alcoholic beverages.

1.21 What is the energy distribution of human speech?—The greatest energy lies between 200 and 4000 Hz, falling off quite rapidly beyond 6000 Hz. This characteristic is shown in Fig. 1-21.

1.22 What does the term timbre mean?—It is the characteristic quality of a musical instrument which permits it to be distinguished from another. Timbre depends on the harmonic or overtone structure of the instrument. If the harmonics are removed by the use of filters, all instruments will sound similar, except for the pitch.

1.23 What is a transducer?—A device actuated by power from one system and supplying power to a second system. The actuating power may be electrical, mechanical, or acoustical. A loudspeaker is a typical example of an electroacoustic transducer.

1.24 What is a passive transducer?—A transducer whose output waves are independent of any source of power and are controlled by the actuating waves.

1.25 What is an active transducer?—A transducer whose output waves are dependent upon sources of power apart from that supplied by any of the actuating waves. The output is controlled by the actuating waves.

1.26 Define the term octave.—An octave is eight notes of a musical scale, and is the interval between two sounds having a basic frequency ratio of two. It is also the pitch interval between two tones such that one tone may be re-

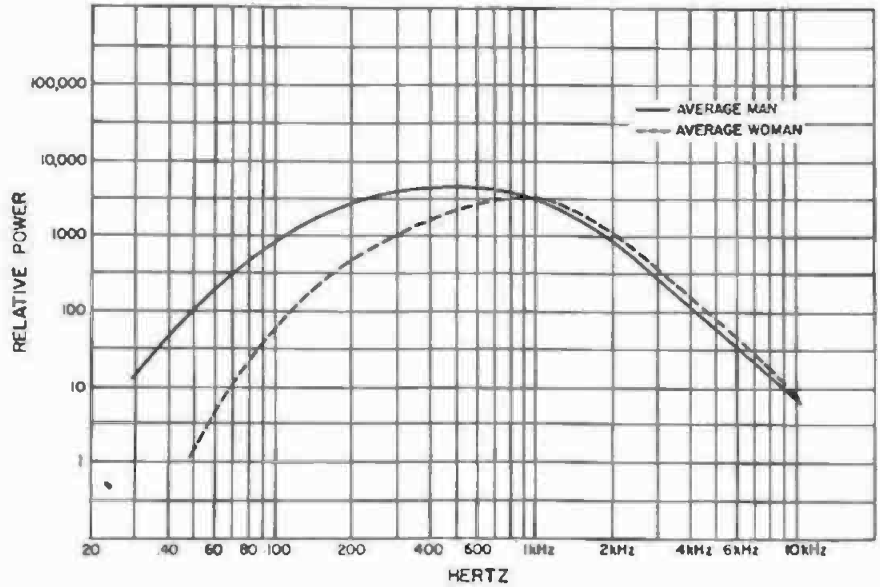


Fig. 1-21. Energy distribution of the human voice.

garded as duplicating the basic musical import of the other tone at the nearest possible higher pitch. An interval in octaves, between any two frequencies, is the logarithm to the base two (or 3.322 times the logarithm to the base 10) of the frequency ratio. The frequency ratio corresponding to an octave pitch interval is approximately, but not always exactly, 2:1.

1.27 What is a semitone, or half-step?—The same as a half-tone or any interval between two tones equal to the 12th root of 2.

1.28 What is a tone?—A sound wave capable of exciting an auditory sensation and having pitch. A sound sensation having pitch.

1.29 What is a chromatic scale?—A musical scale in which the intervals are all half-tones or semitones.

1.30 What is a scale of equal temperament?—A musical scale divided into twelve intervals. It is obtained by alternating the tones from the exact frequency of just intonation as a result of reducing the number of tones per octave.

These frequency ratios are 1, F, F_1, F_2, F_3 . . . to F_{12}, where F is the 12th root of 2 and F_{12} equals 2. The scale consists of 12 equal intervals, including half-tones. In Fig. 1-30 is shown the keyboard of a conventional 88 note piano with the tones indicated in hertz. A (440 Hz) is designated A_4 and the octaves above A_4 are designated A_5, A_6, and A_7. The octaves below A_4 are designated A_3, A_2, A_1, and A_0. The lowest frequency is A_0 or 27.5 Hz. The highest frequency is C_8, or 4186 Hz. The interval between the black and white keys is 100-cents, or an equally tempered semitone. (See Question 1.31.)

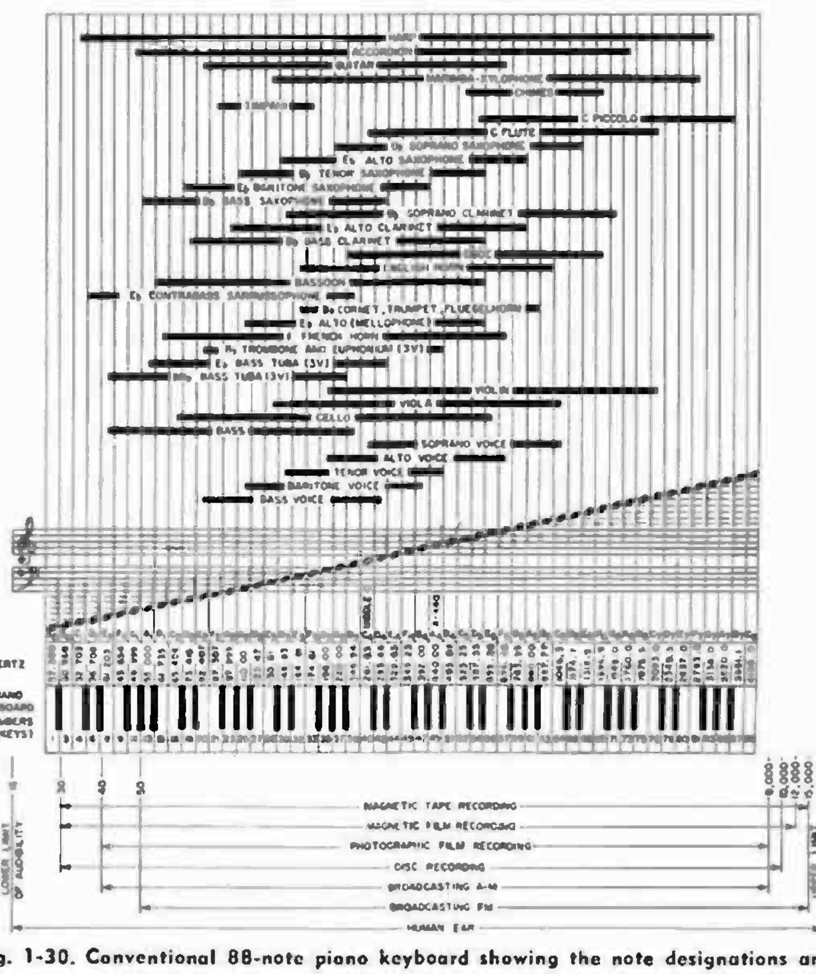


Fig. 1-30. Conventional 88-note piano keyboard showing the note designations and frequency. The audio frequency range of recording and broadcasting stations has been entered for comparison.

The complete spectrum is designated by starting at C_0, 16.351 Hz (not shown), the lowest frequency heard by the average human ear, and counting the octaves above from this frequency up to C^{9}, 16,744 Hz (not shown).

1.31 What is a cent?—An interval between two frequencies whose frequency is the 1200th root of 2. 1200 cents equals 12 equally tempered semitones.

1.32 What is pitch?—The property of a musical tone determined by its frequency and intensity. The higher the frequency the higher the pitch. (See Question 1.77.)

1.33 What is the American standard of pitch?—By agreement, all musical instruments manfactured in the United States are standardized to a frequency of 440 Hz for the tone of A_4. This corresponds to the 49th key on a standard 88-note piano. (See Question 1.30.)

1.34 What is a mel?—A unit of pitch. A 1000-Hz tone at a level of 40 dB above the threshold of hearing is equal to 1000 mels.

The mel scale of pitch is logarithmic above 1000 Hz and approximately linear below 1000 Hz. Any sound heard by an auditor and judged to be "n" times that of 1 mel is "n" mels.

1.35 Define the term microbar.—The microbar is a unit of measurement of sound pressure. One microbar is equal to the sound pressure of 1 dyne per square centimeter, or to a sound level of 74 dB above the threshold of hearing (0.0002 microbar). It is also equal to approximately one-millionth of normal atmospheric pressure. Normal atmospheric pressure equals 1,013,250 microbars. One bar equals 10^6 dynes per centimeter. At one time, the term bar was used to mean 1 dyne per square centimeter. This is no longer used. (See Question 1.82.)

1.36 Define threshold of pain.—The threshold of pain is the minimum value of sound pressure of a given frequency that will cause discomfort or pain to a listener 50 percent of the time, as shown in Fig. 1-36. Discomfort starts around a sound level of 118 dB, if the frequency falls within the range of 200 to 10,000 Hz. Actual pain starts around 140 dB in the same frequency range. It is interesting to note a sound level of 1 pound per-square-foot is equal to plus 147.6 dB; 1 pound per-square-inch to 170.8 dB; and 1 atmosphere to 194.1 dB—all being referred to 0.0002 microbar.

1.37 What does the term damping mean?—It is the introduction of friction or some other type loss, to reduce the vibration of a moving body. Undamped and damped waveforms are shown in Fig. 1-37.

1.38 What is critical damping?—A value of damping which will result in the most rapid response possible without overshooting or no indication of os-

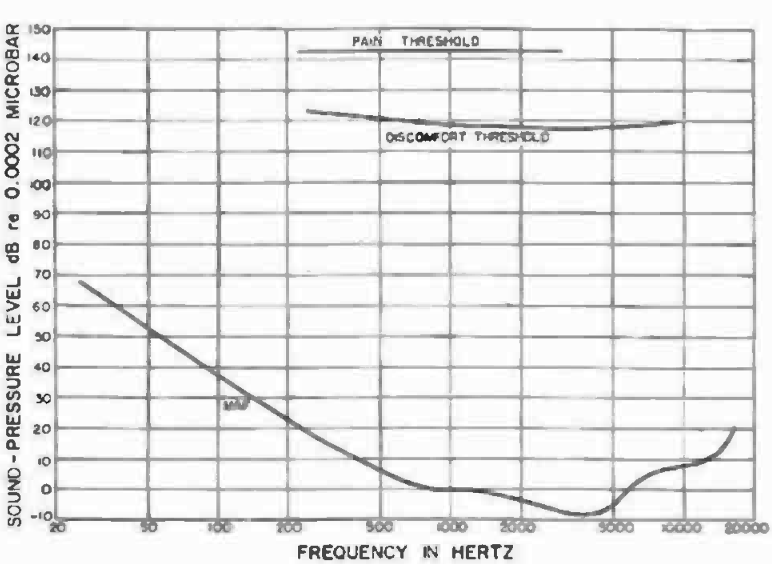
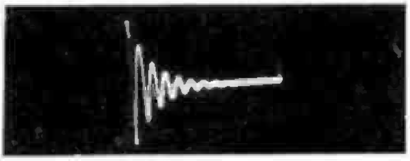
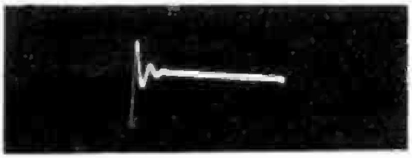


Fig. 1-36. Curves showing the threshold of hearing, discomfort, and pain. (Courtesy, General Radio Co.)



(a) *Undamped.*



(b) *Damped.*

Fig. 1-37. Undamped and damped waveforms.

cillation, is referred to as critical damping. It is the point of change between aperiodic and periodic damping.

1.39 What is periodic damping?—A system in which the damping is so great that, if the system is subjected to a single disturbance, it will come to rest.

1.40 What is aperiodic damping?—Aperiodic damping occurs when the amount of damping is so large that if the system is subjected to a single disturbance, either constant or instantaneous, the system comes to rest without passing through that position. While an aperiodically damped system is not strictly an oscillatory one, it has such properties that it could become an oscillating system if the damping was removed.

1.41 Define the term dyne per square centimeter—One dyne per square centimeter is the unit of force used in acoustical measurements and is the force that will give an acceleration of one centimeter per second, during each second it is operating, to a mass of one gram. One million dynes per square centimeter is equal to one bar. (See Question 1.125.)

1.42 What is a bar?—A unit of sound pressure equal to 10^6 microbars or 10^6 dynes per square centimeter.

1.43 What is superpositioning?—The combination of two or more frequencies, or the superimposing of one or more frequencies upon another frequency. (See Question 1.11.)

1.44 What is a sine wave?—A wave which rises from zero to maximum in one direction, returns to zero in a corresponding, gradual, nonuniform manner, reverses direction and falls below zero to an equal maximum, and returns to zero again during one complete cycle. A typical sine wave development is shown in Fig. 1-44.

1.45 What is a sinusoidal waveform?—An alternating current wave which varies in proportion to the sine of an angle. It is another name for the sine wave described in Question 1.44.

1.46 Define the terms cycles per second and hertz.—A complete turn of events. For a sine wave, it is the complete sequence of the rise from zero to maximum and return to zero, the rise to maximum in the opposite direction, and the return to zero again, as shown in Fig. 1-44.

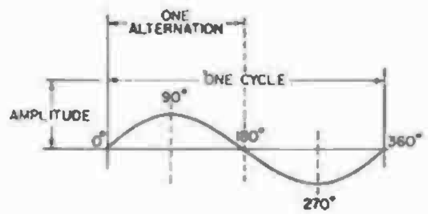


Fig. 1-44. Development of a sine wave.

The common term for measurements, when dealing with frequency, was in cycles per second, abbreviated cps. In 1965, the term cycles per second was changed to hertz in honor of Heinrich Rudolph Hertz, an early German physicist. One cycle per second is stated as 1 Hz; 1 kilocycle as 1 kHz; 1 megacycle as 1 MHz, etc. Either term is used, but *hertz* is preferred.

1.47 What is an alternation?—One-half of a cycle. As an example; a 60-Hz current has 120 alternations per second, similar to that shown in Fig. 1-44.

1.48 What does the term frequency mean?—The number of cycles or vibrations in a given unit of time.

1.49 What is a wavelength?—It is the distance from a given point on one wave to a corresponding point on the next wave, regardless of the frequency. Example: One wavelength in Fig. 1-44 is the distance from zero degrees to 360 degrees.

1.50 State the formula for calculating the wavelength of sound in air.—Experiments indicate the velocity of sound in air, at any frequency, is approximately 1127 feet per second, 343.4 meters per second, or 767.54 miles per hour. The velocity will vary slightly with changes in temperature, increas-

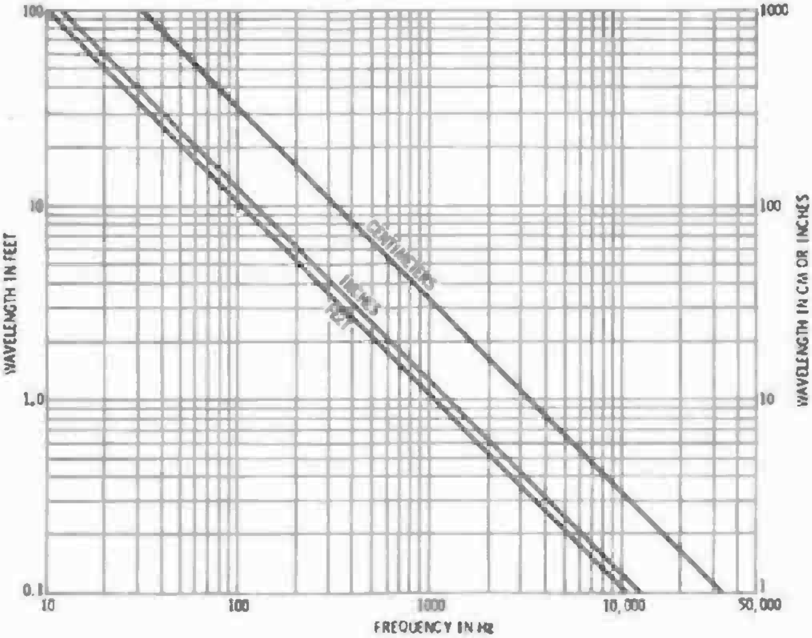


Fig. 1-50. Wavelength of sound in air as a function of frequency. Temperature 20° C @ 760mm pressure.

ing about 1.1 feet per second for 1 degree Fahrenheit, going from 1099 ft per second at 32° F to 1130 feet per second at 70° F. The wavelength of sound in air is equal to the velocity divided by the frequency in hertz. Example:

What is the wavelength of 40 hertz?

$$\lambda = \frac{V}{f} = \frac{343.4}{40} = 8.58 \text{ meters.}$$

where,

V is the velocity of sound in air,
f is the frequency in hertz, at 20° F.

The wavelength for various frequencies in air may be determined from the graph in Fig. 1-50. The graph is entered at the lower edge for a given frequency and followed upward to where it intersects the diagonal line. Wavelength is then read in feet from the left margin, or in centimeters or inches from the right margin.

1.51 Define an eigentone.—It is a resonance in an enclosure, caused by the presence of parallel walls, which causes the generation of standing waveforms. A room of cubical dimensions (dimensions equal in all directions) would have the same resonant frequencies in all directions. This is why good acoustical design does not employ parallel walls, and uses given ratios for length, width, and height.

1.52 What is a fundamental frequehcy?—The principal component of a complex waveform or the component having the lowest frequency. Using the physical definition, the fundamental frequency is called the first harmonic, F_1. In musical terms, the first multiple above the fundamental is called the first harmonic, F^1. (See Question 1.57.)

1.53 What is an overtone?—A harmonic of the fundamental frequency of a complex waveform.

1.54 What is an odd harmonic?—Any frequency that is an odd multiple of the fundamental frequency, such as 1, 3, or 5 times the fundamental.

1.55 What is an even harmonic?—Any harmonic that is an even multiple of the fundamental frequency, such as 2, 4, or 6 times the fundamental.

1.56 What is a subharmonic?—Subharmonics are obtained mathematically by dividing the fundamental frequency by the desired number harmonic. The second subharmonic of 500 Hz is 250 Hz.

1.57 How are the harmonics of a given frequency determined?—Any specified harmonic may be determined by multiplying the fundamental by the number of the desired harmonic. Example: The third harmonic of 100 Hz is 300 Hz. This is in physical terminology. (See Question 1.52.)

1.58 What are the characteristics of the various octaves between 16 and 16,000 Hz?—The frequencies between 16 and 32 Hz are called the first octave and are the frequencies in which the lowest swell tones of an organ are heard, and where the threshold of feeling exists.

Frequencies between 32 and 512 Hz are referred to as the second, third, fourth, and fifth octaves. This is the rhythm section where the lower and upper bass frequencies are found. If the response in the second and third octaves is accentuated, the reproduction will be reverberant and become objectionable to the true music lover. With the proper frequency response, the low frequencies of the drums and piano will be reproduced in their proper perspective.

From 512 to 2048 Hz are the sixth and seventh octaves. If speech is limited to this frequency range, it will have a telephonelike quality, because the human ear starts to approach its maximum sensitivity around 2000 Hz and continues up to 4000 Hz. If the response of the sixth octave is accentuated with respect to other parts of the frequency response, the quality of reproduction will have a hornlike quality. Also, if the response is increased between 1000 and 2000 Hz, the reproduction becomes tinny. Over accentuation of both the sixth and seventh octaves is one of the causes of listener fatigue.

Accentuating the eighth and ninth octaves, 2048 to 8192 Hz, adds presence to the program material and creates the illusion that a person speaking is present in the room. Within these frequencies lie the labial and fricative sounds. Labial sounds are those made with the lips and fricative those cause by the rustling of the breath as sound is emitted by the mouth. As a rule, recording systems are equipped with equalizers for controlling the amount of presence. (See Questions 6.106 and 6.107.)

The tenth octave occupies the region between 8192 and 16,000 Hz. These frequencies add brillance to the reproduction, reproducing the tinkling of bells and the higher frequencies of the triangle, cymbal, and other instruments. It has been proved by exhaustive listening tests that for high quality reproduction of music, the correct balance between the lowest and highest frequencies is obtained when the highest frequency multiplied by the lowest frequency equals 600,000.

Thus, a reproducing system with a low frequency response of 40 Hz and an upper frequency of 15,000 Hz will meet these requirements.

1.59 What is vibrato?—Frequency modulation of a tone or the musical embellishment of a tone. It is employed by vocalists and musical artists to enhance the presentation. The average rate of vibrato is 5 to 7 Hz.

1.60 What is tremolo?—Amplitude modulation of a tone. In an organ, the tone is modulated by a mechanical device.

1.61 What do the terms consonance and dissonance mean?—If two or more tones are sounded simultaneously and they are pleasing to the ear, they are consonant; if displeasing, they are said to be dissonant.

At times, the term mistuned consonance may be used in contrast to beats of nearly identical frequencies, which are termed imperfect unisons. The term mistuned consonants refers to beats between a given frequency and another frequency which is not quite in harmonic relationship with the former, such as 500 Hz and 1002 Hz.

1.62 What is a chord?—A combination of three to seven tones harmonically related and sounded simultaneously.

1.63 Define loudness.—It is the intensity of the sound stimulus and, chiefly, a function of sound pressure. However, it is also dependent on the frequency and the complexity of the waveform.

At times it may be advantageous to express the change in loudness in percentage. To do so, the level of the original sound must be known. The graph in Fig. 1-63A gives an approximate answer for different values of the original sound level. The tabulations in Fig. 1-63B were taken from Fig. 1-63A, for reductions of 6 and 10 dB.

1.64 How are the waveforms of program material classified?—They are of a complex nature and seldom of sine-wave characteristic. Typical complex waveforms are illustrated in Figs. 1-64A and B.

1.65 Describe the construction of a complex waveform.—A waveform con-

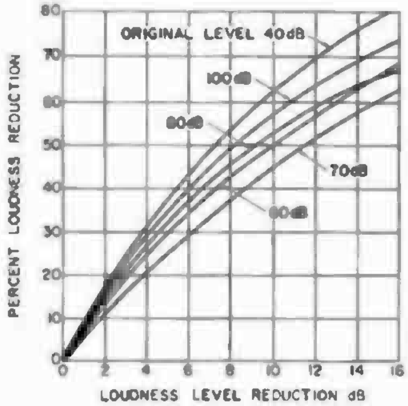


Fig. 1-63A. Percent loudness reduction versus loudness level in decibels.

ORIGINAL LOUDNESS LEVEL (dB)	REDUCTION IN DECIBELS	PERCENT LOUDNESS REDUCTION
40	10 6	62 43
60	10 6	50 33
70	10 6	44 29
80	10 6	53 38
100	10 6	56 40

Fig. 1-63B. Original loudness level versus reduction in decibels/percentage.

sisting of a fundamental frequency and its harmonics, or other frequencies, superimposed. When a complex waveform is viewed on an oscilloscope, it appears as shown in Figs. 1-64A and B. The waveform in Fig. 1-64A is typical of a symphony orchestra, while that in Fig. 1-64B is typical of speech waveforms. It will be noted that neither of the waveforms is of sine-wave character. The speech waveform indicates considerably more amplitude above the line than below; therefore, when transmitting or recording such waveforms, the system must be capable of transmitting or recording the maximum amplitudes without overloading.

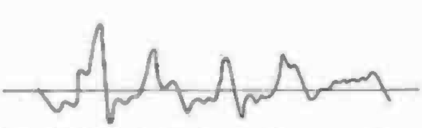


Fig. 1-64A. Waveform of a symphony orchestra as viewed on an oscilloscope.

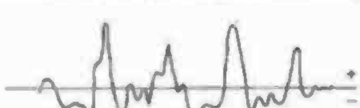


Fig. 1-64B. A typical speech oscillogram showing the energy or amplitude of the positive and negative peaks which are not symmetrical.

Because the maximum peaks are 8 to 14 dB above the average level indicated by a VU meter, it is necessary that this be taken into consideration when setting up a recording channel. If this is not done, serious overload will occur. This latter subject is discussed in Questions 10.27 and 10.28.

1.66 What is a partial?—A physical component of a complex waveform. It may be a frequency higher or lower than the fundamental frequency, and may or may not be harmonically related to the fundamental.

1.67 Define volume range.—It is the difference in amplitude between the softest and loudest passages in a recording or a live pickup. In a recording system, the signal-to-noise ratio of the system determines the maximum volume or dynamic range of that system. For a live pickup, the volume range is limited by the ambient noise level of the recording studio and the system noise combined (Fig. 1-67). Signal-to-noise measurements are discussed in Section 23.

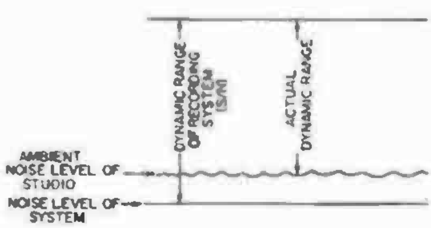


Fig. 1-67. The dynamic, or volume, range of a recording system is limited by its own signal-to-noise ratio, and the ambient noise level of the studio.

1.68 What is the range of the human voice in octaves?—When singing, about two octaves; however, female voices have been known to cover a four-octave range.

1.69 What is the effect of removing the lower frequencies from the human voice?—Because the greater portion of human speech intelligibility lies above

800 Hz, the lower frequencies may be removed without affecting intelligibility. Removing the higher frequencies reduces the intelligibility to the point where speech becomes unintelligible. Articulation tests indicate that, when freqnencies above 1550 Hz are removed, the intelligibility falls off 65 percent.

The reproduction of speech may be limited to a bandwidth of 250 to 3500 Hz with a high degree of intelligibility. For motion-picture dialogue recording, frequencies below 800 Hz are slowly rolled off to where 100 Hz is down 8 to 12 dB, with reference to 1000 Hz. The midrange high frequencies are accentuated 4 to 6 dB to add presence to the speaking voice. This characteristic is shown graphically in Fig. 18-81.

1.70 Does the human ear hear all frequencies with the same intensity?—No, the human ear is less sensitive at both the lower and upper ends of the frequency spectrum, and this characteristic varies with both age and sex as shown in Fig. 1-99A and Fig. 1-99B. The human ear shows its greatest sensitivity between 500 and 6000 Hz. As an example, at the lower intensities, the judged loudness of a 1000-Hz tone varies approximately in proportion to the square of the sound pressure. At the higher intensities, the loudness varies approximately as two-thirds power of the sound pressure.

This relationship should set to rest the contention that the decibel is used in acoustic measurements because the human ear responds to the intensity of sound logarithmically. At no level is a logarithmic characteristic indicated; actually, the loudness is approximately proportional to the sound pressure raised to the 0.6 power.

1.71 What is the frequency range of a telephone system for the transmission of speech?—For quality reproduction of the human voice, a telephone receiver must exhibit good frequency response across a bandwidth of about 3000 Hz, or from 200 to 3400 Hz. The intensity of normal conversation, heard at a distance of 3 feet, is on the order of 60 to 70 dB above the threshold of hearing. The frequency response for a modern telephone receiver and for one manufactured in 1905 is shown in Fig. 1-71.

1.72 What is meant by an inharmonic frequency?—A frequency which is not a rational multiple of another frequency.

1.73 What is instantaneous speech power?—The rate at which sound energy is being radiated by a speech source at any given instant.

1.74 What is peak speech power?—The maximum value of the instantaneous speech power within a given time interval.

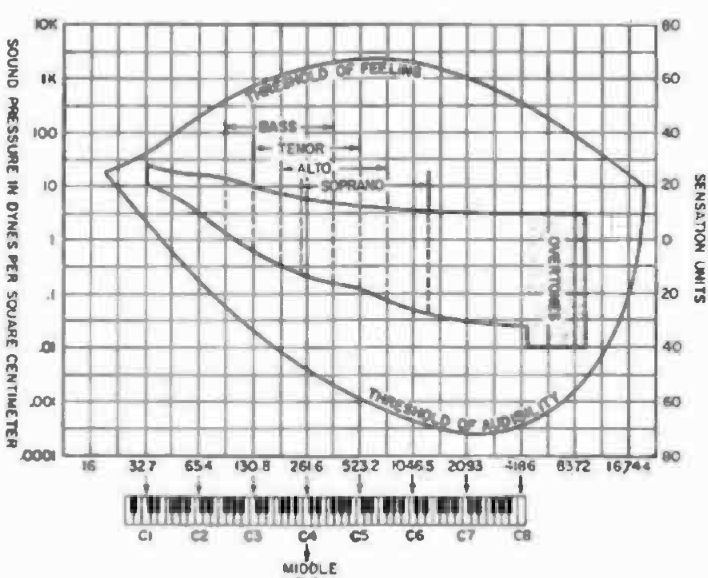


Fig. 1-70. Human ear frequency characteristics compared to the frequency range of a piano.

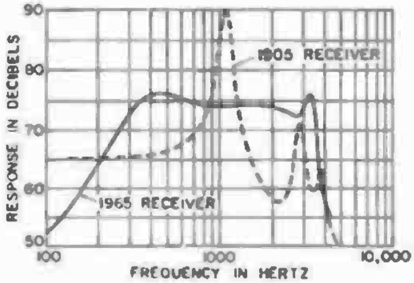


Fig. 1-71. Frequency response of a modern telephone receiver compared to one manufactured in 1905. (Courtesy, Bell Telephone Record)

1.75 ***What is average speech power?*** —The average value of the instantaneous speech power for a given interval.

1.76 ***What are equal loudness contours?***—A group of sensitivity curves plotted for the human ear showing its characteristic for different intensity levels between the threshold of hearing and the threshold of feeling. These data are often referred to as equal loudness contours. The reference frequency is 1000 Hz.

The curves of Fig. 1-76A made by Fletcher and Munson of the Bell Telephone Laboratories, are generally accepted throughout the sound industry as a basis for the design of devices concerned with human hearing. The curves given in Fig. 1-76B were made by Churcher and King of England; those in Fig. 1-76C hy Robinson and Dadson also of England; and those of Fig. 1-76D by Pollack of the United States. Although these curves differ somewhat, they all give the same general information, and are in fairly close agreement with the Fletcher-Munson curves, except for the differences at the higher and lower frequencies. The dotted line in Fig. 1-76C is the minimum audio frequency (MAF) that can be heard by the human ear. The curves of Fig. 1-76D by Pollack were made in 1952 using relatively narrow bands of random noise.

Referring to Fig. 1-76A, it will be observed that the contours are not equally spaced, but generally converge at the lower frequencies. This characteristic causes a change in the quality of reproduced sound when the volume level is changed. As an example: Consider a system carrying two pure tones of equal loudness, 1000 Hz and 50 Hz. Let the 1000-Hz tone have an intensity level of plus 80 dB (re 0.0002 microbar). Following this curve out to 50 Hz it indicates that an intensity level of plus 85 dB is required to make the 50-Hz tone sound as loud as the 1000-Hz tone, or a difference of 5 dB. Assume now that the gain of the system is lowered to where the 1000-Hz tone has an intensity level of plus 40 dB. By following this curve out to 50 Hz, it can be seen that the intensity must be increased to 72 dB to equal the loudness of the 1000-Hz tone, or an increase of 32 dB. It will be noted that in lowering the level, the balance between the low- and middle-range frequencies has been destroyed, and to the ear it appears that the low frequency response is lacking. This is the principal reason for including a loudness control in home reproducing equipment. It also emphasizes the need for establishing a monitor level in a recording studio and maintaining it from day to day to assure uniformity of product. (See Question 5.65.)

Weighting networks used in sound-level meters (see Question 2.93) are based on the Fletcher-Munson measurements, but modified to take into account the random nature of a sound field in an enclosure.

1.77 ***Why do sounds heard by the human ear increase and decrease in pitch as the intensity of the sound is varied?***—It is caused by the nonlinearity of the human ear. It has been determined experimentally that the pitch is related to the basilar membrane, a delicate part of the cochlea, a spiral, cylindrical tube forming the innermost of the three portions of the labyrinth of the human ear. Because of the structure of this membrane, the relationship between pitch and frequency is not exactly linear.

If the intensity of 100 Hz is increased from a loudness of 40 to 100, the pitch will be decreased about 10 percent. At a frequency of 500 Hz, the pitch will be changed about 2 percent for the same increase of intensity.

1.78 ***What is a consonant?***—A speech sound characterized in enunciation by constriction of the breath channel. Letters such as P, G, N, L, R, and W are called consonants.

1.79 ***What are sibilants?***—High-frequency sounds uttered with a hissing effect. The letters S and Z and the combinations SH, ZH, and CH are typical examples of sibilant sounds. These sounds are a constant source of annoyance to the recording engineer.

For motion-picture recording, sibilant sounds are attenuated by the use of a "de-esser" installed in a compressor amplifier, as described in Questions 18.89 and 18.91.

1.80 What is the sound pressure one foot from the mouth of a person speaking in a normal tone of voice?—Approximately one dyne per square centimeter. For a shout, the pressure may rise to 10 dynes per square centimeter, while a whisper may fall to 0.10 dyne per square centimeter.

1.81 How does a typical speech waveform appear on an oscilloscope?—A typical waveform was shown in Fig. 1-64B. It will be noted that the amplitudes of the positive peaks are greater than for the negative peaks, because the human voice is not of a true sine-wave character.

1.82 Define the threshold of hearing.—It is the sound level for a given frequency that the average human ear can just hear 50 percent of the time. The threshold of hearing is relative to a sound pressure of 0.000204 dyue per square centimeter, which equals 0.0002 microbar or 10^{-16} watt per square centimeter. (See Questions 1.35 and 1.41. The point in space must be specified.)

1.83 What is the dynamic range of the human ear?—120 dB, or a ratio of one trillion-to-one. A high quality magnetic-tape recording system has only a dynamic range of 60 dB, or one million-to-one.

1.84 What is an articulation test?—A quantitative measurement of the intelligibility of human speech. The test is conducted by reading selected sentences to a group of listeners. A score is kept by each listener as to the intelligibility of each sentence, word, and syllable. Such tests are used in the testing of telephone and other communication equipment.

1.85 What is the average deflection of the human ear drum for sounds of different intensities?—The variation in deflection is from 10 to 40 millionths of an inch.

1.86 How are acoustic powers added?—If the level is given in decibels, it is converted to power, added or subtracted as the case may be, and then converted back to decibels.

1.87 How much must the loudness of a sound be increased to make it sound twice as loud to the human ear?—Physiologists have devised a scale of loudness, which rank-orders sound from soft to loud in units of sones. The physical quantity measured is the sound pressure level. One thousand Hz is used as a reference frequency, at a sound pressure of 40 dB above the threshold of hearing (0.0002 microbar); this is taken to be 1 sone. Doubling the loudness is equal to 2 sones. The relationship of sones versus sound pressure is shown in the graph of Fig. 1-128. At the low intensities, the loudness of the 1000-Hz tone varies approximately in proportion to the square of the sound pressure, while at the higher intensities it varies approximately two-thirds power of the sound pressure. (See Question 1.70).

1.88 If two tones of the same frequency and intensity are sounded simultaneously, is the sound intensity twice as loud as for a single tone?—No. The overall level of intensity will only be iucreased 3 dB. Fig. 1-88 is a chart for adding acoustic powers. Assume two tones, one plus 30 dBm and one plus 22 dBm, are to be added. What is the acoustical power for the two, in decibels? Since 30 dBm minus 22 dBm is a difference of 8 dB, enter the chart at the bottom for 8 dB and follow the line upward to the intersection of the curve. The value read at the left of the chart, in this case 0.65 dB, is added to the higher power. Thus, the overall level for the two powers is 30.65 dBm.

This chart may be used for adding measured noise levels or for any problem of a similar nature.

1.89 What is a whole tone?—The interval between two sounds whose basic frequencies are a ratio approximately equal to the sixth root of two.

1.90 Define velocity.—It is the time rate of change of position. Unless angular velocity is stated, the term is understood to mean linear velocity.

1.91 What is the velocity of sound through solid materials?—

See Fig. 1-91.

1.92 What is the frequency range through water?—From 2 Hz to 50 megahertz (50,000,000 Hz).

1.93 Does the velocity of sound depend on frequency?—No. The velocity

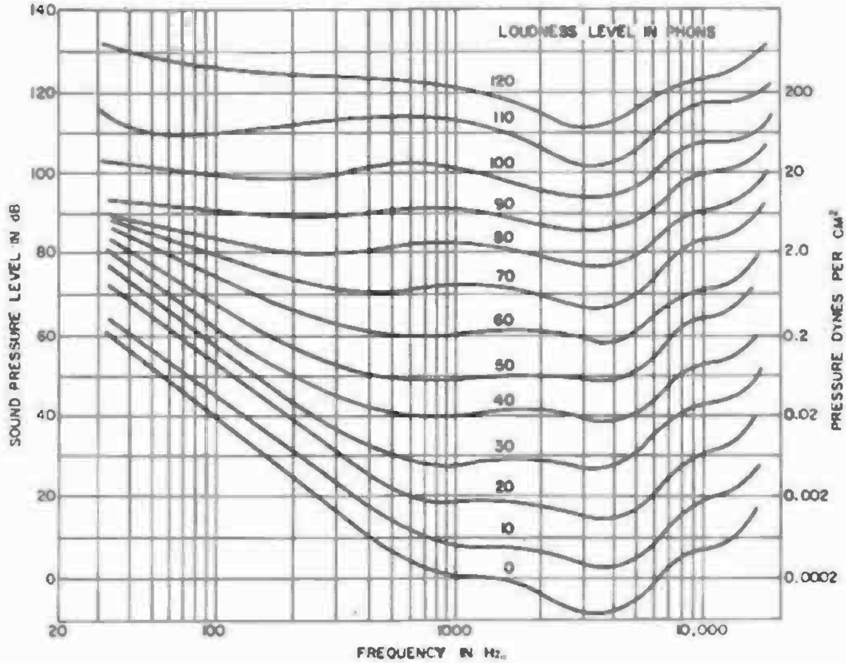


Fig. 1-76A. Fletcher-Munson curves (USA). (Courtesy, Acoustical Society of America)

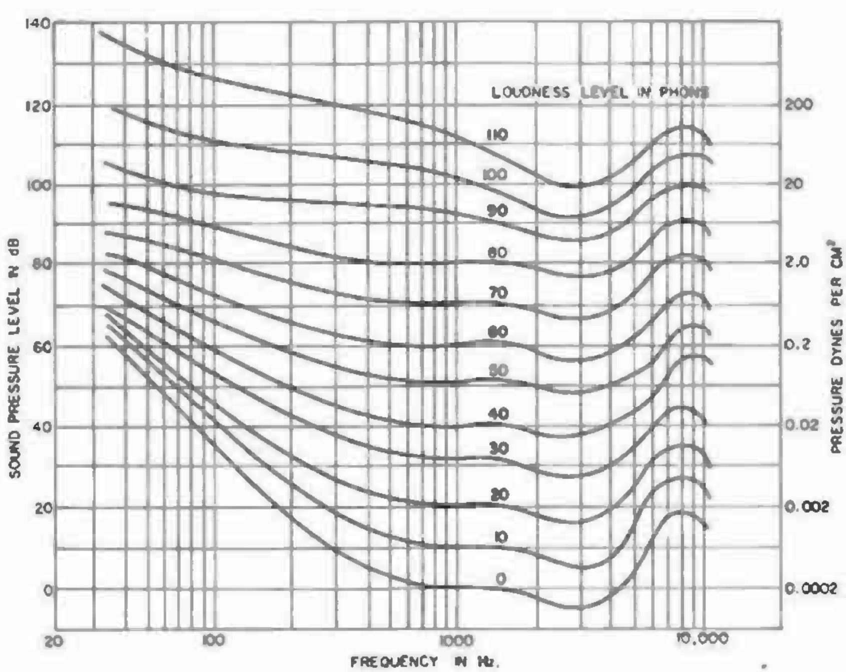


Fig. 1-76B. Churcher-King curves (England). (Courtesy, Acoustical Society of America)

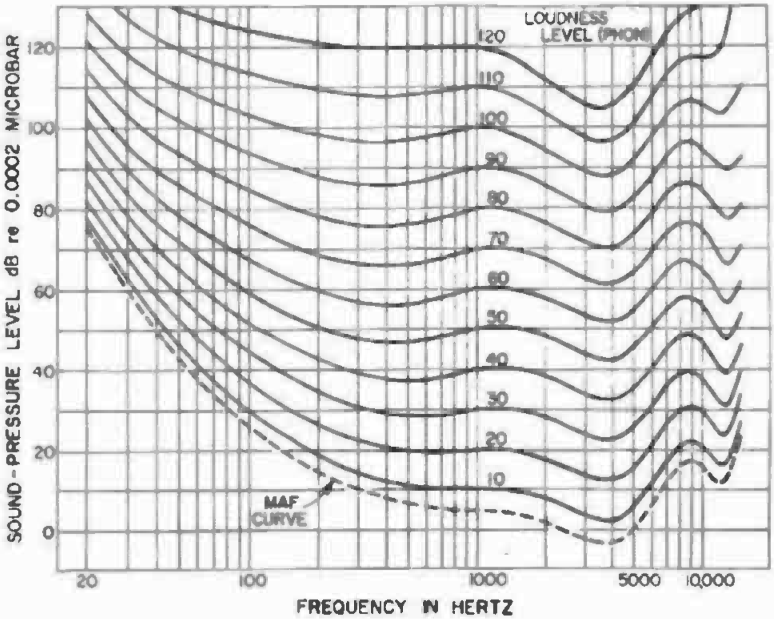


Fig. 1-76C. Robinson and Dadson free-field equal loudness contours. Observer facing the source of sound.

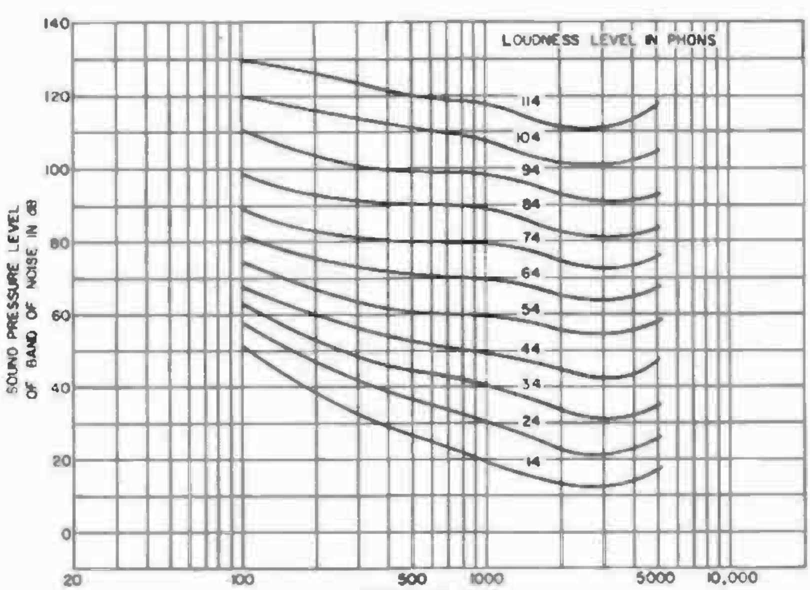


Fig 1-76D. Equal loudness contours for relatively narrow bands of random noise, 300 mels in width (after Pollack).

Material	*Feet per second*
Aluminum	16,740
Brick	12,500
Copper	11,670
Carbon dioxide	846
Glass	16,830
Iron	5130
Lead	4026
Pine (direction of grain)	10,900
Silver	8553
Water	4728
Steel	16,500

Fig. 1-91. Velocity of sound.

of sound is governed by the medium of transmission, the temperature, and the intensity of the sound.

1.94 Show the attenuation of high frequency sound waves in air.—Sound waves are affected by distance, humidity, and frequency. The attenuation in air for several frequencies, at different humidities, is shown in Fig. 1-94.

1.95 What is the decay rate of sound?—The time required by a sound to diminish from its original intensity to the threshold of hearing, or one-millionth of its original intensity.

1.96 How is decay rate measured?—Generally by an oscillator and an automatic graphic recorder. (See Question 22.112. The recorder measures the decay rate in milliseconds. The intensity level at the end of a given decay period, in decibels, is:

$$dB = 10 \, Log_{10} \frac{P_1}{P_2}$$

where,

P_1 is the original intensity in watts,
P_2 the diminished intensity in watts.

1.97 What is the acceleration of a sound wave?—The time-rate of change of the velocity of the point.

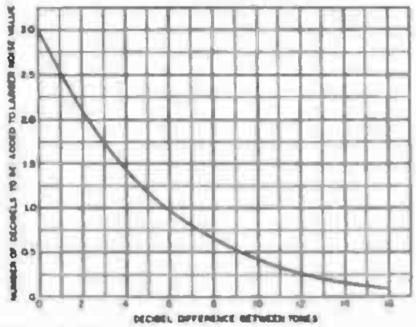


Fig. 1-88. Chart for adding acoustic powers or noise values.

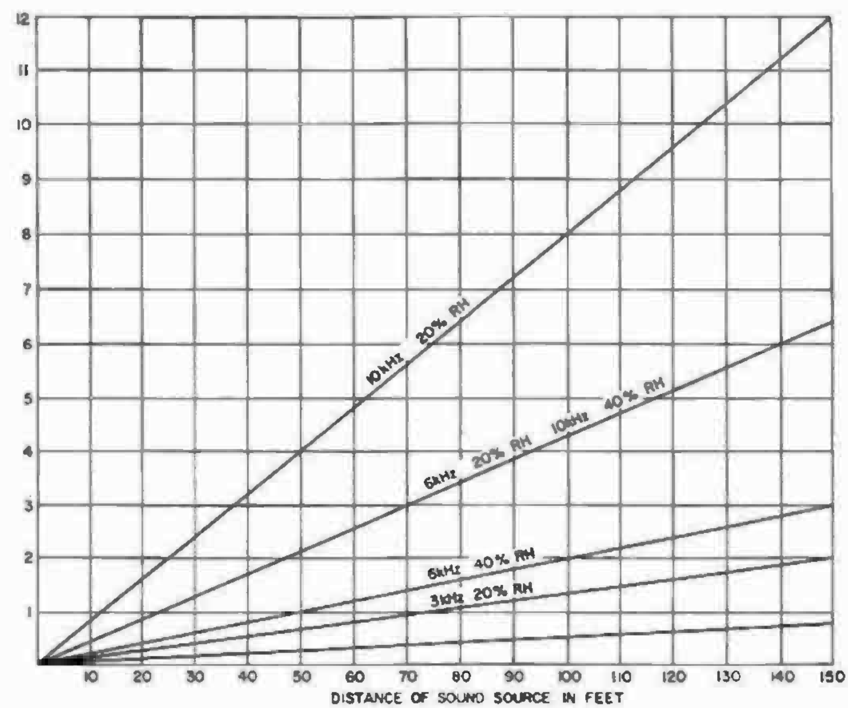


Fig. 1-94. Absorption of high-frequency sound waves in air for a given distance and relative humidity.

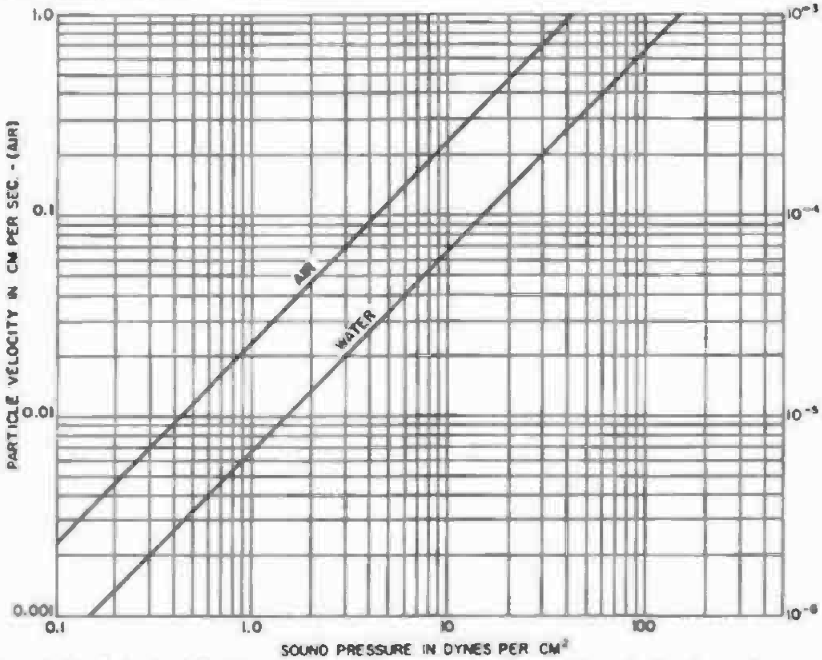


Fig. 1-98. Relationship between sound pressure and particle velocity in a plane wave in air and water at 20° Centigrade.

1.98 What is particle velocity?—It is the velocity of a given infinitesimal part of the medinm, with reference to the medium as a whole, due to the sound wave. The unit of measurement is the centimeter.

This relationship is shown graphically in Fig. 1-98. For example, to find the velocity for a sound pressure of 3 dynes/cm² in water, enter the chart at the lower left edge at 3 dynes, and follow the vertical line upward until it intersects the diagonal line for water; read the velocity at the right margin 0.02 cm/second.

1.99 Show the hearing losses with age for the human ear.—In 1939 The Bell Telephone Laboratories conducted tests at both the New York World's Fair and San Francisco Exposition to determiue the effect of age on the hearing of the human ear. Over one-half million records were made in five age groups, 10-19, 20-29, 30-39, 40-49, and 50-59. The results of these tests are shown in Fig. 1-99A and Fig. 1-99B. It will be noted that the high-frequency loss is less with age for women as compared to men; however, this is reversed at the lower frequencies.

1.100 What law does the growth and decay of a musical instrument follow?—Generally, the exponential law affecting the time required for a tone to reach its fullest intensity and then decrease to a given level of intensity.

1.101 What is meant by the term in phase?—The state or condition existing when two devices or sound waves are in perfect synchronization.

1.102 Explain the relationship of acoustic power to the intensity of sound.—The intensity of sound, independent of its frequency, is proportional to the average of the square of the pressure

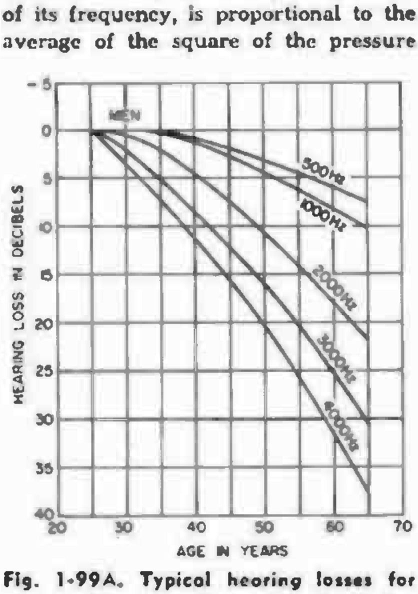


Fig. 1-99A. Typical hearing losses for men. (Courtesy, General Radio Co.)

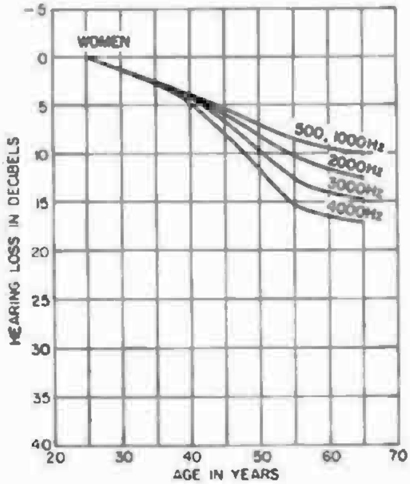


Fig. 1-99B. Typical hearing losses for women. (Courtesy, General Radio Co.)

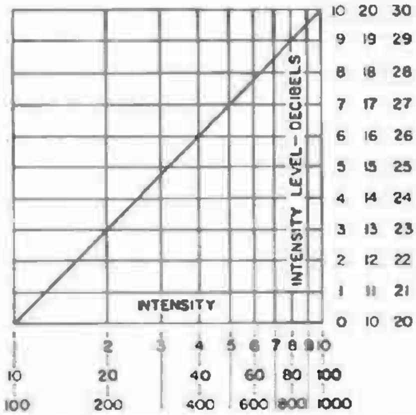


Fig. 1-103. Graphical plot showing the relationship of intensity to intensity level.

taken over a complete pressure cycle. Intensity is defined as the power in watts that is transmitted across one square centimeter of wavefront, perpendicular to the direction in which the sound is traveling. The power of even the most intense sound expressed in watts is extremely small. The unit of acoustical intensity is 10^{-16} watts per square centimeter. This intensity is slightly less than the least intensity of a 1000-Hz tone which is audible to the human ear. Even a painfully intense sound has the intensity of only 1/1000 watt per square centimeter. The square root of the average square (root-mean-square) of the sound pressure that corresponds to 10^{-16} watt per square centimeter, is 0.0002 dyne per square centimeter. A dyne is the force equal to 1/980 of the weight of a gram. Therefore, an rms sound pressure of 0.0002 dyne per square centimeter is equal to about two-millionths of a gram weight per square centimeter. The intensity of sound is generally expressed in 10^{-16} watt units, and sound pressure in 0.0002 dyne per square centimeter.

1.103 Define intensity and intensity level.—As an acoustical device, the human ear is unsurpassed in the range of intensities to which it will respond, without being damaged, as well as in its extreme sensitivity to faint sounds. However, the ear is relatively insensitive to changes in intensity. Roughly speaking, the intensity of a sound must be increased by approximately 26 percent, in order for the ear to register a change in the loudness sensation produced.

The graph in Fig. 1-103, illustrates the relationship between intensity and intensity level. The ordinates of the points on the straight line are ten times the logarithms of the numbers of the horizontal scale. Thus, intensity level is plotted versus intensity. The first vertical scale corresponds to levels of intensities from 1 to 10, the second from 10 to 100, the third from 100 to 1000, etc. The graph may be extended to give the intensity level of any intensity by multiplying the intensity by 10, which raises the intensity level by 10 dB. Thus, the intensity level for 2.5 is 4 decibels; for 25 it is 14 dB; for 250 it is 24 dB; for 2,500,000 it is 64 dB.

1.104 What is the relationship of distance to the intensity of a sound wave?—Sound in free space follows an inverse-square law; that is, the intensity of the sound varies as the square of the distance. If an observer moves to twice the distance from the sound source, the intensity will decrease to one-quarter the original intensity. If the distance is increased to ten times, the intensity will be reduced to one one-hundredth that of the original intensity.

This relationship exists because the area covered by the sound wave increases as the square of the distance. The sound energy is spread out over a greater area; therefore, the energy is decreased inversely by the same amount.

1.105 What is the relationship of a sound wave in air to a sine wave?—The energy of the sound wave in air is proportional to the rms value of the sine wave.

1.106 What is static pressure?—The pressure existing in a medium when no sound waves are present. The pressure is expressed in microbars.

1.107 What is instantaneous sound pressure?—It is the total instantaneous pressure at a point minus the static pressure at that point. Instantaneous sound pressure is stated in microbars.

1.108 What is the maximum sound pressure?—The maximum absolute value of the instantaneous sound pressure occurring during a given cycle, stated in microbars.

1.109 What is peak sound pressure?—The maximum absolute value of the instantaneous sound pressure in a given interval, stated in microbars.

1.110 What is effective sound pressure?—It is the rms value of the instantaneous sound pressures, over a given time interval, at a given point. In the case of periodic sound pressures, the interval must be an integral number of periods or an interval longer than one period.

1.111 How are sound pressure levels stated?—The level, in decibels, of a sound is 20 times the logarithm to the base 10 of the ratio of the pressure of the sound to the reference pressure. Reference pressures in common use are:

(a) 2×10^{-4} microbar.
(b) 1 microbar.

Reference pressure (a) is the one most commonly used for the calibration of microphones and sound level measuring instruments.

1.112 What is the sound power of a source?—The total sound energy radiated by the source per unit of time. The commonly used unit is the erg per second, but it may also be expressed in watts.

1.113 What is the dynamic range of a full symphony orchestra?—From 20 dB to 100 dB above the threshold of hearing. The lowest level will be affected by the ambient noise level of the auditorium.

1.114 How are sound pressures measured?—In microbars. One microbar equals a pressure of one dyne per square centimeter, or approximately one-millionth of the normal atmospheric pressure.

1.115 What is a sympathetic vibration?—An undamped body set in vibration on a certain frequency by airborne sound waves or building vibrations.

1.116 What are Fletcher-Munson contours?—A group of ear characteristics plotted with reference to frequency and pressure, starting at the threshold of hearing. These curves are illustrated in Fig. 1-76A.

1.117 Show typical sound levels encountered in daily life.—The sound levels encountered in daily life are many and vary over a large range of sound pressures. In Fig. 1-117A are shown typical overall sound levels referenced to the threshold of hearing. Typical acoustic power levels for various sources are shown in Fig. 1-117B. These latter levels bear no simple relation to those shown in Fig. 1-117A.

It is generally more convenient to express the ratio between two sound pressures in decibels. Since sound pressure is generally proportional to the square root of the sound power, the sound pressure ratio for a given number of decibels is the square root of the corresponding power ratio. For example, if one sound pressure is twice another, the number of decibels is 6; if it is 100 times as great, it is 40 dB. Sound pressure can also be expressed as sound pressure level with respect to a reference sound pressure. For airborne sound, this reference is generally 0.0002 microbar. The definition of sound pressure level (SPL) is:

$$\text{SPL} = 20\text{Log}_{10}\frac{P}{0.0002} = \text{dB}$$

where,

P is the root-mean-square sound pressure in microbars.

For example, if the sound pressure is 0.00025 microbar, the corresponding sound pressure level is:

$$\frac{1}{0.00025} = 4000$$

$$20\text{Log}_{10} 4000 = 20 \times 3.60 = 72.0 \text{ dB}$$

A chart for converting decibels to pounds per square inch rms pressure, or vice versa is given in Fig. 1-117C.

1.118 What effect does the nonlinearity of the human ear have on hearing?—If a pure sine wave is impressed on the human ear and the intensity is increased, harmonics of the fundamental frequency appear.

Fig. 1-117A. Typical overall sound levels encountered in daily life. The levels are in reference to 0.0002 microbar. (Courtesy, General Radio Co.)

Since the human ear is nonlinear, multiple tones cause the generation of sum and difference frequencies thus creating distortion (see Question 1.77). Using a harmonic wave analyzer, Newman, Stevens, and Davis were able to detect the presence of 66 different frequencies in the electrical response from the cochlea of a cat's ear, when stimulated with 700 and 1200 Hz at a SPL of 90 dB above the threshold of hearing.

This indicates the complexity of the ear and the great number of frequencies generated for only the application of two frequencies. As the number of tones are increased, so are the sum and difference frequencies. Thus, the number of frequencies generated by the ear is tremendeous when listening to a full orchestra. A table of frequencies generated by a cat's ear is given in Fig. 1-118. This phenomenon is termed *aural harmonics.*

1.119 What effect does nonlinear reproduction have on the human ear?— Nonlinear reproduction induces harmonics not present in the original program material. If the high frequency response is reduced, an increase of distortion can be tolerated.

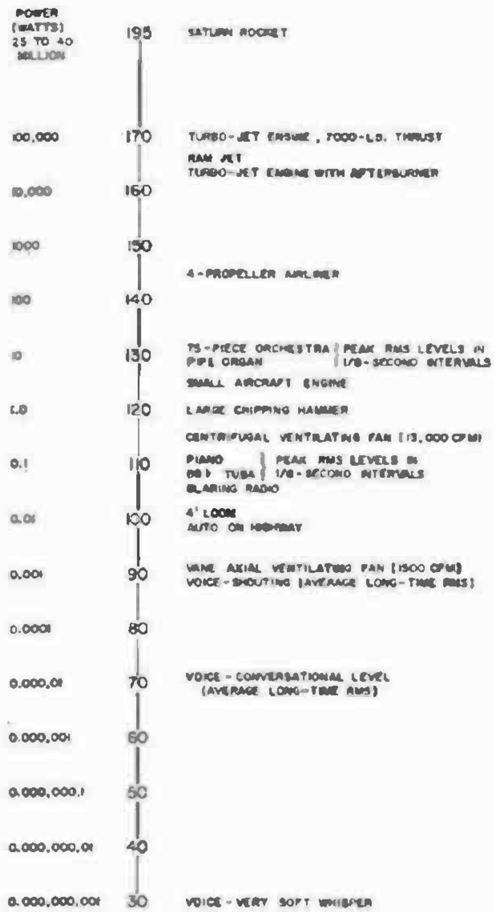


Fig. 1-117B. Typical power levels for various acoustic sources. These levels bear no simple relation to the sound levels of Fig. 1-117A. (Courtesy, General Radio Co.)

Starting from a very low value of distortion and slowly increasing the harmonic distortion will cause the third harmonic to become audible, at low levels, with a value of about 1.25 percent. The second harmonic may be increased to approximately 5 percent before it becomes annoying. Reproducing systems with a high-frequency cutoff of 6000 Hz may tolerate distortion up to 10 percent, while a system extending to 15,000 Hz can not tolerate over 2 percent total harmonic distortion. A good rule to follow is: Before increasing the

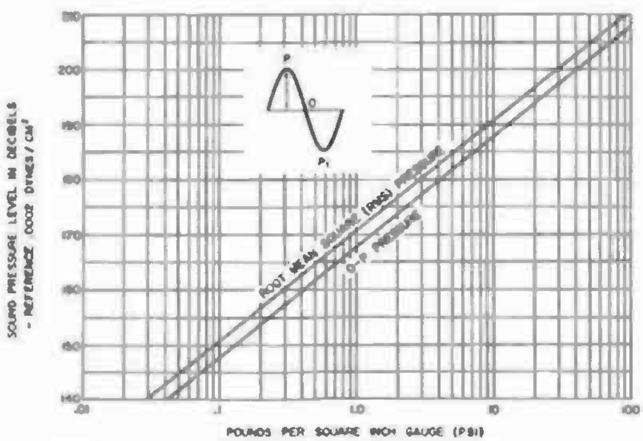


Fig. 1-117C. Conversion chart, decibels to pounds per square inch.

Frequency	Response in % of Fundamentals	Source	Frequency	Response in % of Fundamentals	Source
100	2.0	$3f_1 - 5f_2$	3700	1.1	$6f_1 - 5f_2$
200	7.9	$2f_2 - f_1$	3800	2.2	$2f_1 + 2f_2$
300	3.7	$2f_1 - 3f_2$	3900	0.84	$5f_1 - 3f_2$
400	0.58	$4f_2 - 2f_1$	4000	0.47	$4f_2 - f_1$
500	4.0	$f_1 - f_2$	4100	1.4	$4f_1 - f_2$
700	100.0	f_2	4200	0.6	$6f_2$
800	3.3	$3f_1 - 4f_2$	4300	1.0	$3f_1 + f_2$
900	1.0	$3f_2 - f_1$	4400	0.42	$6f_1 - 4f_2$
1000	1.3	$2f_1 - 2f_2$	4500	2.3	$3f_2 - 2f_2$
1200	100.0	f_1	4600	1.0	$5f_1 - 2f_2$
1300	1.8	$4f_1 - 5f_2$	4700	0.42	$5f_2 - f_1$
1400	3.1	$2f_2$	4800	0.32	$4f_1$
1500	1.3	$3f_1 - 3f_2$	4900	0.33	$7f_2$
1600	0.16	$4f_2 - f_1$	5000	2.7	$3f_1 - 2f_2$
1700	18.0	$2f_1 - f_2$	5100	1.4	$6f_1 - 3f_2$
1800	0.83	$5f_1 - 6f_2$	5200	0.45	$4f_2 - 2f_1$
1900	3.2	$f_1 - f_2$	5300	0.2	$5f_1 - f_2$
2000	0.58	$4f_1 - 4f_2$	5400	0.33	$6f_2 - f_1$
2100	1.1	$3f_2$	5500	0.38	$4f_1 - f_2$
2200	3.2	$3f_1 - 2f_2$	5600	0.67	$7f_1 - 4f_2$
2300	0.17	$5f_2 - f_1$	5700	0.97	$3f_1 + 3f_2$
2400	3.0	$2f_1$	5900	0.33	$5f_2 + 2f_1$
2600	8.0	$2f_2 + f_1$	6000	0.1	$5f_1$
2700	3.5	$4f_1 - 3f_2$	6100	0.28	$7f_2 + f_1$
2800	2.0	$4f_2$	6200	0.45	$4f_1 + 2f_2$
2900	2.5	$3f_1 - f_2$	6400	0.77	$4f_2 + 3f_1$
3000	0.2	$6f_1 - 6f_2$	6500	0.33	$6f_1 - f_2$
3100	10.0	$2f_1 + f_2$	6600	0.12	$6f_2 + 2f_1$
3200	1.3	$5f_1 - 4f_2$	6700	0.25	$5f_1 + f_2$
3300	1.2	$3f_2 + f_1$	6900	0.67	$4f_1 + 3f_2$
3400	1.3	$4f_1 - 2f_2$	7100	0.1	$5f_2 + 3f_1$
3500	0.2	$5f_2$	7200	0.15	$6f_1$
3600	1.8	$3f_1$	7600	0.3	$4f_2 + 4f_1$

Fig. 1-118. Sum and difference frequencies measured from the cochlea of a cats ear, when stimulated by frequencies of 700 Hz (f_2) and 1200 Hz (f_1) (after Newman, Stevens, and Davis).

frequency range of any reproducing system, reduce the harmonic distortion to a negligible amount.

1.120 What is the relationship of a fundamental frequency to its harmonics? —A fundamental tone with its harmonics up to an including the sixth harmonic, is shown in Fig. 1-120.

1.121 What is a free-field?—A sound field free from reflecting surfaces, or sound in free space.

1.122 What is a node or nodal point?—A point on a vibrating body that is free from vibration, or a point of zero potential in an electrical circuit, with respect to ground.

1.123 What is a nodal diagram?— A diagram of a stretched diaphragm showing the points of maximum and minimum vibration for a given set of conditions. Such patterns are made by vibrating a diaphragm having a light covering of sand on its surface. The sand will assume a pattern indicating the various points of vibration.

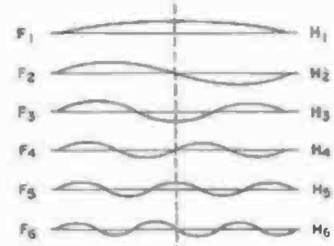


Fig. 1-120. A fundamental frequency and its harmonics.

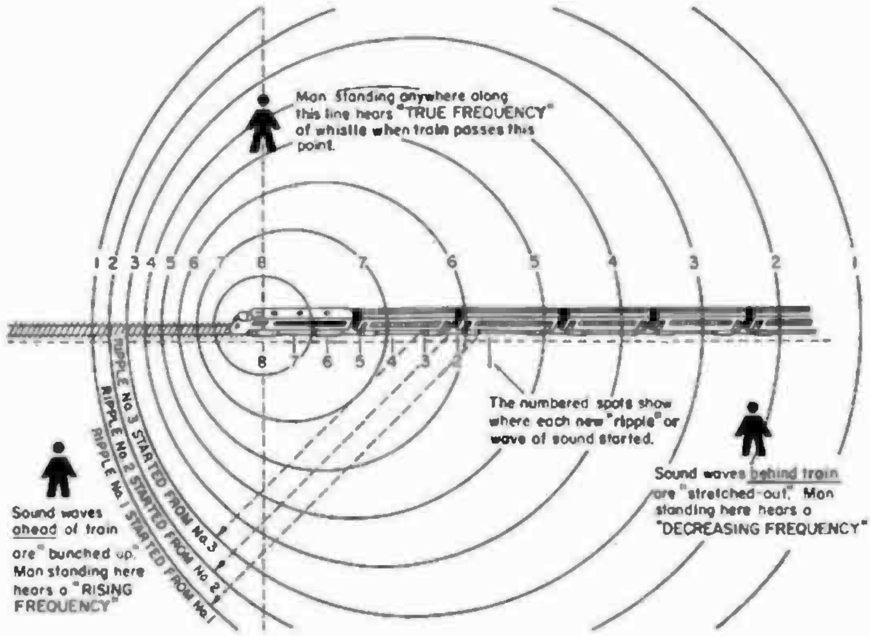


Fig. 1-124. The Doppler effect, showing how the frequency of a train whistle is affected as the train approaches and recedes from the observer.

1.124 What is the Doppler effect?—The change in pitch of a sound heard by an observer when the sound source is in motion. An example of the Doppler effect will be noted when a train blowing its whistle approaches an observer. The sound appears to increase in loudness and pitch. After passing the observer, the pitch and intensity drop quite rapidly until the sound fades out completely.

The increase in pitch is caused by compression of the sound wave as a result of the forward motion of the train being added to the velocity of the sound wave. Conversely, as the train moves away from the observer, the pitch decreases because the speed of recession is subtracted from the normal velocity of the sound wave, resulting in a lower pitch. This effect is illustrated in Fig. 1-124.

1.125 Define the term dyne.—The dyne is a unit of force used in acoustic measurements. Approximately 450,000 dynes equal one pound of force, and 68,944 dynes per square centimeter equal one pound of force per square

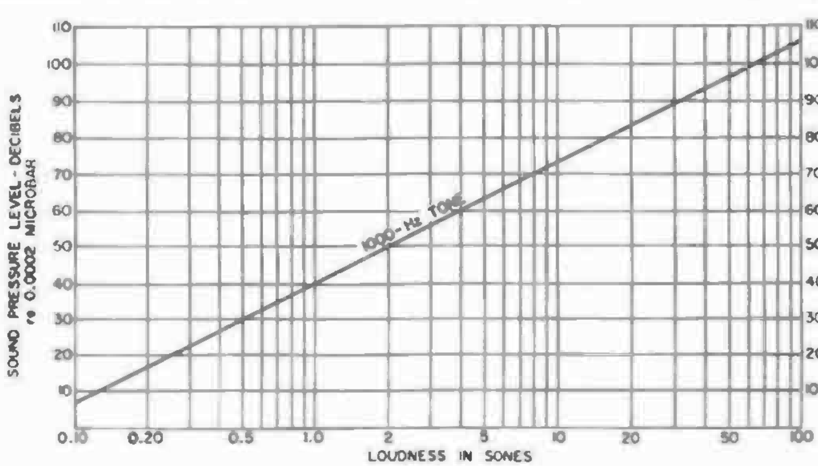


Fig. 1-128. Sound pressure in dB versus sones.

inch. One dyne will exert a pressure of 0.000036 ounce on a surface one centimeter square.

1.126 What are the acoustic pressures generated in average music reproduction?—About 0.005 to 240 dynes per square centimeter.

1.127 What is meant by the term bone conduction?—Sound conduction to the inner ear by the cranial bone rather than by air to the ear drum.

1.128 Define the sone.—It is a unit of loudness graduated in equal steps and is used in the measurement of the human ear characteristics. One sone is equal to 10^{-6} microwatt per centimeter squared. The loudness of a 1000-Hz tone 40 dB above the threshold of hearing is equal to one sone. A tone twice as loud equals two sones, and so on. A millisone is one thousandth of a sone and is often referred to as a loudness unit. A graph for converting decibels to sones and vice versa is shown in Fig. 1-128.

1.129 Define the phon.—The phon is a unit for measuring the loudness level of a pure tone (sometimes called a loudness unit). Because the human ear does not hear on a linear scale, doubling the intensity of a sound does not result in doubling the intensity of the sound at the ear. A true loudness scale would be one that doubles the sensation at the ear when the intensity of the sound is doubled. Such a scale is the loudness scale. Its unit of measurement is the *phon*, with a reference frequency of 1000 Hz.

A graphical plot of phons versus sones appears in Fig. 1-129. A simplified relation between loudness in sones and the loudness level in phons has been standardized internationally (ISO/R131-1959), and is useful to the audio engineer. The relation of sones to phons may be expressed as:

$$S = 2^{\frac{P-40}{10}}$$

where,
S is the loudness in sones,
P is the loudness level in phons.

This relationship is a good approximation to the psychoacoustical data but not accurate enough for research on the subjective aspects of hearing. For example, given a loudness level of 81 phons, in the +1 column in the 80 row, read 17.1 sones.

The loudness scale is employed by comparing the intensity of a tone to the reference frequency. Tones between 800 and 2000 Hz show little difference in loudness. Frequencies that are between 2000 Hz and 8000 Hz show a slight loss. Above a frequency of 8000 Hz, the intensity decreases as the frequency is increased.

Frequencies below 50 Hz require the intensity to be increased 250,000 times to make them equal in loudness to the reference frequency of 1000 Hz. The loudness level in phons of a sound is numerically equal to the sound pressure in decibels relative to 0.0002 microbar at 1000 Hz. A level of 40 phons equals one sone, when referred to 0.0002 microbar.

1.130 What is a summation frequency?—A frequency which is the direct result of two other frequencies being sounded simultaneously, and is the sum of the two frequencies.

1.131 What is the frequency range of musical instruments as compared to the frequency ranges of the human voice, broadcast, and recording systems?

Phons	*0*	*+1*	*+2*	*+3*	*+4*	*+5*	*+6*	*+7*	*+8*	*+9*
20	0.25	0.27	0.29	0.31	0.33	0.35	0.38	0.41	0.44	0.47
30	0.50	0.54	0.57	0.62	0.66	0.71	0.76	0.81	0.87	0.93
40	1.0	1.07	1.15	1.23	1.32	1.41	1.52	1.62	1.74	1.87
50	2.0	2.14	2.30	2.46	2.64	2.83	3.03	3.25	3.48	3.73
60	4.0	4.29	4.59	4.92	5.28	5.66	6.06	6.50	6.96	7.46
70	8.0	8.60	9.20	9.80	10.6	11.3	12.1	13.0	13.9	14.9
80	16.0	17.1	18.4	19.7	21.1	22.6	24.3	26.0	27.9	29.9
90	32.0	34.3	36.8	39.4	42.2	45.3	48.5	52.0	55.7	59.7
100	64.0	68.6	73.5	78.8	84.4	90.5	97.0	104	111	119
110	128	137	147	158	169	181	194	208	223	239
120	256	274	294	315	338	362	388	416	446	478

Fig. 1-129. Table for converting phons to sones or vice versa. (Courtesy, General Radio Co.)

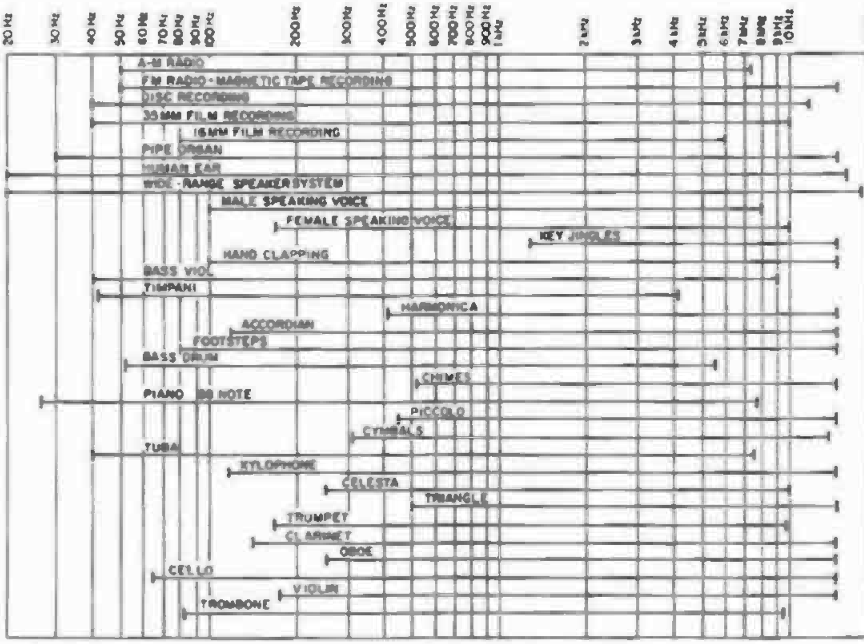


Fig. 1-131. Frequency range required for realistic reproduction of various sound sources.

—A comparison of the frequency ranges of the human voice, recording and broadcast systems, and the human ear is shown in Fig. 1-131. The ranges shown include the fundamental frequencies and their harmonics, with the frequency range indicated for faithful reproduction. Some types of musical instruments generate noise and subharmonics having frequencies below the fundamental frequency. In some instances the fundamental frequency is so low in intensity that it may be filtered out without discerning any change in the characteristics of the instrument. The frequency ranges shown include the more commonly used instruments.

1.132 What are the intensity levels of musical instruments used in an orchestra when referred to the threshold of hearing?

Piano	60 to 100 dB
Organ	35 to 110 dB
Bass drum	35 to 115 dB
Trumpet	55 to 95 dB
Violin	42 to 95 dB
Tympani	30 to 110 dB
Cymbal	40 to 110 dB

The above are intensities measured at a distance of 10 feet from the instrument.

1.133 What value of sound velocity is used in scientific measurements?—For scientific measnrements, the velocity is taken as 1087.42 feet per second, or 331.4 meters per second, at a temperature of 0°C, with zero moisture content at a pressure of one atmosphere. For sound measurements the velocity is taken as 1127 feet per second, or 343.4 meters per second, at a temperature of 20° C. (See Question 1.50.) The speed of sound for any given temperature is:

$$V = \frac{1087 \sqrt{(273 + t)}}{16.52}$$

where,

V is the speed of sound in feet per second,

t is the temperature in degrees Celsius..

1.134 Define the term liveness factor.—This term is associated with the liveness or brilliance of an auditorium or recording stage. Liveness is the sound coming from the source and the reflected sound that an observer hears. The greater the reflected sound, the greater the liveness of the enclosure.

Liveness in recorded or radio sound reproduction creates the illusion that the program is coming from a large auditorium. Liveness may be calculated:

$$L = \frac{1000\,T^2\,D^2}{V}$$

where,

T is the reverberation time of the enclosure in seconds,
D is distance of the sound source from the observer,
V is the volume of the enclosure in cubic feet.

1.135 What is an acoustic shadow? —It is a region of reduced sound pressure, caused by an obstacle in the path of travel of a sound wave. The size and reduction in the intensity of the sound will be governed by the size and shape of the obstacle and the wavelength of the sound wave.

1.136 What is the frequency irregularity of an enclosure?—The irregularity per one-cycle bandwidth expressed in decibels per cycle.

1.137 What is the transmission irregularity of an enclosure?—For an arbitrary band of frequencies, it is the sum of the crest values in decibels, minus the sum of the valley values, also in decibels.

1.138 How much harmonic distortion can the human ear tolerate before the reproduction becomes objectionable? —Experimental data indicate for the modern reproduction, the latitudes shown in Fig. 1-138.

If the high frequencies are cut off at 2750 Hz, up to 15 percent may be tolerated, or about 48 percent intermodulation distortion for the same cutoff frequency.

Adequate phsychological tests have not been made to determine the amount of intermodulation distortion the human ear will tolerate. However, it will suffice to say, amplifiers having low percentage of intermodulation distortion generally sound cleaner than those having a comparable amount of harmonic distortion. For high-quality reproduction, a reproducing system must have less than 1 percent intermodulation distortion. (See Question 1.144.)

1.139 What are the peak powers in watts reached by instruments used in a symphony orchestra?

36-inch bass drum	24.6 watts
15-inch cymbal	9.5 watts
Snare drum	11.5 watts
Piano	0.267 watt
Piccolo	0.084 watt
French horn	0.053 watt
Violin	0.025 watt

As shown above, a bass drum will generate a peak power of 25 watts while a violin will generate only 0.025 watt. This shows the need for an amplifier of considerable power for driving the loudspeaker system, if the recorded material is to be reproduced with realism and low distortion.

To illustrate this point, if a bass drum is struck simultaneously with a violin, 25 watts of power must be handled at the low frequencies while reproducing only 0.025 watt of power for the violin in the higher frequencies. Both these frequencies must be reproduced in their proper perspective covering a frequency range of 30 to 18,000 Hz.

Although a single diaphragm speaker will have difficulty reproducing this type of program material in its true form, it can be reproduced in a fairly satisfactory manner.

The reason this extreme combination of frequencies and powers can be reproduced at all is that the loudspeaker diaphragm responds to different frequencies over various areas of its diaphragm, vibrating around the apex of the diaphragm for the higher frequencies while radiating the lower frequencies from around the areas near the rim of the diaphragm.

In this manner a single-diaphragm type radiator is able to reproduce two or more tones of widely differing powers and frequencies, in more or less the original relationship.

The foregoing discussion illustrates why a multiple speaker system is desirable, because each speaker in the system is confined to a given frequency band. Thus, the greatest efficiency is obtained from each speaker.

	Music	*Speech*
Acceptable	0.7 percent	0.9 percent
Tolerable	1.3 to 1.8 percent	1.9 to 2.8 percent
Objectionable	2.0 to 2.5 percent	3.0 to 4.2 percent

Fig. 1-138. Latitudes in tolerable harmonic distortion.

To smooth out the frequency response where the loudspeakers cross over, a crossover network is used with a tapering cutoff frequency response. This subject is discussed in detail in Section 20.

1.140 What is white sound?—A complex waveform in which the higher frequencies get successively less in amplitude, with steep wavefronts similar to a sawtooth waveform. White sound includes all sounds perceptible to the human ear. A sawtooth waveform consists of a fundamental frequency and even harmonics, while a square waveform includes only the fundamental and odd harmonics. The sawtooth form of sound is called white noise because it is analogous to white light.

1.141 What is pink sound?—White sound was explained in Question 1.140; it contained all frequencies perceptible to the human ear. When the output of a white-noise generator is viewed on an oscilloscope or a graphic level recorder, it displays a rising characteristic of 3 dB per octave, Fig. 1-141A. To bring the response to an equal energy level (uniform output), a pink-noise filter, having an inverse frequency characteristic, is connected in the output of the signal generator. If the output is now mea-

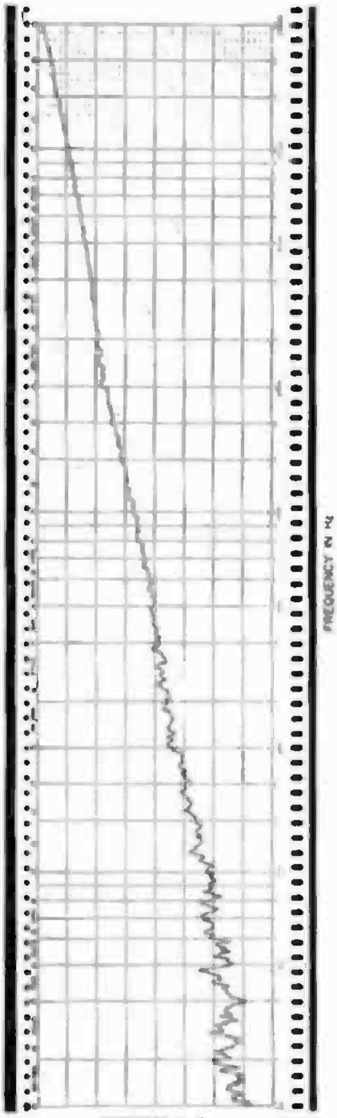


Fig. 1-141A. Output of white-noise generator without pink-noise filter.

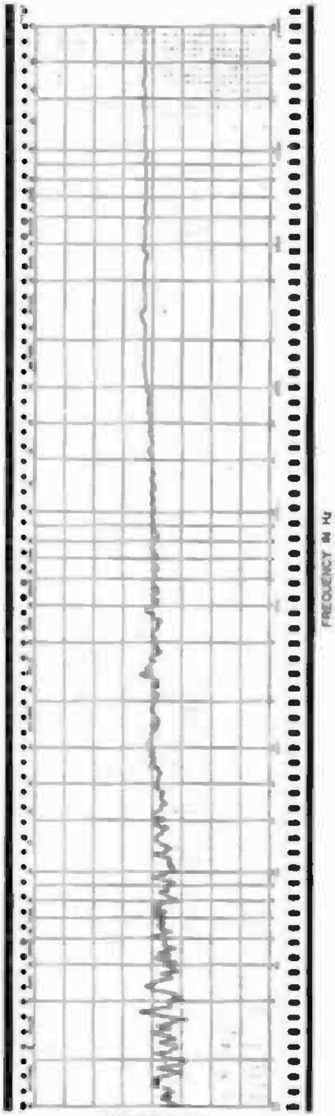


Fig. 1-141B. Output of white-noise generator with pink-noise filter.

sured, it will display a uniform output signal, as in Fig. 1-141B. The signal is now termed pink noise. The sole purpose of the pink filter is for the convenience of measurement. The use of random or white-noise generators is discussed in Sections 22 and 23.

1.142 What is black sound?—It is a term used to denote inaudible sounds. Audible sounds are often referred to as white sound. This latter term is not to be confused with the term white sound as described in Question 1.140.

1.143 What is scale distortion?—When speech or music or both are reproduced in an enclosure such as a room, and at the same acoustic level as the original program material, the reproduced quality will be the same as the original program material.

If the level of reproduction is lowered, a lack of both high and low frequencies will be noted. In some instances a loudness control may be used to correct for this effect. Loudness controls are discussed thoroughly in Question 5.65.

1.144 Explain phase shift, its causes, and the amount that the human ear can tolerate.—Phase shift is caused by the delaying of certain frequencies of a complex waveform in its passage through a sound system or device. The greatest delay is generally at the higher frequencies, particularly harmonics of the fundamental frequency, and is induced by the reactive components of the device or system. If two tones are applied simultaneously to the input and one tone arrives at the output behind the other, phase distortion exists. However, phase distortion is relatively unimportant in a sound system, unless it is great enough to produce a time delay greater than 8 milliseconds at the high frequencies and more than 15 milliseconds below 100 Hz. Phase shift in sound systems is generally measured relative to 1000 Hz. *There is considerable disagreement by engineers, as to how important phase distortion really is.* (See Question 1.138).

1.145 What is listening fatigue and its cause?—The exact cause of listening fatigue is rather vague; however, it is known that such fatigue is not caused entirely by harmonic distortion, because amplifiers having a high degree of listening fatigue will show a low percentage of distortion.

Experience indicates that amplifiers with 1 percent or less harmonic distortion have little effect on the listener. A well-known authority has stated that first order beat-tone intermodulation is of the greatest importance.

This type of distortion may be measured by applying two frequencies not harmonically related to each other to the input of an amplifier and then measuring the sum and difference frequencies at the output. Distortion due to sum and difference frequencies is very annoying because of its nonharmonic relation to the fundamental frequencies of the program material.

1.146 Can sound be transmitted without a medium?—No. This may be demonstrated by the classical experiment of placing an electric bell in an evacuated chamber. If a good vacuum is maintained, no sound will be heard from the bell. If the air is slowly let into the chamber, the bell will be heard, faintly at first and, as more air is let in, the sound of the bell will increase in intensity. With normal atmospheric pressure, the bell will be heard at its normal intensity. The above experiment proves that sound requires a medium for its transmission.

1.147 What is the minimum change in sound level the human ear can detect?—Psychologists have devised various experiments to determine what changes in level can be observed by the average person with good hearing faculties. Under laboratory conditions, when two different levels are presented to the observer, with little time delay between, the observer can detect a difference of 0.25 dB for a 1000-Hz tone at high levels. This sensitivity to change will vary with levels and frequency, but over the range of most interest this differential sensitivity is about 0.25 to 1.0 dB. When the observer is exposed to wide-range random noise (white noise) the detected change is on the order of 0.05 dB, for sound pressures of 30 to 100 dB above 0.0002 microbar (threshold of hearing). Under average conditions, the minimum change likely to be detected is 1.0 dB.

1.148 Define an intertone.—When the human ear hears two tones of nearly the same frequency sounded together, the ear does not recognize them as separate frequencies, but as a single tone. The pitch will lie between the two frequencies, and it is referred to as iutertone.

1.149 Define the term Gaussian noise.—Thermal noise is present in every component of an electronic circuit, and establishes a minimum noise level, under ideal conditions. Thermal noise is said to be white noise and Gaussian. White noise means it has equal power distribution throughout the spectrum, and Gaussian means that the instantaneous magnitude is distributed in accordance with the law of probability, propounded by Karl F. Gauss. (See Question 1.140.)

1.150 What is speed hearing?—It is an electronic device, developed by R. H. Miller of the Bell Telephone Laboratories, which allows hearing of recorded speech at word rates comparable to speed reading. This device uses a harmonic compressor, and has been given to the American Foundation for the Blind. This compressor divides into half the frequency components (harmonics) in a voice recording, while preserving the original time duration. By the doubling of the half-frequency recording, the frequency components are restored to their original values, resulting in normal pitch for a double-speed recording.

The operating principle of the harmonic compressor is as follows: Speech is fed into a bank of 36 bandpass filters, which separate the speech into different frequency components. Output from the filters is fed into 36 frequency dividers, which have the frequencies of the narrow-band signals. The halved-frequency is fed to networks which remove distortion and combine the 36 halved signals into one, where the frequency components are one-half the original input values. This harmonically compressed signal is then recorded on magnetic tape where, by doubling the speed, its halved-frequencies are restored to their original values. Thus the syllabic rate is doubled, without doubling the pitch of the speech.

Speakers who record for the blind speak at an average rate of 160 to 170 words per minute. Doubling the speed without increasing the pitch results in word rates of 320 to 340 words per minute. This compares to average speed-reading rates of 300 to 400 words per minute.

1.151 What is a voice print?—It is a system developed by Lawrence G. Kersta of Bell Telephone Laboratories, for the positive identification of voices, and is similar to the taking of fingerprints. Because each individual develops his own approach to the pronounciation of a given word or sound, and is influenced by the physical characteristics of the vocal cavities and vocal cords, voices can be identified. It is claimed that over 97 percent accuracy was achieved in 50,000 voice prints by this system of identification.

1.152 Can decibels be added or subtracted directly?—No, being a logarithmic value, they cannot be added or subtracted directly. To simplify such operations, the graph in Fig. 1-152 may be used. To add decibels, the graph is entered at Numerical Difference Betweeu the Two Levels Being Added. Follow the line to its intersection with the curved line, then read the Numerical Difference Between Total and Larger Level. Add this value to the larger level to determine the total level. As an example, assume 75 dB and 80 dB are to be combined; the difference is 5 dB. The 5 dB line intersects the curved line at 1.2 dB on the vertical scale. Thus, the total value is 80 + 1.2 or 81.2 dB.

To subtract decibels, enter the graph at Numerical Difference Between Total and Larger Levels, if the value is less than 3 dB. If the value is between 3 and 14 dB, enter the graph at Numerical Difference Between Total and Smaller Levels. Follow the line corresponding to this value to its intersection with the curved line, then either left or downward read Numerical Difference Between Total and Larger (or Smaller) Levels. Subtract this value from the

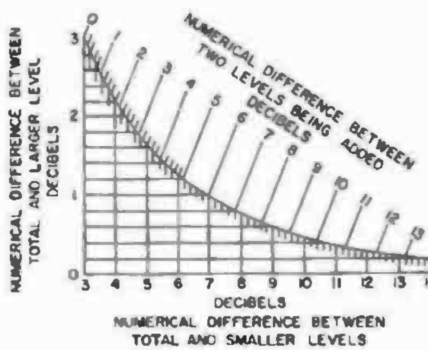


Fig. 1-152. Graph for adding or subtracting decibels (after Musa). (Courtesy, General Radio Co.)

total level to determine the unknown level. As an example, subtract 81 dB from 90 dB; the difference is 9 dB. The 9 dB line intersects the curved line at 0.6 dB on the vertical scale. Thus, the unknown level is 90 minus 0.6 or 89.4 dB.

1.153 What effect does background noise have on acoustic measurements?—Assume a recording stage is being measured, with a ventilating system in operation. How much noise does the ventilating system contribute to the overall noise level? Two measurements are made, one with the system on and one with it off. To ascertain how much the ventilating system contributes, the graph in Fig. 1-152 may be used.

The stage without the ventilating system measures 33 dB (using a sound level meter) and with the ventilating system operating, 37 dB. The difference is 4 dB. Referring to the chart and entering the bottom at a value of 4 dB, following the vertical line to where it intersects the curved line, at the left margin is read 2.2 dB. This value is subtracted from the total measurement (37 dB) leaving a total noise level of 34.8 dB. Therefore the SPL is only increased 2.2 dB when the ventilating system is in use.

1.154 Define the term "G."—The quantity "G" is the acceleration produced by the force of gravity, which varies with latitude and elevation of the point of observation. By international agreement, the value 9.80665 cm/sec^2 equals 386.087 in./sec^2 equals 32.1739 ft/sec^2, has been chosen as the standard acceleration of gravity.

1.155 What is the relationship between sound pressure level (SPL) and acceleration in G's?—A table of sound pressure levels (SPL) in decibels (re 0.0002 microbar) relative to the acceleration in G's is given in Fig. 1-155.

Level in dB	*Accel in G's*	*Level in dB*	*Accel in G's*
44	.000398	92	0.100
45	.000447	93	.112
46	.000501	94	.126
47	.000562	95	.141
48	.000631	96	.159
49	.000708	97	.178
50	.000794	98	.200
51	.000891	99	.224
52	.00100	100	.251
53	.00112	101	.282
54	.00126	102	.316
55	.00141	103	.355
56	.00159	104	.398
57	.00178	105	.447
58	.00200	106	.501
59	.00224	107	.562
60	.00251	108	.631
61	.00282	109	.708
62	.00316	110	.794
63	.00355	111	.891
64	.00398	112	1.00
65	.00447	113	1.12
66	.00501	114	1.26
67	.00562	115	1.41
68	.00631	116	1.59
69	.00708	117	1.78
70	.00794	118	2.00
71	.00891	119	2.24
72	.0100	120	2.51
73	.0112	121	2.82
74	.0126	122	3.16
75	.0141	123	3.55
76	.0159	124	3.98
77	.0178	125	4.47
78	.0200	126	5.01
79	.0224	127	5.62
80	.0251	128	6.31
81	.0282	129	7.08
82	.0316	130	7.94
83	.0355	131	8.91
84	.0398	132	10.0
85	.0447	133	11.2
86	.0501	134	12.6
87	.0562	135	14.1
88	.0631	136	15.9
89	.0708	137	17.8
90	.0794	138	20.0
91	.0891	139	22.4
		140	25.1

Fig. 1-155. Relationship of sound pressure level (SPL) and acceleration in G's. (Courtesy, General Radio Co.)

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Section 2

Acoustics, Studio Techniques, and Equipment

While acoustical engineering is a science within itself, it plays an important role in the daily routine of the audio engineer. This section sets forth the basic principles of acoustics, as applied to sound waves in open air or enclosures. It also deals with treatments of rooms and stages used for recording and reproduction, reverberation characteristics, noise reduction coefficients of acoustic materials, and the shape of enclosures for optimum characteristics. Design factors are given for dubbing and looping stages, soundlocks, monitor rooms, projection facilities, reverberation chambers, and auxiliary equipment, diffusers, flats, microphone booms, tempo regulators, and synthetic reverberation units. Techniques for setting-up stages, and microphone placement for small intimate groups or large symphonic orchestras are discussed. Explanations of anechoic and reverberant chambers, Rayleigh disc, Helmholtz resonator, weighted curves, ambiophonic reproduction, wave trains, sound power, and sound levels are given.

2.1 ***Define the term acoustics.***—It is a science dealing with the production, effects, and transmission of sound waves; the transmission of sound waves through various mediums, including reflection, refraction, diffraction, absorption, and interference; the characteristics of auditorium, theaters, and studios, as well as their design.

2.2 ***Define acoustic impedance.***—It is the force per unit area on a given surface of a sound medium divided by the flux through that surface. Expressed in ohms, it is equal to the mechanical impedance divided by the square of the surface. The unit of measurement is the acoustic ohm.

2.3 ***Define acoustic ohm.***—A unit of acoustic resistance. It is equivalent to a sound pressure of one dyne per square centimeter producing a volume velocity of one cubic centimeter per second. It is also used when referring to acoustic impedance or reactance.

2.4 ***What are the preferred frequencies for acoustical measurements?***—In January 1960, The Acoustical Society of America sponsored the USASI(ASA) Standard S1.6-1960, covering the frequencies recommended for acoustical measurements. These frequencies are given in Fig. 2-4 and the order of preferences is indicated by the size and style of type. It will be noted that the order frequencies used for measurements, such as 256, 512, 1024, etc., have now been changed to 250, 500, 1000 Hz, etc.

Frequencies for use with bandpass filters are also given and are geometric center frequencies of the bands. For example, the lower and upper cutoff frequencies for an octave band filter centered on 8000 Hz are, respectively, 5600 Hz and 11,200 Hz, these being the preferred frequencies in the series of half-octave intervals. For certain measurement purposes, it may be convenient to depart slightly from the regular geometric series in order to obtain the nearest round-number approximation.

For audiometry, (see Question 22.42) in addition to octavely spaced frequencies, such as 1000, 2000, and 4000 Hz, the frequencies 3000 and 6000 Hz have been used rather than 2800 and 5600 Hz or 3150 and 6300 Hz.

2.5 ***What is an acoustic labyrinth?***—A specially designed baffle arrangement for use with a loudspeaker to re-

Preferred Frequencies	*1*	*1/2*	*1/3 Octave*	*Preferred Frequencies*	*1*	*1/2*	*1/3 Octave*	*Preferred Frequencies*	*1*	*1/2*	*1/3 Octave*
16	X	X	X	160			X	1600			X
18				180		X		1800			
20			X	200			X	**2000**	X	X	X
22.4		X		224				2240			
25			X	**250**	X	X	X	2500			X
28				280				2800		X	
31.5	X	X	X	315			X	3150			X
35.5				355		X		3550			
40			X	400			X	**4000**	X	X	X
45		X		450				4500			
50			X	**500**	X	X	X	5000			X
56				560				5600		X	
63	X	X	X	630			X	6300			X
71				710		X		7100			
80			X	800			X	**8000**	X	X	X
90		X		900				9000			
100			X	**1000**	X	X	X	10000			X
112				1120				11200		X	
125	X	X	X	1250			X	12500			X
140				1400		X		14000			
160			X	1600			X	**16000**	X	X	X

Fig. 2-4. Table of preferred frequencies in Hz, at various intervals, for acoustical measurements and for center frequencies of filter passbands.

inforce the low-frequency response and prevent cavity resonance. Such a device is shown in Fig. 20-66.

A labyrinth is not always in the form of a baffle, but could be a tube filled with hair-felt, as is sometimes used in microphones.

2.6 ***What is an acoustic line?***—An acoustic equivalent of a sound chamber at the rear of a loudspeaker.

2.7 ***What is acoustic response?***—It is a measurement of the *reverberation* characteristics of an enclosure, which might be an auditorium or stage.

2.8 ***What is an acoustic pickup or sound box?***—A nonelectric pickup for reproducing disc records. A needle or stylus is connected by mechanical linkage to a mica or dural diaphragm. Flexing of the diaphragm, caused by the movement of the needle in the sound track of the record, disturbs the air in a horn to which the sound box is coupled. Movement of the air in the horn produces sound waves which are heard by the ear.

2.9 ***What is an acoustical equalizer?***—A small metal tube at the rear of a microphone to release the pressure behind the diaphragm, thus preventing mechanical distortion of the diaphragm which would, in turn, produce electrical distortion.

2.10 ***What is acoustic treatment?***—The application of acoustic or sound-absorbing material to a room or enclosure to obtain the desired acoustic characteristics.

2.11 ***What is an acoustic feedback?***—An audible howl or singing noise caused by sound waves feeding back from a loudspeaker to a microphone. It is generally caused by placing a loudspeaker too close to a microphone. Acoustic feedback can also be caused by sound leaking through air ducts.

2.12 ***What is meant by the term "a brilliant stage"?***—A stage in which the high frequencies predominate. Such stages are also referred to as being live or hot. This characteristic is caused by hard or reflective surfaces of the stage, such as ceilings, walls, floors, etc.

2.13 ***Define the term noise-reduction coefficient.***—It is the attenuation afforded by the acoustic treatment in an enclosure by the materials involved. In computing the reverberation time of an auditorium, stage, or theater, the absorption coefficients at a single frequency of 500 Hz is used (originally 512 Hz was used). In rating the effectiveness of absorbents in the reduction of room noise, the average coefficients of frequencies of 250, 500, 1000, and 2000 Hz are used. The average is then termed the noise-reduction coefficient or NRC. (See Question 2.4.)

2.14 *What is the effect of a high ambient noise level on human beings?*—Workers become irritable and fatigued; also, permanent injury to the hearing may result.

2.15 *What frequencies are most annoying to the human being?*—Frequencies above 2000 Hz.

2.16 *What is a flutter echo?*—A multiple echo in which the reflections occur in rapid succession. If the echo is periodic and in the audible range, it is referred to as a musical echo.

2.17 *What does the term "tubby" mean?*—Reproduction lacking in definition, or an accentuation of the low frequencies resulting in a barrellike reproduction.

2.18 *What is a dead room?*—One in which an overamount of sound-absorbing material has been used so that most of the high frequencies are absorbed. The reproduction from a room of this type will be dull and lacking in presence.

2.19 *What does the term "hangover" mean?*—Acoustically, it is undesirable reflections causing excessive echoes. In an amplifier, hangover is caused by a low internal damping factor. (See Question 12.177.)

2.20 *What is masking?*—The inability of an auditor to hear certain sounds because of the presence of other sounds. Masking is most noticeable at the higher frequencies.

2.21 *What takes place when a sound wave is reflected?*—When a sound wave is traveling through a medium such as air and encounters another medium such as water, cold air, or a solid object, and the second medium is larger in comparison to the wavelength of the emitted sound, part of the sound is reflected back from the object in a manner similar to a beam of light. The balance of the sound is absorbed into and transmitted by the second medium.

If the sound wave strikes the second medium at an angle, a large part will bounce off and will be reflected at an angle which is exactly equal to the angle of incidence. See Fig. 2-21A. If the emitted sound wave is in an enclosure similar to that shown in Fig. 2-21B, an observer situated as shown will hear reflected sound as well as the direct sound from the sound source.

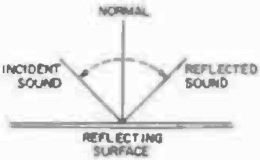

Fig. 2-21A. Reflected sound.

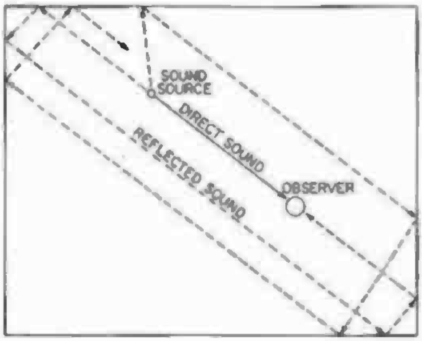

Fig. 2-21B. Reflected and direct sound in an enclosure.

Reflected sound has the effect of increasing the intensity, causing out of phase conditions, and adding reverberation. All of the above effects vary with the acoustical treatment of the enclosure. If the reverberation is excessive, the intelligibility is reduced.

2.22 *What is confusion or scattering?*—The jumbling of sound waves resulting from interference caused by objects in the path of transmission producing unintelligibility.

2.23 *Define interference.*—Interference is caused by sounds coming from different directions, or by reflection and mixing with the original sound. Under such conditions, the intensity of the sound may be increased or decreased, depending on the phase relationship of the waveforms at any given instant.

2.24 *What is rarefaction?*—The state of being less dense. The opposite of compression.

2.25 *Define diffraction.*—When sound encounters an object in its normal path of travel, it bends around the object, causing eddy currents behind the object. This is diffracted sound. Low frequencies bend around an obstacle more easily than high frequencies. This phenomenon is also called scattering.

2.26 *What is refraction?*—It is a change in the direction of a sound wave, caused by the nature of the medium of transmission. This can be caused by air temperature, since the

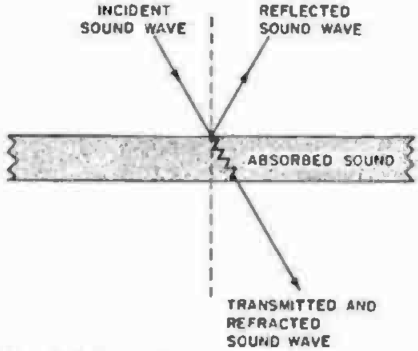

Fig. 2-26. Incidence, reflection, absorption, and transmission of a sound wave striking a flat surface. The term refracted is also applied to the transmitted sound wave because of its change of direction.

velocity of sound increases as the temperature increases. Refraction also takes place when a sound wave strikes a surface such as water or a wall, as shown in Fig. 2-26.

2.27 ***What is the angle of refraction?***—The angle measured between a perpendicular erected at the point of contact with a surface and a wave, ray, or beam refracted from that surface.

2.28 ***What is dispersion?***—The separation of a complex sound wave into its frequency components, caused by a change in velocity. This action is analogous to sunlight being passed through a prism. (See Fig. 2-28.)

2.29 ***What is acoustic absorptivity?***—The ratio of sound energy absorbed by the surface of a given material to that which arrives at the surface from the source. A porous material will break up a wave train, slow the waves down, and, finally, absorb them. The action is similar to water on blotting paper.

2.30 ***What standard is used for comparison when rating the coefficient of absorption of acoustic materials?***—One square foot of air, free from reflections.

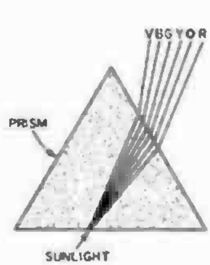

Fig. 2-28. Dispersion of sunlight through a prism.

2.31 ***Is the absorption coefficient of acoustic materials the same for all frequencies?***—No. It is not constant and varies with frequency and the angle of incidence. Generally, the data given for acoustic materials are for a frequency of 500 Hz. However, data for other frequencies are available from the manufacturer of the material. A table of absorption coefficients is given in Question 2.32.

2.32 ***What are the absorption coefficients for general building materials and furnishings?***—See Fig. 2-32.

2.33 ***What is a sabin unit?***—A unit of absorption equivalent to the absorption of 1 sq. ft. of surface which will absorb all incident energy. The unit is named for its originator, Wallace C. Sabine.

2.34 ***Define the term reverberation period.***—It is the time required for a sound in an enclosure to die away to one-millionth of its original intensity, or decrease 60 dB. The reverberation time of any enclosure may be calculated by the formula:

$$T = \frac{V \times 0.049}{AS}$$

where,

- T is the reverberation time in seconds,
- V is the volume of the enclosure in cubic feet,
- A is the average absorption coefficient of the enclosure,
- S is the total surface area in square feet.

Reverberation is the persistence of sound within an enclosure after the original sound has ceased. Reverberation may also be considered as a series of multiple echoes, decreasing in intensity, so closely spaced in time as to merge into a single continuous sound and eventually be completely absorbed by the treatment of the enclosure and to a degree, by dissipation of the energy into the air. However, this latter factor is generally ignored and only the wall treatment is considered.

If a loudspeaker is placed in a room and a continuous frequency applied, a wave train will be built up, spreading in all directions. Upon striking the boundaries of the room, the wave train is partially absorbed and partially re-

Material	125 Hz	250 Hz	500 Hz	1000 Hz	2000 Hz	4000 Hz
	Coefficients					
Brick, unglazed	0.03	0.03	0.03	0.04	0.05	0.07
Carpet, heavy on concrete	0.02	0.06	0.14	0.37	0.60	0.65
Carpet, with latex backing on 40-oz hairfelt of foam rubber	0.08	0.27	0.39	0.34	0.48	0.63
Concrete block, coarse	0.36	0.44	0.31	0.29	0.39	0.25
Light velour, 10 oz per sq-yd in contact with wall	0.03	0.04	0.11	0.17	0.24	0.35
Concrete or terrazo	0.01	0.01	0.015	0.02	0.02	0.02
Wood	0.15	0.11	0.10	0.07	0.06	0.07
Glass, large heavy plate	0.18	0.06	0.04	0.03	0.02	0.02
Glass, ordinary window	0.35	0.25	0.18	0.12	0.07	0.04
Gypsum board, nailed to 2 by 4 studs on 16-inch centers	0.29	0.10	0.05	0.04	0.07	0.09
Plaster, gypsum, or lime, smooth finish on tile or brick	0.013	0.015	0.02	0.03	0.04	0.05
Plywood, ⅜-inch	0.28	0.22	0.17	0.09	0.10	0.11
Air, Sabins per 1000-cu. ft.	—	—	—	—	2.3	7.2
Audience, seated in upholstered seats, per sq. ft. of floor area	0.44	0.54	0.60	0.62	0.58	0.50
Wooden pews occupied, per sq. ft. of floor area	0.57	0.61	0.75	0.86	0.91	0.86
Chairs, metal or wooden, seats unoccupied	0.15	0.19	0.22	0.39	0.38	0.30

Coefficients above were obtained by measurements in the laboratories of the Acoustical Materials Association. Coefficients for other materials may be obtained from Bulletin XXII of the Association.

Fig. 2-32. Absorption coefficients for different materials.

flected not once but hundreds of times. Thus, the average intensity of the sound is built up to a steady state, in which the rate of emission just equals the rate of absorption at the boundaries. This indicates that time is required to set the body of air in an enclosure in motion.

If the source of sound is now cut off, the sound does not cease immediately, but generally dies away. The average intensity at any one instant decreases at a rate which is proportional to the average intensity at that instant. This indicates the logarithm of the average intensity is decreasing at a uniform rate, or the drop in intensity level expressed in decibels is proportional to the time measured at the instant of cutoff of the sound source.

Fig. 2-34C shows an oscilloscope display of a decay recorded in an enclosure. It can be observed that the decrease in amplitude with time follows approximately the average curve of the right half of Fig. 2-34A, but with a larger fluctuation thereafter.

Reverberation has considerable ef-

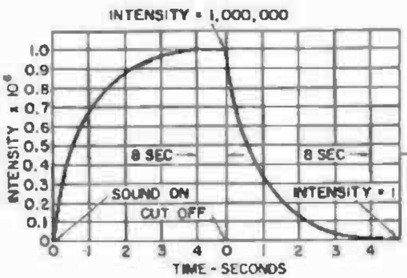

Fig. 2-34A. Typical decay curve. Plotted time versus intensity.

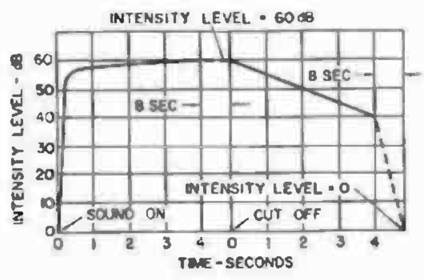

Fig. 2-34B. Curve of Fig. 2-34A plotted time versus intensity in decibels.

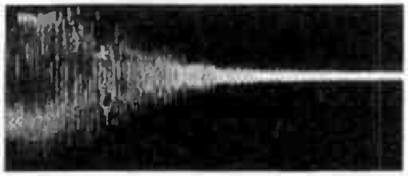

Fig. 2-34C. Oscilloscope display corresponding to Fig. 2-34A.

fects on speech. If the auditor is close to the speaker, no great difficulty will be experienced in clearly understanding the speaker. If, however, the speaker raises his voice, each syllable is prolonged, running into succeeding syllables, with resulting confusion and loss of intelligibility. For music, individual notes are prolonged by the reverberation and have the effect of a piano played with the loud pedal held down continuously. To arrive at the reverberation time for a given enclosure, the area of each surface in the enclosure is multiplied by its absorption coefficient, and the sum of these, plus the absorption due to objects such as chairs, drapes, people, etc, gives the average absorption coefficient in the formula.

2.35 ***What is the recommended reverberation time for theaters, auditoriums, recording and broadcast studios?***— A graphical presentation of the recom-

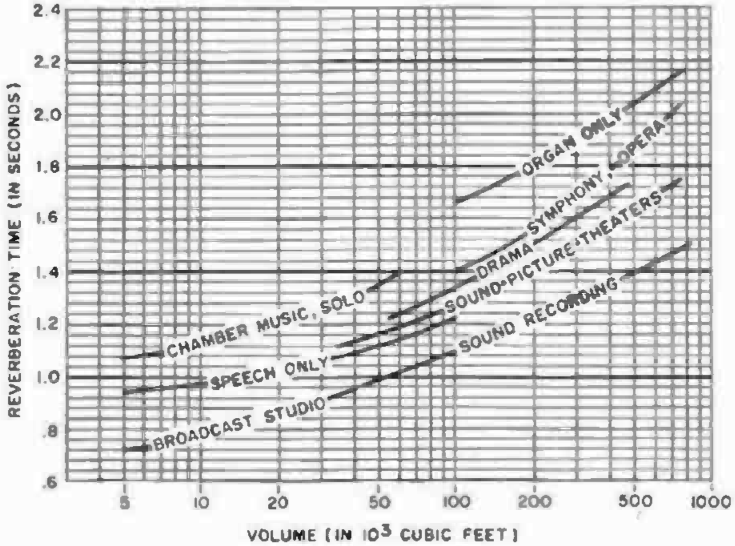

Fig. 2-35A. Recommended reverberation time for various types of auditoriums, at a frequency of 512 Hz.

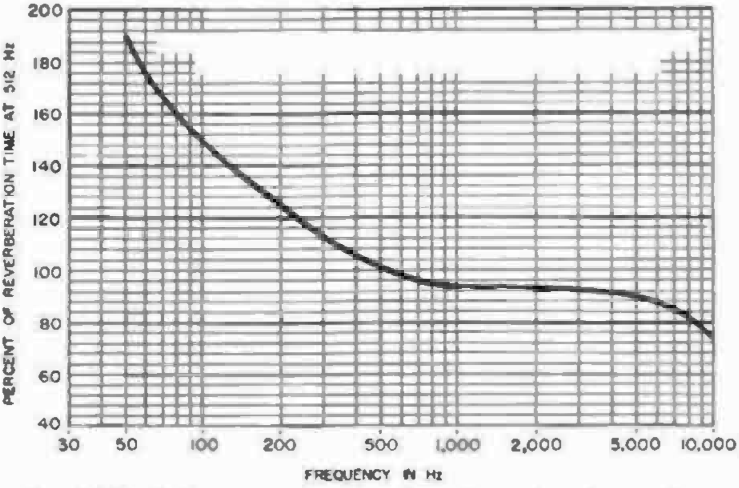

Fig. 2-35B. Optimum reverberation times for various frequencies.

mended reverberation time for various types of enclosures is shown in Fig. 2-35A. Fig. 2-35B shows the optimum reverberation time for different frequencies. In Fig. 2-35C is plotted the optimum reverberation versus room volume in cubic feet, at 512 Hz. (See Question 17.136).

2.36 ***Define optimum reverberation time, and tell how it is governed.***—Optimum reverberation time is the most desirable reverberation time for an enclosure of given dimensions. This is governed by the cubic volume of the enclosure and the absorption factors of the walls, ceiling, floor, and other furnishings.

2.37 ***How is artificial reverberation added to an auditorium, and what is the purpose?***—The basic purpose of adding artificial reverberation to an existing auditorium or stage is to control the acoustics electronically, and improve the overall acoustics of the enclosure.

It is common practice in broadcast and television studios to overtreat the studio to reduce the noise created by the movement of equipment and actors around the stage. If the treatment is carried to the extreme, as it sometimes is, the sound reproduction is flat to both the listener in the studio and over the air. It also has a pronounced effect on the musicians as the reproduction of their instruments does not sound normal and this leads to difficulties. To overcome this difficulty and still retain the heavy acoustic treatment, controlled electronic reverberation is induced into the enclosure or stage. This electronic system of reverberation is termed ambiophony, and has been used quite successfully by the British Broadcasting Co., and in several auditoriums and music halls in the United States.

Artificial reverberation can be achieved in several ways. One system employs an electromechanical device, using amplifiers and rods or springs for the delay networks. The electroacoustic system is similar, except echo chambers are used for the delay. A third system employs magnetic recording techniques. The signal is recorded on an endless magnetic tape, and the signals from several playback heads spaced at various distances from the recording head are combined. The all-electronic system employs logic circuits to provide the delay.

A typical installation might consist of as many as 60 loudspeakers placed around the walls and ceiling of an enclosure. The speakers radiate the orchestral music, through the time delays, and radiate about as much sound as the walls normally would if they were not highly absorbent. Each speaker has a time delay of such a value that it radiates at the approximate time it would take the sound to reach that particular speaker position in the studio. The randomness of true reverberation is achieved by not connecting the speakers according to the delay appropriate

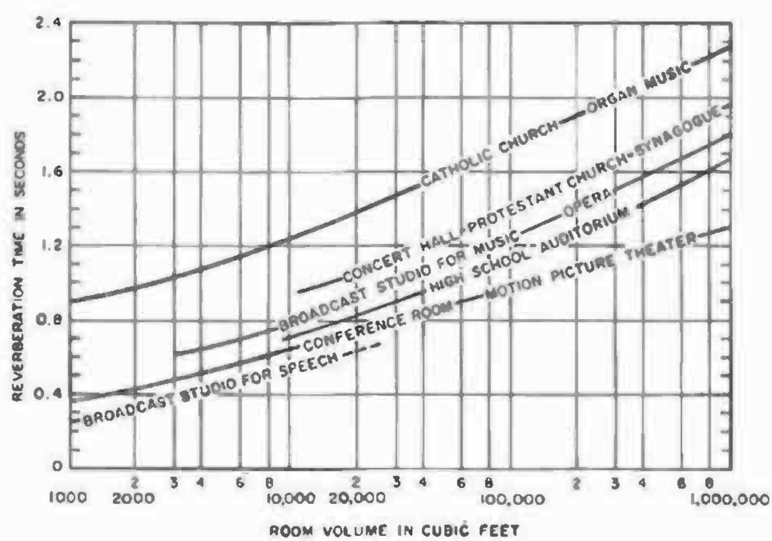

Fig. 2-35C. Reverberation time versus cubic volume, at 512 Hz.

to their exact position, although they generally approach their true position.

Because some regeneration is picked up by the microphones supplying the input to the reverberation system, care must be taken in adjusting the level of the system. Two completely separate microphone systems are used for feeding the broadcast and ambiophony systems, with the necessary equalization. Such installations are still in the experimental stage and each installation becomes a specialized system. Magnetic reverberation units are discussed in Question 17.136.

2.38 What requirements constitute an ideal studio?—According to Wallace C. Sabine, an outstanding authority in the field of acoustics and architectural design, they are: (1) that the sound in the studio be sufficiently loud; (2) that the components of a complex sound wave maintain their relative intensities; (3) that successive, rapidly moving sounds, either music or speech, be clear and distinct from each other; and, (4) that all extraneous noises must be reduced to a negligible amount.

2.39 What shape rooms are to be avoided when constructing enclosures for recording and reproduction of sound?—Flat, untreated surfaces; concave surfaces, as they concentrate and focus the sound; parallel walls, because they produce standing waves; and cubical shaped enclosures, because they produce standing waveforms. (See Questions 2.41 and 2.49.)

2.40 What are the preferred ratios of dimensions for studios and auditoriums?—The height, width, and length should be in the following ratios:

1. Small rooms ....1 to 1.25 to 1.6
2. Long rooms ....1 to 1.25 to 3.2
3. Average shaped rooms 1 to 1.6 to 2.5
4. Low ceiling rooms ...1 to 1.25 to 3.2

Number 3 indicates the preferred dimensions.

To minimize the effects of standing waveforms set up by parallel surfaces in an enclosure, it is desirable to choose the major dimensions that are not integral to each other. Resonance effects in an enclosure introduce frequency discrimination and create peaks and valleys in the characteristics of the room. Secondly, it introduces a persistence or hangover effect in the sound at or near resonance. Frequency discrimination results in a hollow sounding characteristic, especially if the resonant frequencies are widely separated (walls close together) which is generally the case in small rooms. The effects of resonance may be reduced by absorption, by changing the dimensions of the room, and by changing the shape of the reflecting surfaces.

By proportioning the three major dimensions in a ratio of the cube root of two (or multiple), good distribution of the natural resonances may be obtained. As an example using a ratio of 1:1.6:2.5 (average shaped room) for a ceiling height of 16 feet, the width is 25.6 feet (16×1.6) and it is 40 feet in length (16×2.5). For a small room using ratios of 1:1.25:1.6 with a ceiling of 12 feet, the dimensions are 12 ft $\times$ 15 ft $\times$ 19.2 ft, whereas in an average room the ratios change and the dimensions are then 12 ft $\times$ 19.2 ft $\times$ 30 ft. Diffusion and control of the high and low frequencies may be obtained by the use of polycylindrical diffusers explained in Question 2.76.

2.41 What are standing wave trains?—When a sustained tone is emitted in an enclosure consisting of parallel walls, a standing wave train is set up. Standing waves are created when two wave trains, moving in opposite directions, interfere. Walking along the room produces the sensation of an increase and decrease in the intensity of the sound. This sensation is noted because of passing through the zero and maximum peaks of the wave.

When a reflected waveform exactly matches compression with a rarefaction of the original sound wave, the sound waves reinforce themselves as they are reflected back and forth, thus increasing the amplitude. Reinforcement at critical frequencies can result in an increase of 20 dB or more. Serious reinforcement can occur when the wavelength is twice the ceiling height. Standing waveforms can also be generated at harmonic frequencies of the fundamental frequency.

Standing waves may be prevented by nonparallel walls, multilevel ceiling sections, and polycylindrical diffusers on the walls and ceilings, as in Fig. 2-66B.

2.42 What is the effect of a long reverberation time?—Both speech and music will be blurred and may become unintelligible, because of the overlapping of successive sounds.

2.43 Explain the purpose of diffusion in a studio. Diffusion in a studio improves the acoustical response because the energy in the room is not reduced and the reflections which occur per unit of time are increased. Thus, the intensity level of the individual reflections is reduced, and the reverberation characteristic smoothed out, resulting in a higher intelligibility and an added definition to both music and speech.

2.44 What is the frequency range required for high quality reproduction of speech and music?—For speech only, 100 to 6000 Hz with a volume range of 40 dB. For music, 40 to 15,000 Hz, with a volume range of 70 dB. Most audio systems designed for high-quality reproduction will reproduce up to 20,000 Hz and higher, some extending up to 100,000 Hz. The question then arises: why such a wide frequency range? A wide frequency range is required for music to reproduce inaudible frequencies which beat with frequencies in the audible range, producing sum and difference frequencies. Such frequencies lend realism to the reproduction. However, to make use of such wide frequency bauds, the harmonic distortion and intermodulation distortion must be reduced to negligible amounts. Also, the distortion due to phase shift must be at a minimum and the frequency characteristics uniform.

2.45 What is the recommended cubic footage per person for 35-mm and 70-mm motion picture theater projection?—In the past years the cubic footage per person recommended for theaters projectiug 35-mm films was approximately 125 cubic feet per person. However, with the advent of wide-screen projection systems and stereophonic sound reproduction, the space per person has been increased 250 percent or more. This is particularly true for theaters built for 70-mm projection.

Seating arrangements used in North America generally employ a layout whereby the aisles are placed down the center of the seating area or about one-third from the side walls. In this arrangement, erroneous localization of stereophonic sounds often results in the picture action and sound not coinciding. This has been overcome to a great extent by the use of *Continental* seating arrangements (as used in Europe) whereby the aisles are placed along the *side walls* rather than in the seating area.

Using the above arrangement optical distortion of the picture is reduced, and better sound reproduction is obtained for those seated at the sides of the theater. This has also led to the redesigning of the floor rise, and spacing of the seats. Nonparallel walls and special treatment of the side and rear walls reduce the effect of flutter echoes. Details of this type seating arrangement are given in the reference.

Review rooms and small theaters used on motion picture lots where the seating capacity is between 20 and 40 people generally employ about 250 to 500 cubic feet per person.

2.46 What effect does an audience have on the acoustics of a theater?—Unless the house has been specially designed, the effect may be considerable. The projectionist generally increases or decreases the reproduced sound level when the audience decreases to about half-house. Modern theaters have overcome this problem by using seats which have an absorption coefficient equivalent to the average person, thus reducing the need for frequent changes in the sound level.

2.47 What amplifier power is recommended for motion picture theaters?—The power requirements are given on the graphs of Fig. 2-47A and Fig. 2-47B. The shaded portion of the curve indicates the minimum and maximum recommendations. Since the average theater amplifier system must have a fairly wide dynamic range and be capable of handling heavy sound effects, an amplifier on the heavy side should be selected rather than one that will just meet the requirements. Present-day theaters require at least 40 watts of power with low distortion and noise. It is not uncommon in the larger theaters to find amplifier installations with 100 to 250 watts of power output. Power output versus seating capacity of a theater is shown in Fig. 2-47A, and power versus volume in cubic feet is shown in Fig. 2-47B. These data are based on recommendations of the Motion Picture Research Council.

2.48 Define acoustic reflectivity (reflective coefficient).—The acoustic re-

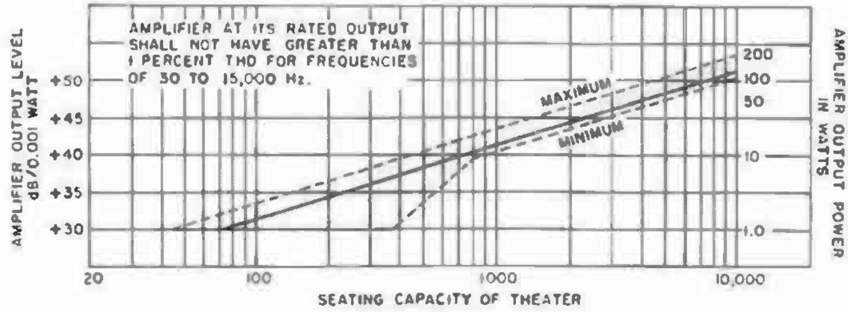

Fig. 2-47A. Relationship of optimum amplifier power output versus theater seating capacity.

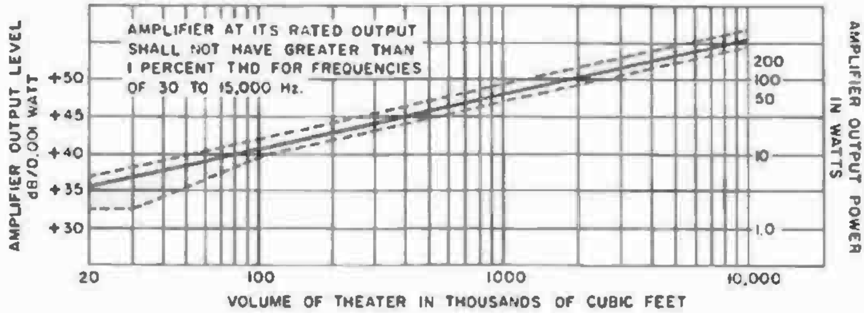

Fig. 2-47B. Relationship of optimum amplifier power-output level to theater volume in cubic feet.

flectivity of a surface is not a generator; it is the ratio of the rate of flow of sound energy reflected from a surface to that of the incident rate of flow, and may be calculated:

$$\frac{\text{incident rate of flow of sound energy}}{\text{reflected rate of flow of sound energy}}$$

Unless otherwise specified, all possible directions of incident flow are assumed equal, and values given apply to a portion of an infinite surface thus eliminating edge effects.

2.49 ***What shape walls are recommended for theater construction?***—The walls should be nonparallel or convex shaped, as shown in Fig. 2-49. The cubical volume should be in accordance with the seating capacity recommendations given in Question 2.45.

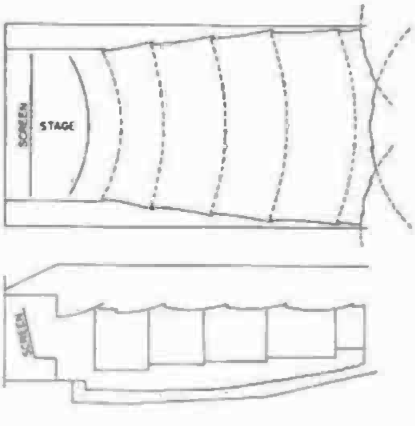

Fig. 2-49. Recommended wall shapes for motion-picture theater construction.

The auditorium width should be from 50 to 70 percent of the length and the ceiling height not more than 40 percent of the length. Nonparallel surfaces should be employed. The walls and ceilings must be broken up thoroughly to diffuse the sound.

The average absorption per square foot of floor space should be the same as for the ceilings and walls. The seats should be well upholstered and the aisle carpets Ozite-lined. The backstage area should be so shaped and acoustically treated that resonant reinforcements of sound will not be reflected to add distortion to the reproduction in the auditorium.

2.50 ***How are low-frequency vehicle and earth rumbles prevented from being***

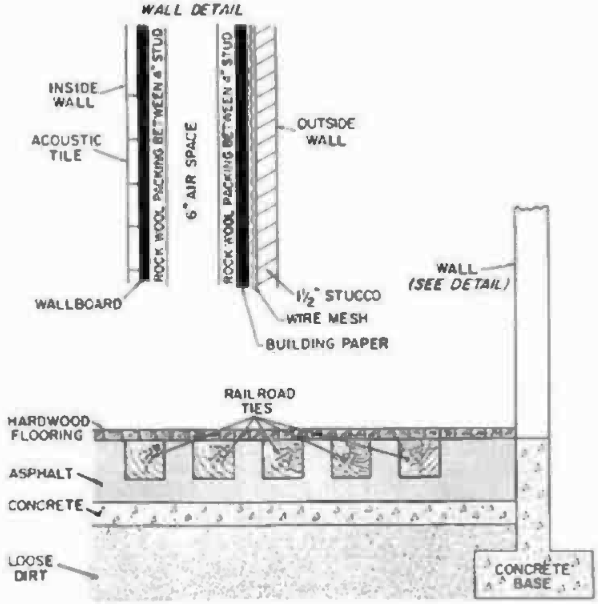

Fig. 2-50A. Cross-sectional construction of a motion-picture sound-stage wall, floor, and foundation.

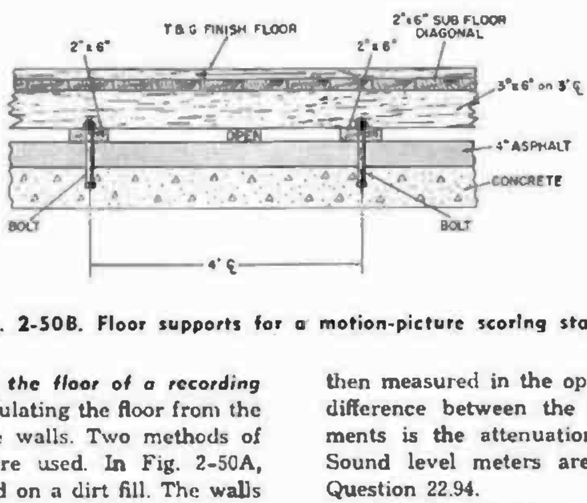

Fig. 2-50B. Floor supports for a motion-picture scoring stage.

transmitted to the floor of a recording stage?—By insulating the floor from the earth and side walls. Two methods of construction are used. In Fig. 2-50A, the floor is laid on a dirt fill. The walls are supported on rockers set on a cement foundation. In Fig. 2-50B, the floor is supported on 3×6-inch beams supported by 2×6-inch members laid on an asphalt fill on a cement foundation.

The interior construction of a typical scoring stage is described in Question 2.66.

2.51 How is the attenuation of the wall illustrated in Fig. 2-50A measured?—A Klaxon horn is placed against the outside wall and the sound transmission through the wall is measured with a sound level meter at a point near the wall. The sound level of the Klaxon is then measured in the open air and the difference between the two measurements is the attenuation of the wall. Sound level meters are discussed in Question 22.94.

2.52 Describe the use of lead sheeting for acoustic treatment.—For many years, it has been known that lead can be used in the acoustical treatment of an enclosure, and that the greater the weight per square foot of isolation, the greater the transmission losses. The acoustical efficiency of any material as a sound barrier depends not only on its weight, but on its stiffness. Lead is classed as a heavy limp material, having a density of two to three times that of most building materials, and 10 to 15 times that of wood. Lead is a limp material, in an acoustical sense. It has been demonstrated that if two equally effective barriers are constructed, one of lead and the other of conventional building materials, the lead barrier will be the lighter of the two.

If the transmission loss of a lead partition is plotted against frequency (Fig. 2-52A) it will show the loss increases with frequency up to the point where the transmission loss reaches approximately 55 dB. Above this frequency, a dip in the transmission loss occurs. At the lower frequencies, the loss is set by the

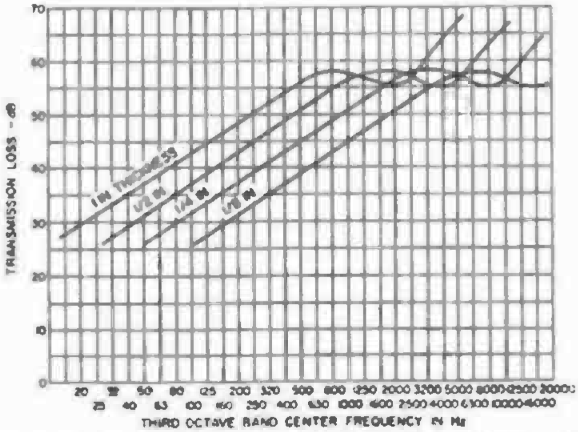

Fig. 2-52A. Transmission loss through solid chemical lead walls. (Courtesy, Lead Industries Association)

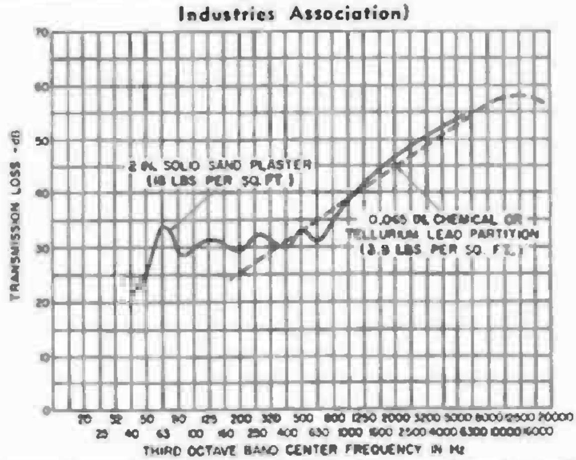

Fig. 2-52B. Comparison of a 2-inch solid sand-plaster wall and a wall of 0.065-inch lead. (Courtesy, Lead Industries Association)

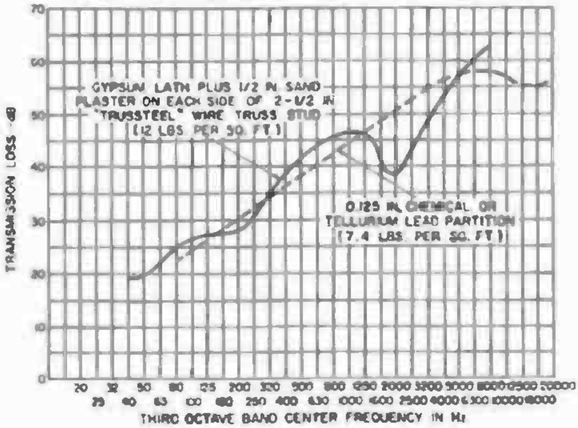

Fig. 2-52C. Comparison of plaster wall and 0.125-inch lead wall. (Courtesy, Lead Industries Association) weight of the lead. The dip in the characteristic at the higher frequencies is caused by the stiffness of the lead.

For existing enclosures, lead cloth, similar to leaded vinyl plastic may be used in wall paper fashion. Lead sheets 1/16-inch thick can be obtained with wood veneer laminated to one side of the sheet, with a pressure sensitive adhesive on the opposite face.

Fig. 2-52B is a plot of a 2-inch solid sand-plaster wall, compared to an 0.065-inch lead wall. In Fig. 2-52C a lead wall 0.125-inch thick is compared to a plaster and stud wall, and in Fig. 2-52D a staggered stud wall is compared to a 0.23-inch lead wall. Lead is also useful in isolating floors and stages from building structures, and increasing the transmission loss through monitor rooms and vocal room walls. It is also excellent for reducing the noise of rotating machinery. A suggested barrier for high-intensity low-frequency sounds is shown in Fig. 2-52E. The lamination consists of vinyl film, 1/8-inch lead sheet, 3 inches of low-density Fiberglas, 1/8-inch lead, and a final cover of vinyl plastic.

2.53 ***What differences may be expected between the theoretical and practical designs for an enclosure of given dimensions?***—The absorptivity of acoustic materials varies with the angle of incidence. Using the absorption coefficients supplied by the manufacturer of the particular material at hand, results in a mean value. Actual measurements made in an enclosure may not coincide with the theoretical reverberation time. This may be due to the lack of sound diffusion causing constant reflection angles.

2.54 ***What effect does a highly polished surface have on sound waves?***—It reflects the sound waves in a manner

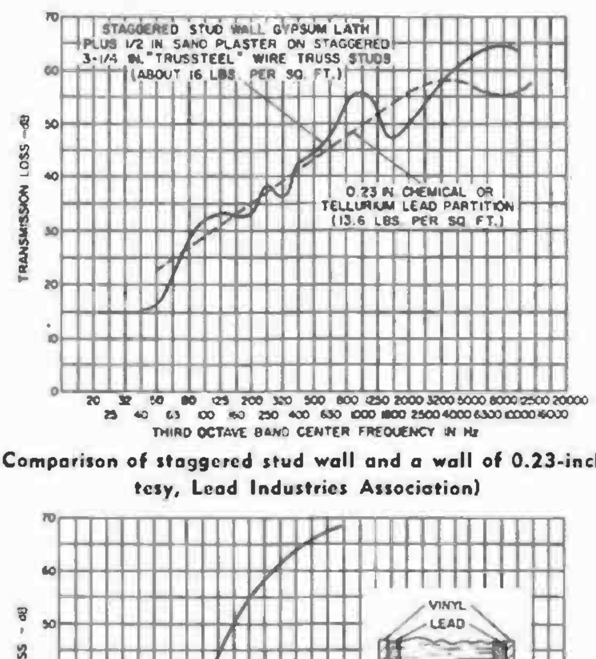

Fig. 2-52D. Comparison of staggered stud wall and a wall of 0.23-inch lead. (Courtesy, Lead Industries Association)

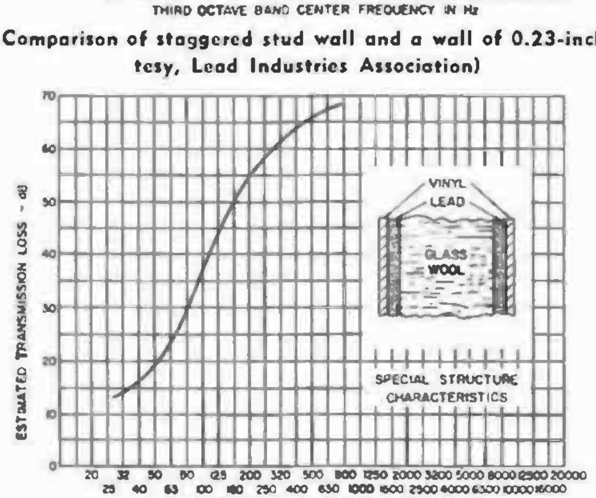

Fig. 2-52E. Suggested construction of a barrier for high-intensity, low-frequency isolation. (Courtesy, Lead Industries Association)

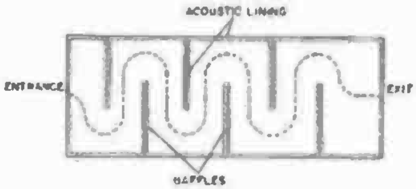

Fig. 2-55. An air duct with sound baffles.

similar to a mirror. (See Fig. 2-21A.)

2.55 *How can sound be prevented from entering an enclosure through an air duct?* —The duct is lined with an acoustic sound-absorbing material such as rockwool or *Ultracoustic*. Baffles covered with the same material are placed at intervals in the duct, as shown in Fig. 2-55.

2.56 *Define acoustic transmittivity.* —The acoustic transmittivity of an interface or spectrum is the ratio of the rate of flow of transmitted sound energy to the rate of incident flow. All directions of incident flow are assumed to be equally probable.

2.57 *Show the construction of a monitor room glass window.* —A typical installation is shown in Fig. 2-57. The glass panels are set in either cork or rubber seals at both the top and bottom, to prevent vibration of the glass from building noises. The glass must be of plate, at least ⅜-inch thick, and preferably ½-inch thick. The panel on the studio side is tilted at the top about 5 degrees to reduce light reflections, and

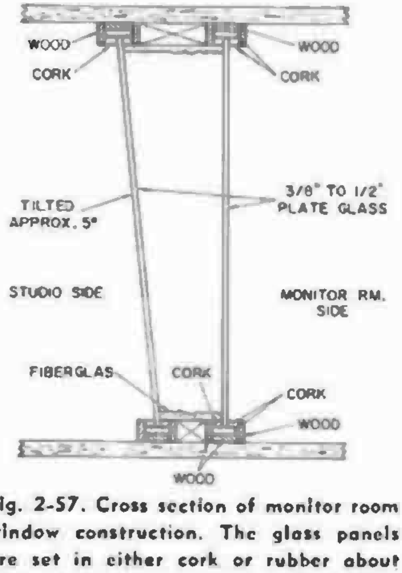

Fig. 2-57. Cross section of monitor room window construction. The glass panels are set in either cork or rubber about 6 to 8 inches apart at the bottom.

to prevent it from acting as a direct reflector in the studio. By using the construction shown, it is possible to achieve 50-dB reduction in sound transmission at 500 Hz.

2.58 *Show the reverberation characteristics for a music-scoring stage and a production-shooting stage.* —For music-recording stages, the desired reverberation will be somewhere between 0.8 and 1.8 seconds, depending on the cubic volume of the stage. After the acoustic treatment has been completed, reverberation measurements are made and additional treatment in the form of live and dead panels are added to bring the measured characteristic nearer the desired characteristic. If polycylindrical diffusers were not included in the original design, they may have to be added to secure the proper diffusion, and to control the low-frequency end of the frequency spectrum. Generally the final reverberation characteristic will be somewhat of a compromise between the desired and a practical characteristic. In Fig. 2-58A, the reverberation characteristic is shown for a music-scoring stage of 210,000 cubic feet, at Republic Corporation, North Hollywood, California. The actual measured reverberation time is plotted versus the ideal and desired characteristic, for comparison. Interior views of this stage are shown in Figs. 2-66A and 2-66B.

Motion picture production-shooting stages are treated in a different manner. These stages are designed to have a high rate of attenuation to outside noises and to prevent reflection from the walls, also to deaden sounds generated within the stage itself. As a rule, a production stage has a reverberation period of 0.6 second at 500 Hz. In the final analysis, the motion picture set will control the acoustics of the sound pickup by its construction materials. The reverberation characteristics for Stage 19 at Republic Corporation are shown in Fig. 2-58B. The volume of this stage is 600,000 cubic feet.

In both stages, the reverberation characteristics will be altered when they are in use. For the music stage, the reverberation characteristic at the higher frequencies will be decreased by the presence of the musicians and their equipment, while the production stage varies from time to time as sets are moved or changed and the number of

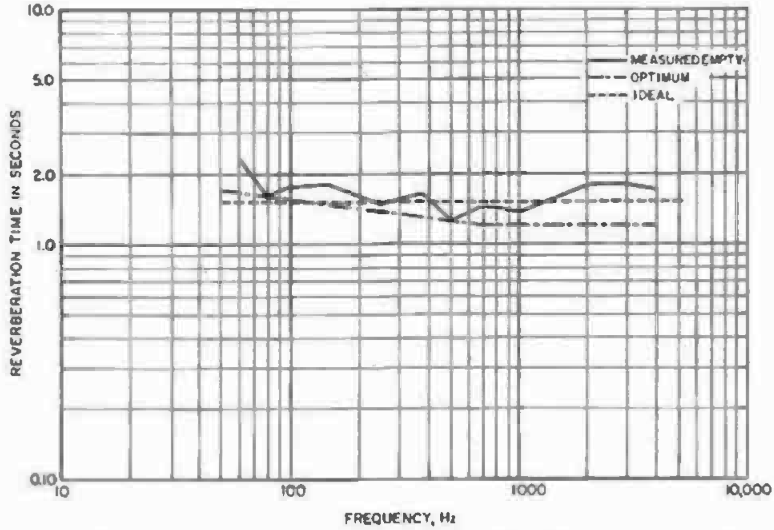

Fig. 2-58A. Reverberation characteristics of Republic Corporation music-scoring stage shown in Fig. 2-66A and 2-66B.

personnel and amount of equipment are varied.

2.59 ***How are the acoustic characteristics of a sound stage measured?***— The acoustic characteristics of a sound stage may be measured in several different manners. The simplest method is shown in Fig. 2-59A. Here a warble tone oscillator is applied to a power amplifier and a loudspeaker system, and frequencies of interest projected into the enclosure. The output from the loudspeaker is picked up, using a sound level meter, and the measurements plotted frequency by frequency. This method is not too accurate, and will give only a general idea of the enclosure characteristics, indicating the peaks and valleys caused by the generation of standing waveforms in the enclosure. If the frequency response of the loudspeaker system is known, this may be taken into consideration when plotting the final results. A measurement made in this manner does not give any indication of the reverberation characteristic, only the frequency characteristic.

A second method (Fig. 2-59B) employs a random-noise generator and a power amplifier to drive a loudspeaker. The projected white noise (if a pink-noise filter is used, see Questions 1.41 and 1.42) is picked up with an octave band analyzer, and applied to a high speed graphic level recorder. In this latter system, the white noise is broken down into frequency bands of ⅓ octave or less, and the sound levels measured. The final characteristic is then plotted in third-octave bands to show the characteristic of the enclosure.

To measure the reverberation period, tones are radiated from a loudspeaker, either by the use of a tone-burst generator, or by hand keying the frequencies of interest. When the tone-burst method is used, an audio oscillator is connected to the tone-burst generator input, as shown in Fig. 2-59C. The generator is set for the desired time interval of tone. The signal is picked up by

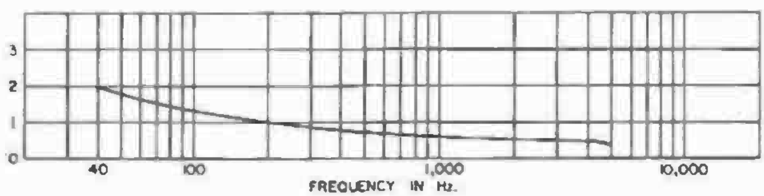

Fig. 2-58B. Reverberation characteristics of Republic Corporation sound Stage-19. Volume 600,000 cubic feet.

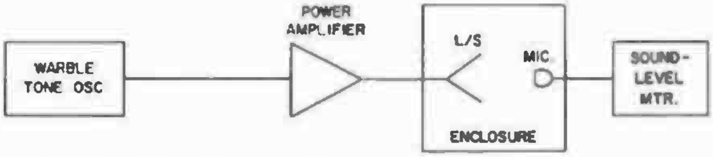

Fig. 2-59A. Setup for measuring acoustic characteristics using a warble oscillator and sound-level meter.

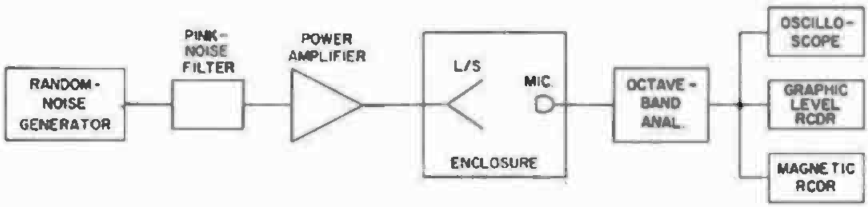

Fig. 2-59B. Setup for measuring the characteristics using a random-noise generator, pink-noise filter, and an octave-band analyzer.

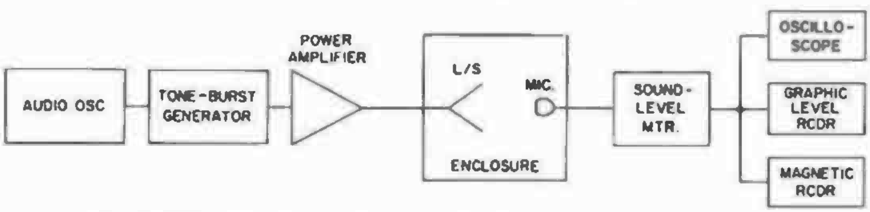

Fig. 2-59C. Setup for making reverberation measurements using a tone-burst generator.

the sound level meter, and recorded on the graphic level recorder, or observed on an oscilloscope.

Another method often used is to fire a pistol in the enclosure, and record the shot on a magnetic recorder. The tape is then played backward through a bandpass filter and an integrating network and the reverberation time computed. Magnetic-tape recorders may be used for recording acoustic tests, provided they have the required dynamic range and are not driven into overload. The reverberation characteristic shown in Fig. 2-58B was made by means of filtered gun shots and the use of a sinusoidal tone. The difference between the two methods was on the order of 10 percent.

It is desirable, although not absolutely necessary, that the frequency characteristics of the loudspeaker system be measured before making acoustical tests, and also that the amplifier used for driving the loudspeaker has a flat response below and beyond the frequencies used in the measurements. It is also desirable that the output from the signal generator supplies a constant signal level to the amplifier input.

The loudspeaker characteristic can be measured by using a constant input to the amplifier, while observing the frequency response on the sound level meter. Knowing the characteristic of the loudspeaker will many times account for peaks and valleys in the final measurement. The sound level meter microphone should be placed close enough to the loudspeaker to eliminate the effect of reflections from surrounding objects, when measuring the characteristics of the loudspeaker. A better method would be to measure its response in the open air.

2.60 ***What is a scoring stage?***—A music-recording stage. This term originated in the motion picture industry.

2.61 ***What are the essential differences between a stage designed for recording music and one designed for recording dialogue?***—A stage designed for recording music is much brighter and larger, and has a longer reverberation time than one designed principally for dialogue recordings. Dialogue stages are rather dull and have a short reverberation period.

2.62 ***How is separation obtained between a vocalist and an orchestra?***—By the use of a separate microphone placed behind an acoustical flat. A glass window in the flat permits the vocalist to observe the orchestra conductor for cueing. A rug is placed under the microphone to prevent reflections from the floor.

2.63 ***How can an artist be cued when singing with an orchestra?***—A single headphone is sometimes used; however, this is not always satisfactory because some artists can not sing and hold the headphone. Also, it bothers them to have one ear covered. A method which has proved to be quite successful is shown in Fig. 2-63. Two small loudspeakers without baffles are mounted on a stand supported at ear level, about six inches from the artist. The sound level from the speakers is held quite low and the microphone is placed at a position for minimum pickup. The loudspeakers must be electrically in phase.

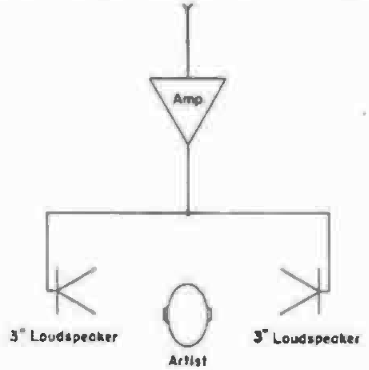

Fig. 2-63. Loudspeakers for cuing an artist singing with an orchestra.

2.64 ***What is a choir room?***—A room adjacent to a scoring stage, with a glass panel in the wall separating the two. The choir to be recorded is placed in the choir room for better separation and control. A loudspeaker operating in the choir room is used for cuing purposes. Headphones, fed from the monitor system, are provided for the conductor so that he may hear the overall mix.

2.65 ***How may a choir and orchestra be recorded if a choir room is not available?***—The choir is separated from the orchestra by acoustical flats. Two microphones, separated by about 10 feet, and placed forward of the group, are used for the choir pickup. The microphones must be electrically in phase.

2.66 ***Show the interior construction of a motion picture scoring stage.***—In Figs. 2-66A and B are shown two views

Fig. 2-66A. View of Republic Corporation scoring stage looking toward the screen. Polycylindrical diffusers may be seen along the sidewalls and on the ceiling.

Fig. 2-66B. Rear view of Republic Corporation scoring stage showing polycylindrical diffusers on rear wall and ceiling.

of the music scoring stage at Republic Corporation, North Hollywood, California. It will be noted the stage has been diffused by the use of polycylindrical diffusers on the walls and ceiling, and that the stage is tapered toward the screen. The walls are constructed in a manner similar to that shown in Fig. 2-50A. The outside walls are coated with a layer of stucco plaster supported on wire mesh. Under the wire is a layer of building paper. All this is supported on four-inch wooden studs. The space between the studs is filled with rock wool. Next, is an air-filled space of six inches and then four-inch studding filled with rock wool. Again, building paper and, finally, the interior finish, which consists of acoustic tile and other materials.

2.67 ***What are the recommended reverberation characteristics for recording stages?***—Scoring stages: fairly live and well diffused. Scoring-stage monitor rooms: slightly deader than a theater. Recording stages: similar to a medium-size theater.

2.68 ***If motion picture projection is used in conjunction with a scoring or dubbing stage, how are the walls of the projection room treated?***—To provide fire protection and, at the same time, a high degree of acoustic isolation, the walls of the projection room are constructed using eight inches of concrete. The interior of the booth is treated as prescribed by fire regulations. Two pieces of optical glass are used in each porthole to isolate the sound of the projectors from the stage. The wall facing the interior of the stage is treated, as are the other walls. As a rule, polycylindrical diffusers are placed horizontally across the face of the projection room wall in the stage just above the portholes to break up the flatness of the surface, as shown in Fig. 2-66B.

2.69 ***Describe how the entrance doors to a production stage are constructed.***—The exterior doors are constructed similar to a walk-in refrigerator door, with interlocking edges similar

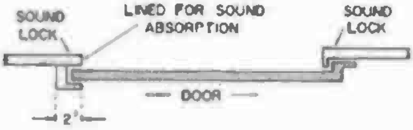

Fig. 2-69. Plan view of a typical sound-stage door installation. The sound lock on the left side starts at the floor and continues to the top and across to the right side. Rollers at the bottom support the weight of the door when it is opened.

Fig. 2-72A. Teleprompter unit with script. The "hot line" appears to the right of the plastic pointer.

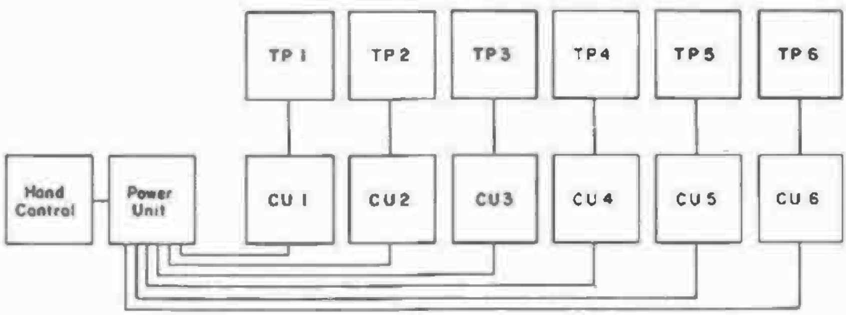

Fig. 2-72B. The block diagram for a six-unit Teleprompter system. The power unit at the left will supply sufficient power for one to six repeater units.

Fig. 2-72C. Two Teleprompter control units and one power unit.

to a bank-vault door. The outer door leads into a small vestibule, with an inner door placed far enough away from the exterior door to prevent opening both doors at the same time. As a rule, signal lights are placed both outside and inside the stage to indicate when a door is being opened. The interior door may be a regular door that has been treated acoustically on its inner surface. The interior of the vestibule is heavily treated acoustically.

Large doors used for bringing in sets and equipment generally weigh several thousand pounds. The most economical design is a concrete slab, reinforced with steel rod, which may be cast in one piece. Hydraulic seals are provided at the sides and bottoms, and interlocking sound barriers for the top and sides (Fig. 2-69).

2.70 ***How is the sound of footfalls eliminated when making dolly shots during production?***—By means of heavy woolen socks worn over the shoes of the operating crew.

2.71 ***When using a loudspeaker for acoustic measurements on a stage, how is the formation of standing wave trains prevented?***—By use of a warble-tone oscillator or film reproduced by the projection system. As a rule, frequencies above 1000 Hz do not require warbling, unless the enclosure is quite small. Standing wave trains are discussed in Question 2.41; stage measurements in Question 2.59; and warble oscillators in Question 22.52.

2.72 ***What is a teleprompter?***—It is a device beyond the sight of the audience, for presenting written material such as a script to an actor or speaker for the purposes of prompting. It is used in production of both television shows and motion pictures.

A teleprompter system consists of one to six variable-speed prompter units such as that shown in Figure 2-72A, which carry the script on a paper roll; a power supply unit; and a hand control for varying the speed of the

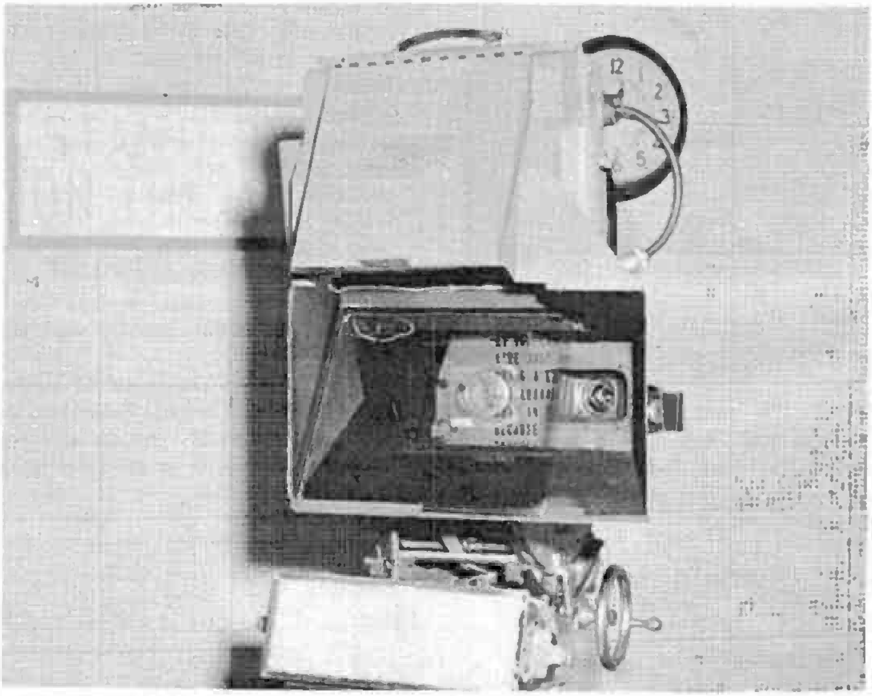

Fig. 2-72D. A Teleprompter unit mounted on a standard motion-picture camera. The image of the script may be seen on a glass plate in front of the Tellens unit in front of the camera lens.

prompter transport systems and reversing their direction for rewinding. The speed of the teleprompter units may be controlled from a few inches per minute to several feet per second, permitting the operator of the system to follow the action on a set, or to adjust the speed of the prompter transport system to accommodate the talking speed of the speaker. The speed of the prompter units is controlled in such a manner that the "hot line" is always in line with a large plastic pointer at the left of the script.

A block diagram for a typical system employing six prompter units is shown in Fig. 2-72B, while Fig. 2-72C is a view of two control units, a power supply, and the variable-speed hand control.

The script is typed on a special typewriter called a Videoprinter which has type 52 times larger than that normally used on a standard typewriter. This size type permits the script to be read up to distances of 25 feet, and more, which is adequate for most purposes.

For scenes which require the narrator to peer directly into the camera lens, a special type mounting for the prompter unit, called a Tellens, has been developed which can be mounted on a standard motion picture camera as shown in Fig. 2-72D. The image of the script is projected downward from the prompter unit onto a piece of clear glass set at an angle of 45 degrees in front of the camera lens. As the person being photographed looks into the camera, the image of the script is seen on the glass plate in front of the camera lens.

The reversing switch on the hand control permits the operator to quickly rewind the script to any particular line in a matter of seconds. Script changes are retyped and fastened to the script by plastic tape. Each prompter unit is well lighted and designed to be mounted using a stand or special support on the camera dolly.

2.73 ***What is a diffuser?***—A wooden panel with an uneven surface as shown in Fig. 2-75. These devices are constructed of ¼-inch plywood and placed in different positions near the source of sound pickup to add reverberation and a degree of liveness to a pickup in a large stage.

2.74 ***Describe the construction of acoustic flats.***—Acoustic flats are used on music scoring and looping stages for the purpose of separating a vocalist, a group of singers, or an actor in the case of looping. The flats are constructed of 2-inch × 4-ft × 84-inch wooden frames, diagonally braced and filled with rock wool or Fiberglas. The exterior surface treatment is varied; that is, one flat has a plywood and acoustic tile surface, another a soft surface of cheesecloth with Fiberglas backing and acoustic tile. Pull-pin hinges are provided for locking the flats together, similar to a Japanese screen, to form a semicircle. Thus, the interior acoustics of the enclosure may be varied to acquire the necessary acoustical environment. A window is provided in one flat for the vocalist to watch the conductor, or for an actor to view the screen when looping. A carpet is placed on the floor to prevent reflections and foot noises. Constructional views are given in Fig. 2-74A, for a typical group of such flats.

For a large group of voices to be separated from the orchestra, several panels are used, similar to those discussed, but constructed somewhat differently. Each panel is 4 ft wide × 12 ft in height, and contains a glass panel, with plywood the first 4 ft up from the floor, as shown in Fig. 2-74B. The frames are constructed of 2 × 4's with pull-pin hinges. The interior surfaces of the panels are crossed-braced and filled with rock wool or similar material to prevent resonance effects and to increase the isolation. In setting up the panels for use, care must be taken that no two surfaces are parallel, as shown in Fig. 2-74C. Carpet is placed on the floor to prevent reflections and eliminate foot noise. Only about 10 to 15 dB of isolation may be expected from an enclosure of this nature.

2.75 ***What is a splay?***—A curved surfaced diffuser as shown in Fig. 2-75.

2.76 ***Describe the construction of a polycylindrical diffuser.***—Polycylindrical diffusers consist of a plywood panel bent in the form of a convex surface. Such devices are used in rerecording stages, music-scoring stages, broadcast studios, and in some instances theaters. Their purpose is to provide a maximum

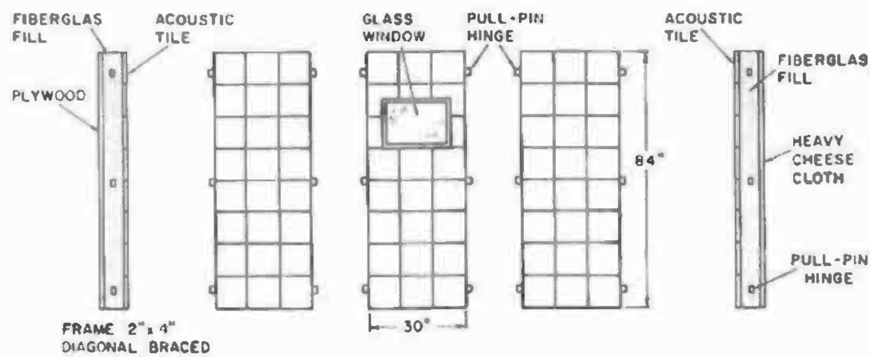

Fig. 2-74A. Acoustic flats used for separation of a vocalist, or for looping.

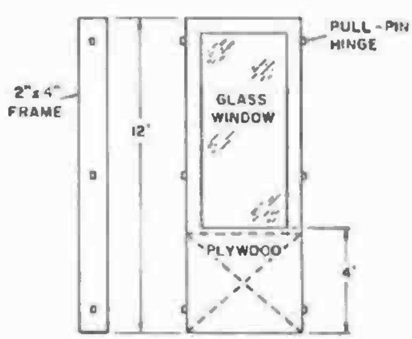

Fig. 2-74B. Glass-paneled flats for enclosing a choral group.

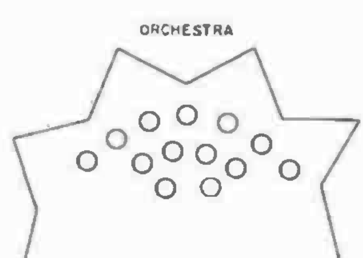

Fig. 2-74C. Ten to twelve such panels are placed around the group, as shown in the floor plan.

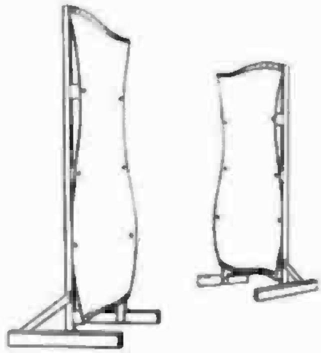

Fig. 2-75. Curved surface diffusers or splays.

of diffusion and aid in the control of the low-frequency reverberation characteristics of the enclosure. The diffuser, because of its curved surface, reflects the high frequencies at many different angles. At the low frequencies, the panel acts as a diaphragm and breathes, dissipating the low-frequency energy in the form of heat. The construction of such diffusers is given in Fig. 2-76A, with the details of their measurements in Figs. 2-76B and C.

When finished, the surfaces of the panel may be painted with a bright glossy paint and then laquered. A dull surface paint must not be used, as a highly reflective surface is necessary. The internal space is filled with loose rock wool, or a backing of 2-inch Fiberglas, placed on the supporting wall. The curved panel must be securely braced; the braces are randomly spaced to prevent selective resonance. The curvatures of the panels between the various diffusers are varied also, and their axes disposed to be mutually perpendicular in the three orthogonal planes. Diffusers may be used in both the vertical and horizontal planes as well as across ceilings and back walls. For a studio 40 feet in length, two such diffusers will be required for each wall, running vertically from the floor to the ceiling, placed directly across from each other, as shown in Figs. 2-76D and E. In small rooms, the ceiling diffusers may be omitted if the ceiling surface is at different angles or broken up. For large stages, the general plan of that shown in Figs. 2-66A and B may be followed.

Concave surfaces must be avoided, as they are points of concentration, and focus the sound rather than diffuse it. It should be mentioned that sheet *Masonite* cannot be used successfully in the construction of polycylindrical diffusers since the material is not stiff enough. To have the proper stiffness, the plywood must be at least $\frac{1}{4}$ inch in thickness.

2.77 ***Show the polar frequency response of a polycylindrical diffuser compared to a flat-surfaced splay.***—A polar plot for a flat and curved baffle is shown in Fig. 2-77. It will be noted that the angular reflection of the curved surface, with respect to frequency, covers over 100 degrees, while the flat panel covers only 40 degrees. This illustrates very clearly the advantages of a curved surface over that of a flat surface.

2.78 ***What is a baffle plate?***—A partial plate placed in an air duct to prevent exterior noise from entering an enclosure. Baffle plates also reduce the rushing sound of the air as it is forced through the duct. The construction of such a duct is shown in Fig. 2-55.

2.79 ***What is an echo?***—The repetition of a sound caused by reflection from a surface. To be an echo, the reflected sound must be $\frac{1}{20}$ of a second or longer behind the original sound.

2.80 ***What is an echo chamber?***—A highly reverberant room which is loug and narrow and has hard walls. A loudspeaker is placed at one end of the room and a microphone at the other. The sound to be reverberated is sent into the loudspeaker, picked up by the microphone, amplified, equalized, and mixed with the original program material. Because of the hard walls, multiple reflection echoes are produced. Typical echo chamber designs are shown in Fig. 2-80. At (a), the microphone is separated from the loudspeaker by a partition running almost the full length of the room, to secure a greater delay. In (b) is shown a plain room with the loudspeaker at the farthest end. For this type of operation, the microphone and speaker are moved to secure the desired results. At (c), a movable partition has been installed at the center of the room and is remotely controlled to alter the length of the reverberation period. Sketch (d) is a similarly constructed room, except the end and side walls are set at an angle.

Echo effects may also be generated by the use of 1-inch pipes, or larger, ranging from 25 to 100 feet in length. In the early days of radio, such systems were constructed by coiling the pipes several feet in diameter, and placing microphones at intervals of 25 feet to achieve the delay times. The sound was introduced into the pipe at one end, using a horn-type driver unit. For a distance of 50 feet the delay is 1/20 second, and 1/10 second for 100 feet. The disadvantage of this method of creating reverberation is the tremendous loss of high frequencies and the amount of equalization and amplification required

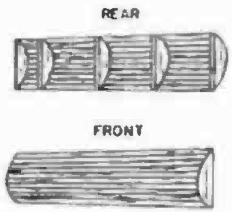

Fig. 2-76A. Front and rear views of a polycylindrical diffuser, showing the interior construction.

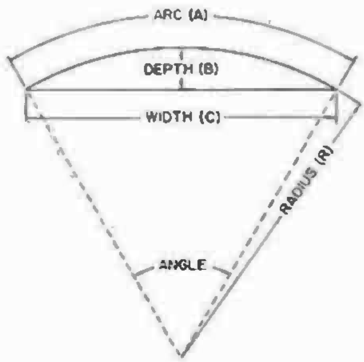

Fig. 2-76B. Basic plan for a polycylindrical diffuser.

	60°			90°			120°			180°
Arc (a)	*Width (c)*	*Depth (b)*	*Radius (r)*	*Width (c)*	*Depth (b)*	*Radius (r)*	*Width (c)*	*Depth (b)*	*Radius (r)*	*Radius (r)*
16"	15¼	2	15¼	14⅜	3	10⅜	13⅛	3⅞	7⅝	5⅛
32"	30⅝	4⅛	30⅝	28¾	6	20¾	26¼	7⅝	15¼	10⅛
48"	45⅞	6⅛	45⅞	43¼	9	31¼	39½	11½	22⅞	15¼
64"	61⅛	8¼	61⅛	57⅝	12	41⅝	52⅝	15¼	30½	20⅜
80"	76⅜	10¼	76⅜	72	15	52	65¾	19⅛	38⅛	25½
96"	91⅝	12¼	91⅝	86⅜	18	62⅜	78⅞	23	45¾	30½
112"	107	14⅜	107	100¾	21	72¾	92⅛	26¾	53⅜	35⅝
128"	122¼	16⅜	122¼	115¼	23⅞	83¼	105¼	30⅝	61	40¾
144"	137½	18⅜	137½	129⅝	26⅞	93⅝	118⅜	34⅜	68⅝	45¾

Fig. 2-76C. Dimensions for polycylindrical diffusers.

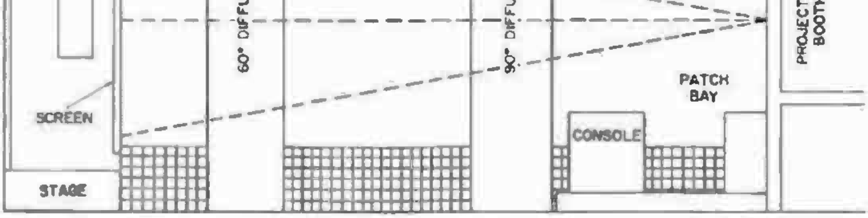

Fig. 2-76D. Cross-sectional view of a typical dubbing stage using polycylindrical diffusers and a broken-up ceiling. A fifth diffuser is placed horizontally across the rear wall about half way between the projection portholes and the ceiling. The wall directly behind the console is broken up in a vertical direction. The console sets on a platform 12-inches high.

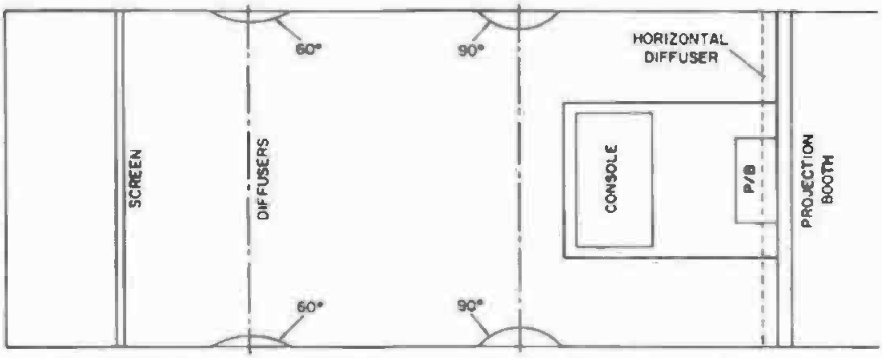

Fig. 2-76E. Floor plan of dubbing stage showing the positions of the polycylindrical diffusers.

to return the signal to a point where good intelligibility is attained. Because of the short duration of the reverberation, the signal picked up by the microphones must be fed back to the driving source, and consequently multiple delays are created. For a pipe 50 feet in length, the loss at 5000 Hz is approximately 23 dB, and for 100 feet it is approximately 45 dB. Because of these problems, this system is no longer used. Artificial reverberation also may be generated electronically or by electromechanical means, as discussed in Questions 2.128 to 2.130. A typical reverberation system using a pipe is shown in Fig. 2-80B.

2.81 ***How is an echo chamber constructed?***—Typical interior dimensions are: Length 18 feet, width 15 feet, ceiling height 12 feet, consisting of about 3200 cubic feet and a reverberation

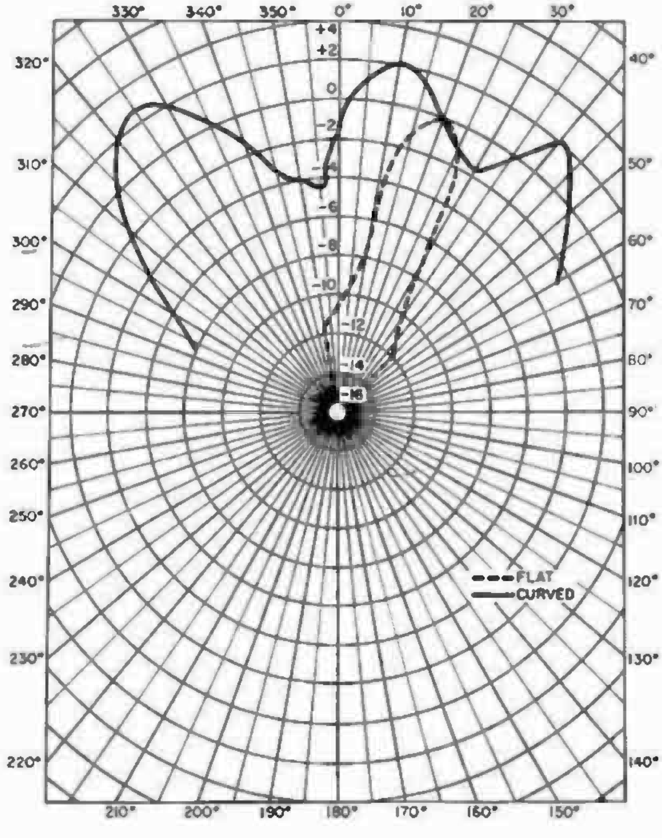

Fig. 2-77. Polar frequency response of a flat baffle as compared to a curved baffle.

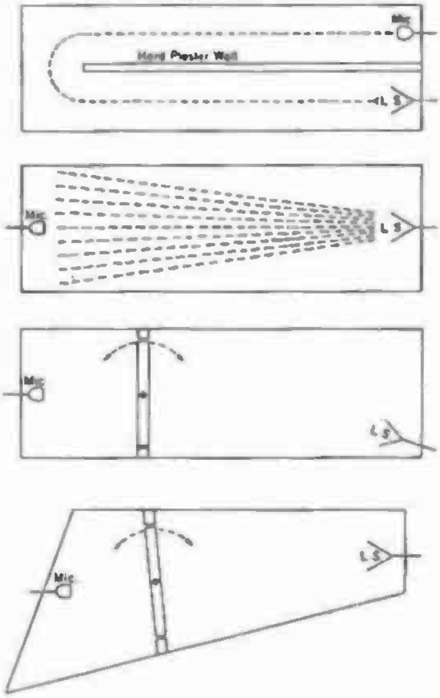

Fig. 2-80A. Echo chamber designs.

time of approximately 3.8 seconds. The floor should be smooth cement and the walls covered with hard, smooth plaster. The ceiling must be nonparallel to the floor. The speaker may be moved and microphones with different pickup characteristics used to secure the desired effect. However, a room with the microphone and speaker in fixed positions and a movable partition are the most convenient to operate. Echo chambers may also be constructed as shown at (d) in Fig. 2-80A. The ideal echo chamber is one constructed using six-inch concrete walls. To be effective, an echo chamber must have a minimum of 2500 cubic feet.

2.82 ***How is an echo chamber connected into a recording channel?***—The output of the source of sound to be reverberated is bridged with an amplifier and applied to the power amplifier driving the loudspeaker in the reverberation chamber, as shown in Fig. 2-82. The reverberated sound is picked up by a microphone, amplified, equalized, and applied to the input of the recording channel through a mixer pot.

A different method of combining the signal in the recording channel is shown in Fig. 9-49. When combining the reverberated sound with the original material, the reverberated sound is mixed at a level approximately 20 dB below the original. This gives the best results, with the highest intelligibility for speech.

A separate echo chamber is required for each microphone to be reverberated.

2.83 ***What is an anechoic chamber?*** —An enclosure in which the reflected sound is negligible. Such rooms are used for measuring the characteristics of microphones, loudspeakers, and other acoustic transducers, and to provide environmental conditions similar to the outdoors. Fig. 2-83A shows an anechoic test chamber at the Bell Telephone Laboratories, Murray Hill, New Jersey. To eliminate surfaces that would reflect

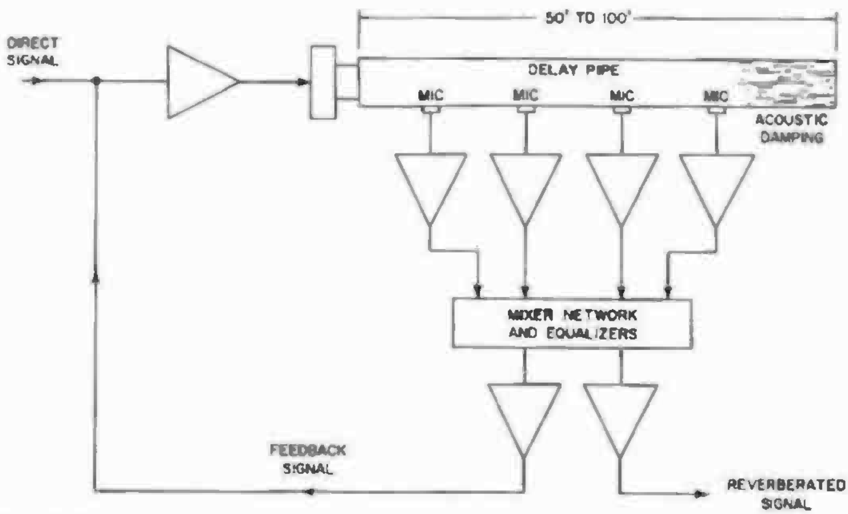

Fig. 2-80B. System for generating artificial reverberation using pipe delay lines.

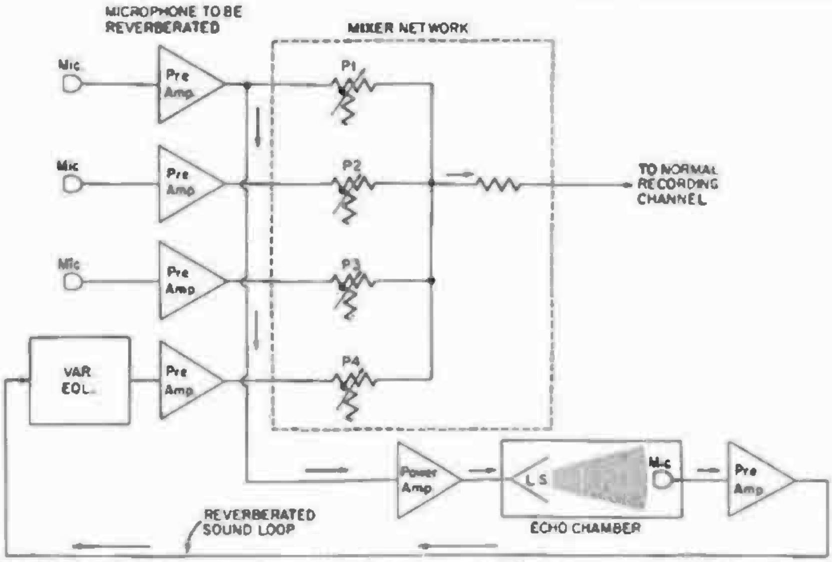

Fig. 2-82. Block diagram for connecting an echo chamber to a recording channel.

sound, the walls, ceiling and subfloor are lined with sawtooth wedges of Fiberglas to a depth of five feet. The flooring consists of high strength steel cables $^{9}/_{100}$-inch thick, strung under tension in two-inch mesh. The chamber dimensions are 35 × 28 × 28 feet. The volume is 27,440 cubic feet. With the interior treatment as shown, the absorption at the walls is 99.98 percent of all incident sound energy in the audible range.

In Fig. 2-83B, is shown a corner of the anechoic chamber of the Electro-Voice plant in Buchanan Mich., used for engineering development and quality control. The main chamber is 35 × 26 × 26 feet and has a volume of 23,660 cubic feet. The walls are treated in the same manner as described for the chamber in Fig. 2-83A. In addition to the main chamber, there are two smaller adjoining chambers; one houses a standard loudspeaker calibrated from 20 to 20,000 Hz, traceable to the National Bureau of Standards; the second is a smaller chamber that will accommodate loudspeakers up to and including a diameter of 30 inches. A standard exponential horn is also available for testing high-frequency driver units.

The ventilation of the main chamber is through three wedges located in widely spaced positions to prevent reflections. Twenty-two circuits are available from the chamber for connections to signal sources and test equipment. Exterior to the chambers are automatic curve tracing and polar pattern graphic recorders. The entire facility is set on its own foundation, which completely isolates it from the main building.

Two chambers, similar in construction are in use at the Altec-Lansing plant in Anaheim, California. One is used for engineering and a second for routine production testing. Fig. 2-83C is a view of the test equipment outside the chamber used for research and development. The equipment includes preamplifiers, power supplies, graphic level recorder, oscillator frequency counter, and other devices. The equipment shown in Fig. 2-83D is used for routine testing of microphones and other devices. The device to be tested is supported by a pulley arrangement and cords for orienting and ease of operating. The test equipment consists of a graphic level recorder, oscillator, preamplifier, vacuum-tube voltmeter, and associated equipment.

2.84 ***Describe the construction of an anechoic chamber.***—Anechechoic chambers are enclosures that are echo-free, within a specified frequency range. To achieve this condition, the sound energy absorption must be between 99 and 100 percent. Or to state it in another way, the sound-pressure reflections must be between 10 percent and zero. The point at which the energy absorption drops below 99 percent or the pressure reflection exceeds 10 percent is known as the low-frequency cutoff frequency. The curves in Fig. 2-84A show the low-fre-

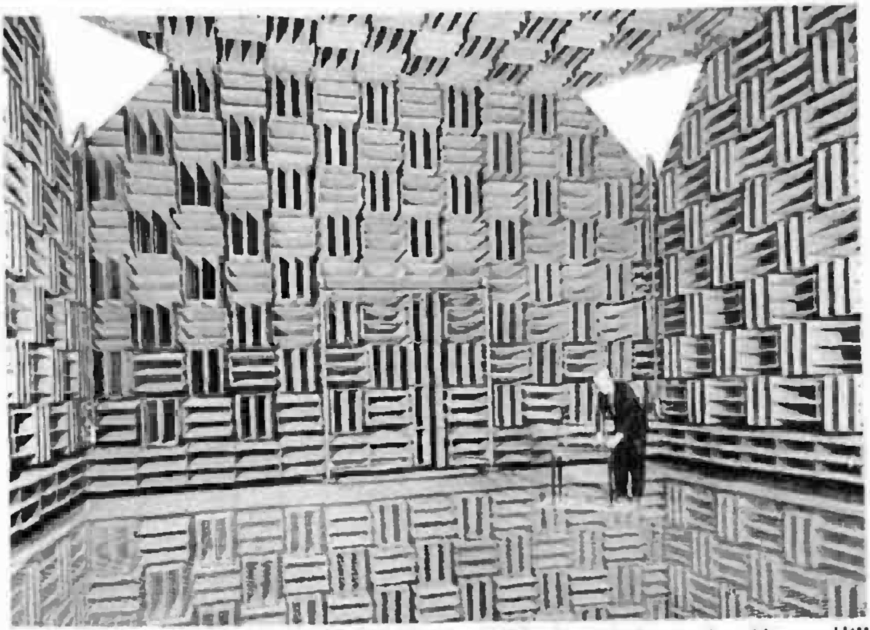

Fig. 2-83A. The anechoic chamber at the Bell Telephone Laboratories, Murray Hill, New Jersey.

Fig. 2-83B. Testing a microphone in the Electro-Voice anechoic chamber.

quency cutoff points for 60, 130, and 250-Hz. The table of cutoff frequeucies recommended for different types of testing is given in Fig. 2-84B.

The actual dimensions for any type anechoic chamber will be determined by the type equipment to be tested. For general acoustical research, the minimum free-field dimensions (the distance from the edge of one wedge to another on the opposite wall) cannot be less than one wavelength of the cutoff frequency, and the largest dimension not less than half a wavelength of the cutoff frequency. The principal factor affecting the free-field dimension is the size of the equipment under test. Measurements at a specified frequency should be taken no closer than a quarter wavelength from the sound radiating surface and no less than one quarter wavelength from the points of the surrounding wedges.

To obtain a satisfactory environment, the anechoic wedges are generally installed in an attenuating structure similar to that of Fig. 2-84C. The outside enclosure may be built on the ground or on an existing floor, or floated on springs, lead, or some type of acoustic

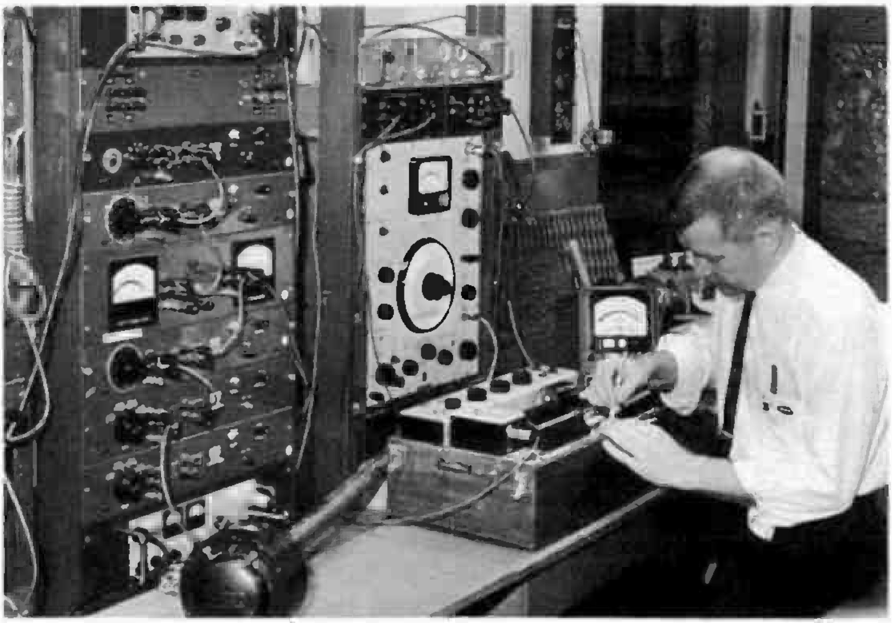

Fig. 2-83C. Test equipment for research and development of loudspeakers and microphones, outside the anechoic chamber in the Altec-Lansing plant.

Fig. 2-83D. Routine production test equipment outside the Altec-Lansing anechoic chamber. The device to be tested is hung from a pulley-and-cord arrangement for ease of operation and orientation.

isolation material. The door in the outer structure must have an attenuation characteristic equal to the outer walls. The inner surface of the chamber door is be treated in the same manner as the inner walls, in order that the anechoic characteristics are not compromised.

To achieve adequate ventilation, twelve complete changes of air per hour will be required, with controlled humidity and temperature. The air from the ventilating system is introduced through a plenum silencer into the chamber through special ventilating wedges (Fig. 2-84D). Velocities up to 250 fpm may be fed to the interior, without introducing noise. A graphical plot of the lower cutoff frequency versus the depth of treatment of the chamber is given in Fig. 2-84E. Typical wedges are shown in Fig. 2-84F.

The question is sometimes raised, what effect if any, does the grating used for the floor have on measurements made in an anechoic chamber? The study of the effects of the floor becomes quite complex. In practice, the floor is a metal grating or nylon cables suspended from the walls. If a grating is used, the openings are generally 1 or 2 inches in depth and 2 inches in length. Thus the slit width is less than half the wave length for frequencies below 3000 Hz. The open area is often in excess of 90 percent of the total area, and consequently when a sound strikes the grating, under normal incidence the reflection is negligible. However, there is a certain amount of reflection. This subject has been treated extensively by Ingard. The reader is referred to the reference at the end of this section.

A portable anechoic chamber, manufactured by the Eckel Corporation, is shown in Fig. 2-84G. The surfaces of the *Fiberglas* wedges are covered with No. 2, 19-gauge hardware cloth. The interior of the chamber shown has a free-field space of approximately 21 cubic feet.

2.85 ***What is the purpose of adding reverberation to program material?***—To enhance the original material by adding brilliance and to create the illusion that the material was originally recorded in a large auditorium. If properly controlled and equalized, the results can be very pleasing. Reverberation is also used for producing sound effects, such as to simulate a large room or cavern. Equalization may be necessary as the reverberation chamber distorts the original frequency characteristic.

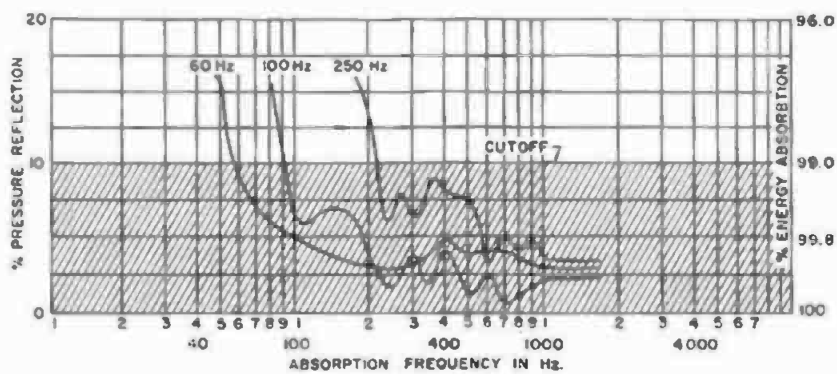

Fig. 2-84A. Absorption characteristics for typical wedges of the Eckel Co.

Type of Testing	Low-Frequency Cutoff Point
Aircraft	75 Hz
Audio	125–400 Hz
Automotive	60 Hz
Electrical Communications	60-200 Hz
Jet Engine	150 Hz
Machinery	75-150 Hz
Musical Instruments	60 Hz
Psycho-Acoustic	150 Hz
Transformer	100 Hz

Fig. 2-84B. Recommended cutoff frequencies for different types of testing.

2.86 ***What is the optimum number of musicians for a given size studio, based on the program material?***—The number of musicians will vary, depending on the program material and the size of the room or studio. Based on experience, the recommended number is given in Fig. 2-86.

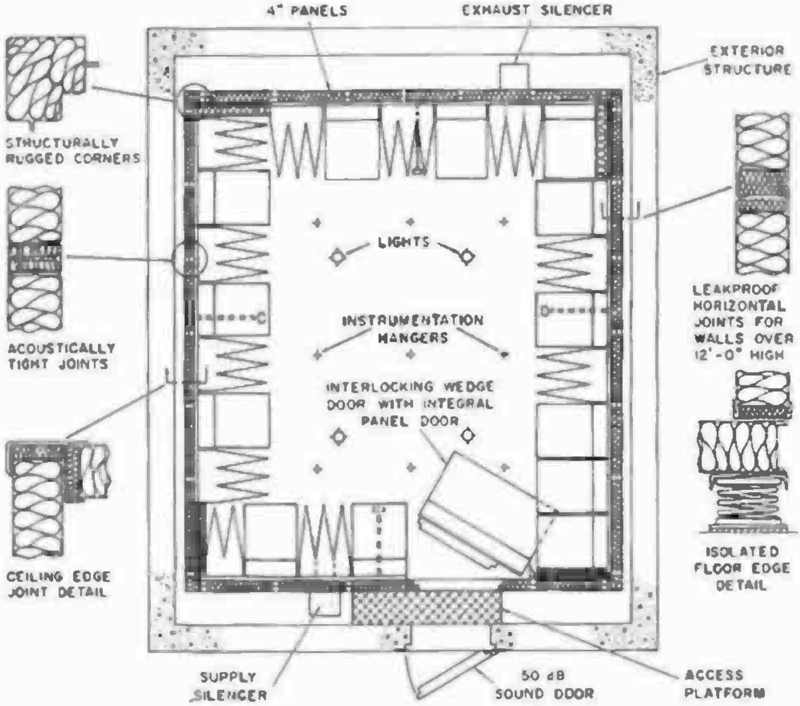

Fig. 2-84C. Typical plan for an anechoic chamber. The interior door is designed so that the wedges mounted on its inner surface interlock with the wedges on the wall to permit the door to open parallel to the wall. (Courtesy, Eckel Co.)

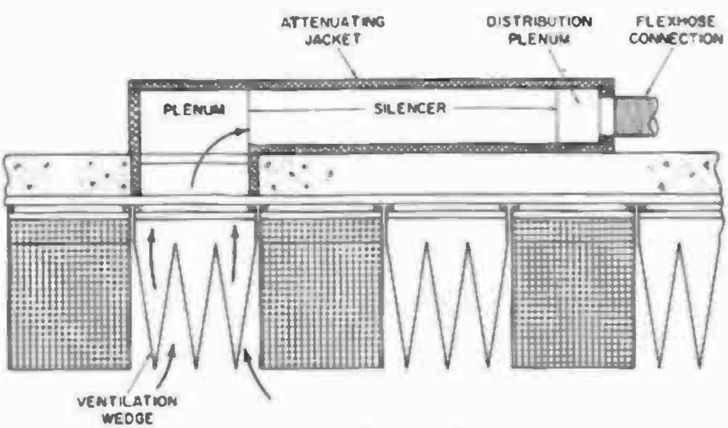

Fig. 2-84D. View showing the installation of the air silencer and plenum.

Volume in Cubic Feet	*Broadcast Studio*	*Auditoriums*	*Scoring Stages*
10,000	12	—	—
20,000	25	6	10
50,000	50	9	22
100,000	130	19	36
200,000	250	31	70
500,000	—	62	140
1,000,000	—	105	240

Fig. 2-86. Recommended number of musicians will vary.

2.87 ***What are acoustic pendants?***— Octahedral-shaped devices suspended from the ceiling of a recording stage or auditorium to break up reflections from a flat or curved ceiling. The pendants are constructed from ⅜-inch plywood and vary in height from 1 to 3 feet. The diameter at the center varies from 6 to 12 inches. These devices should be hung in a scattered manner from the ceiling at varying heights, if the ceiling is flat. If the ceiling is curved, as shown in

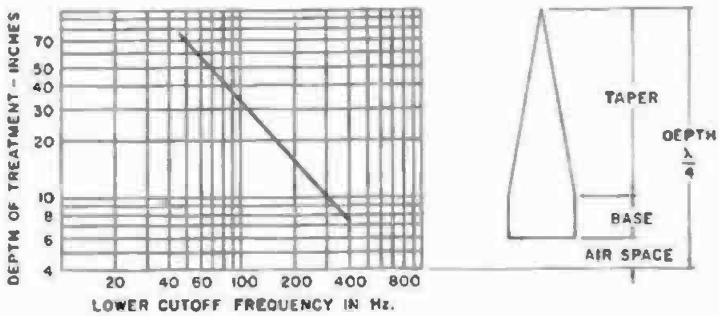

Fig. 2-84E. Lower cutoff frequency versus depth of treatment on the interior walls of an anechoic chamber.

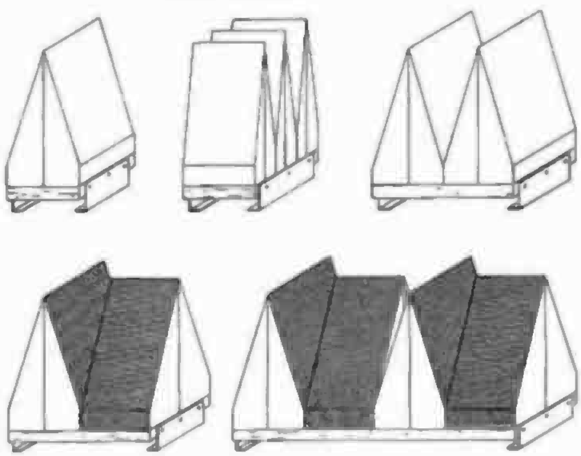

Fig. 2-84F. Typical *Fiberglas* wedges used for the interior treatment of anechoic chambers, with and without hardware screen.

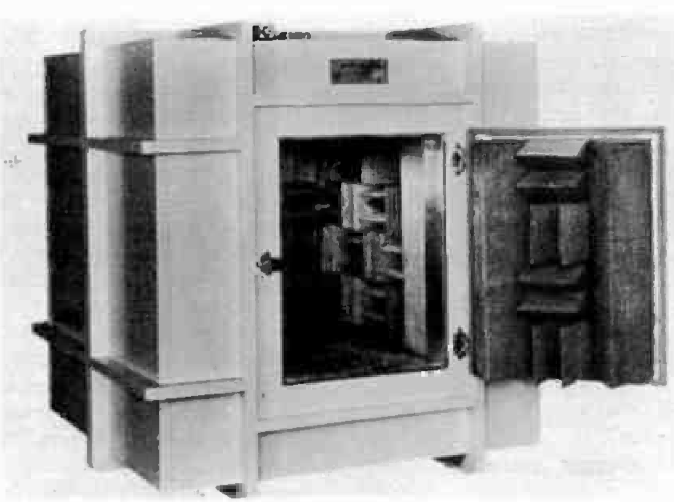

Fig. 2-84G. Portable anechoic chamber manufactured by the Eckel Corporation.

Fig. 2-87, they may be hung at a uniform height. It may be desirable, in some instances, to cover every other one with *Celotex* to provide soft and hard surfaces. The whole purpose is to break up concentrated reflections from the overhead to the side walls and floor. Rooms having strong concen-

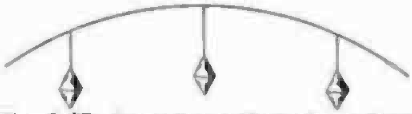

Fig. 2-87. Acoustic pendants hung from a curved ceiling to break up reflections.

trated reflections from the ceiling may be altered to have quite even reflective qualities by the use of these acoustic pendants.

2.88 ***What are the middle-range frequencies?***—Frequencies between 400 and 3500 Hz.

2.89 ***Define the terms sound-pressure level and sound-power level.*** — Sound-pressure level (SPL) is used to express the level of a sound that has been measured using a sound level meter. The reference for these measurements is 0.0002 microbar. Sound level meters do not measure acoustic power. Sound-power level (PWL) is the acoustical power generated by a sound source. Sound-pressure level may be expressed as:

$$SPL = 20 \, Log_{10} \frac{P}{0.0002} \, dB$$

where,

P is the root-mean-square sound pressure in microbars.

Because the range of acoustic powers met with in daily life are on the order of one billion-billion (10^{18}) to one, it is quite convenient to relate these powers using the decibel, which is logarithmic. The reference for these measurements is 10^{-12} watt. The power level may be calculated:

$$PWL = 10 \, Log_{10} \frac{W}{10^{-12}} \, dB$$

where,

W is the acoustic power in watts.

Since 10^{-12} watt corresponds to a level of minus 120 dB, the above equation may be more easily expressed:

$$PWL = 10 \, Log_{10} \, W + 120$$

where,

W is the acoustic power in watts.

Since 10^{-12} is a power ratio corresponding to minus 120 dB, the quantity 10 Log_{10} W which is the value of decibels corresponding to the numerical value in watts, can be readily obtained from the decibel tables that are given in Section 25. As an example: 0.2 watt corresponds to a power level of $-17 + 120 = 103$ dB. Power levels for sounds that are frequently encountered are given in Fig. 1-117.

2.90 ***What is a Rayleigh disc?***—A device invented by Lord Rayleigh in 1882 for measuring sound pressures. The instrument consists of a small light-weight disc, suspended vertically by means of a quartz fiber or annealed bronze wire. A small mirror is cemented to one side of the disc and is supported in a draftless square or tubular chamber constructed of open-mesh silk. When an air stream, whose pressure is to be measured, strikes the surface of the disc, the disc tends to align itself with the direction of fluid and pressure flow. A beam of light focused on the mirror surface permits the angle of deflection to be accurately measured by means of a calibrated lens system also focused on the mirror.

Such a device may be used for calibrating the frequency response of microphones, with a high degree of accuracy, by interposing the disc assembly in an airstream path between a microphone and loudspeaker unit. The loudspeaker unit is excited by an oscillator. Since the speaker characteristics must be isolated from the measurement, the pressure generated by each frequency of interest is adjusted for the same deflection of the disc, so that the microphone under calibration sees the same pressure at all frequencies. The microphone output is then amplified, and the electrical characteristics are measured in the usual manner. The loudspeaker must be of good quality and of low distortion. The microphone preamplifier must have a uniform frequency characteristic and low distortion.

Initially the disc is set to an angle of 45 degrees, which is the angle of the greatest sensitivity, by adjusting the suspension head at the top of the disc enclosure. When the disc is subjected to an air stream, it assumes an angle different than 45 degrees. The angle of deflection is held constant to a reference frequency for each frequency of interest used in the calibration. Barnes and West have shown that if the resonance of the disc falls within the audio spectrum, it can produce errors up to 10 percent in the vicinity of resonance. As an example, a disc with a diameter of 1.6 cm and a thickness of 0.0045 cm has a resonance of 1800 Hz. However, the addition of the mirror to the one side of the disc reduces the resonance somewhat. Since barometric pressure and temperature affect the measurement, they are noted and corrections are made to the final plot.

2.91 ***What is a Helmholtz resonator?***—A cavity resonator which is open to the outside air through a single small hole. Air blowing across the hole will cause a sound to be generated at a frequency dependent on the volume of the cavity. A cross-sectional view of such a resonator is shown in Fig. 2-91.

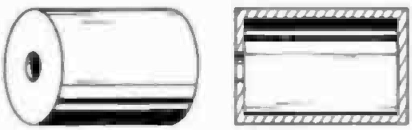

Fig. 2-91. Helmholtz resonator consisting of a cavity with a single hole open to the outside media.

2.92 ***What causes an organ pipe to vibrate?***—Organ pipes are constructed with a reed at the lower end which vibrates when air is forced into the pipe. Sound waves are set up which travel up and down the pipe which is of such length that the wave of air will just have time to travel the full length of the pipe during the interval required by the reed to make one complete cycle. When this occurs, the frequencies of the reed and sound wave synchronize. This is called resonance. At the top end of the pipe is a movable plug for tuning the pipe to the exact frequency. A 16-Hz organ pipe is 32 feet long.

2.93 ***Define the term weighted curve and its usage.***—A weighted curve is a frequency-response curve with a special characteristic. Weighted curves are more commonly associated with sound-level meters used for acoustical measurements. (See Question 22.94.)

Sound-level meters manufactured *prior to 1961* employed frequency characteristics as given in Fig. 2-93A *and are now considered to be obsolete.* In 1961 the industry adopted the American Standard ASA (now USASI) S1.4-1961 shown in Fig. 2-93B. There are also two international standards, ISO R123 and ISO R179. Both the American and International Standards agree within 0.20 dB.

Given in Fig. 2-93C is the USASI Standard S1.4-1961 with the frequency response extended at the lower end to 10 Hz, and at the upper portion to 20,000 Hz. Also included is an additional characteristic curve, N (*it is not a standard*). This latter characteristic is a *proposed* characteristic to be used for the measurement of broadband noise such as encountered in the noise measurement of jet aircraft engines.

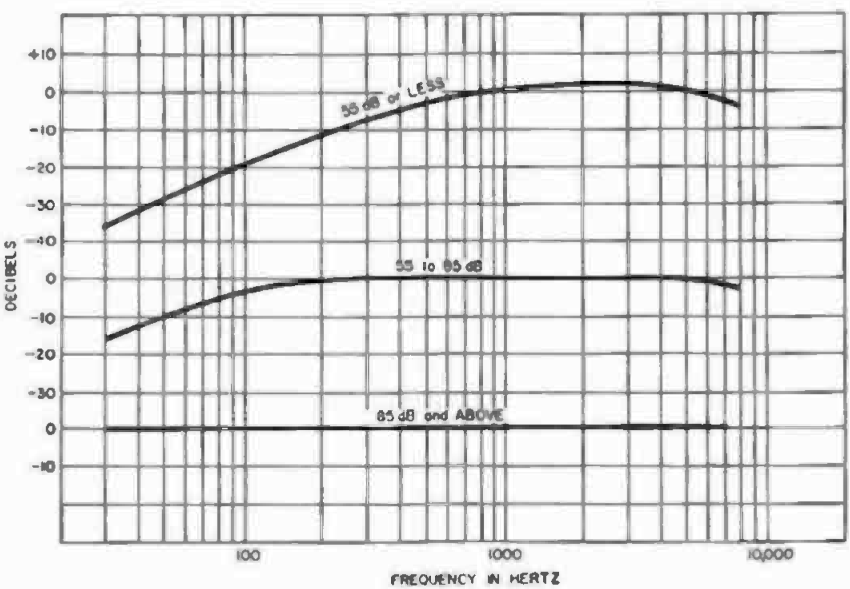

Fig. 2-93A. Weighted curves for sound-level meters manufactured prior to 1961 (ASA 224.3-1944).

The selection of a weighted curve for a particular measurement, is generally predicated on the type noise to be measured. As an example, curve "A" of Fig. 2-93B is often employed when making *speech-interference* measurements, while curve "B" is used for traffic surveys. Whenever a single measurement is made, the reading should be described—curve "A" 45-dB weighted curve, and for "B" 20-dB weighted curve. In the past, for single readings the following rule was used. For levels below 55 dB, characteristic "A" was selected; curve "B" was used for levels of 55 to 85 dB; and above 85 dB, characteristic "C" was used (flat). However, curve "A" is the one most generally employed. Readings taken with curves "A" and "B" are referred to as *sound-level readings*, and those made using curve "C" (flat) are referred to as *sound-pressure levels.*

It is recommended in general that readings be made using all three characteristics to indicate the frequency distribution of the noise. With the same level on all three characteristics, the sound generally predominates *above* 600 Hz. For levels using the "C" characteristics which are higher in level compared to curves "A" and "B," the greater part of the noise is probably *below* 600 Hz (see Question 5.98).

2.94 ***Define the term ambiophony.***—It relates to a method of inducing artificial reverberation into an auditorium, by use of loudspeakers in the ceiling and around the walls. Each group of loudspeakers is delayed in their reproduction the appropriate amount of time, corresponding to the normal delay for its position. Thus, an auditorium may be treated in such a manner that reverberation is induced only by the electronic reverberation system. (See Question 2.37.)

2.95 ***What is prescoring?***—Music recorded prior to the shooting of a motion picture or television scene. The

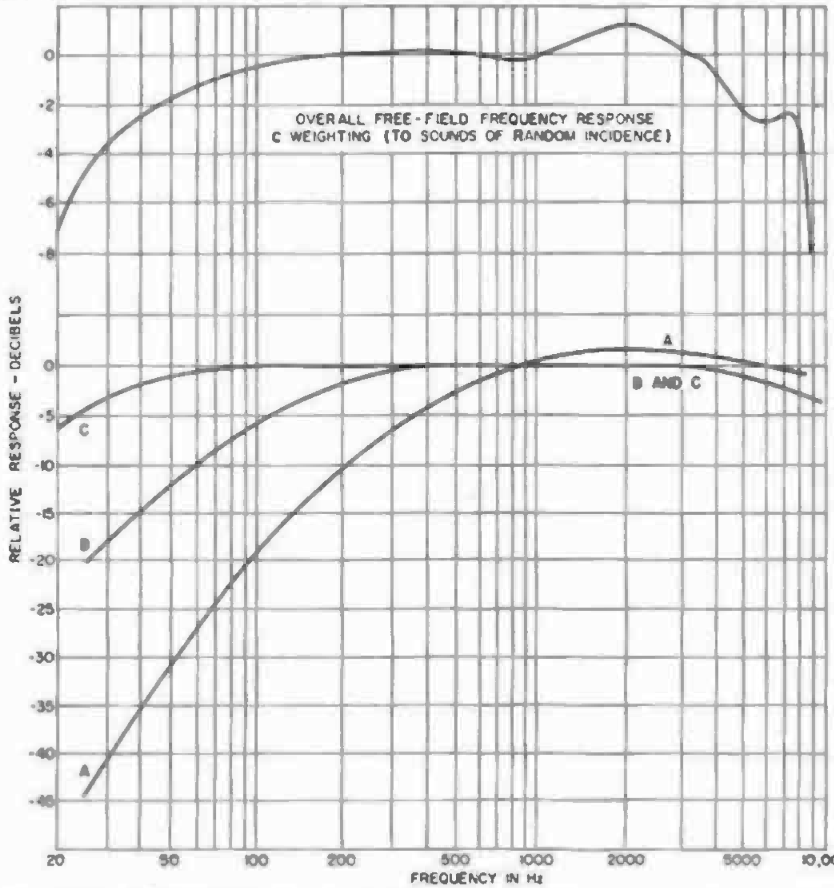

Fig. 2-93. USASI (ASA) Standard S1.4-1961. Weighted curves for sound-level meters. Used in all instruments manufactured since 1961.

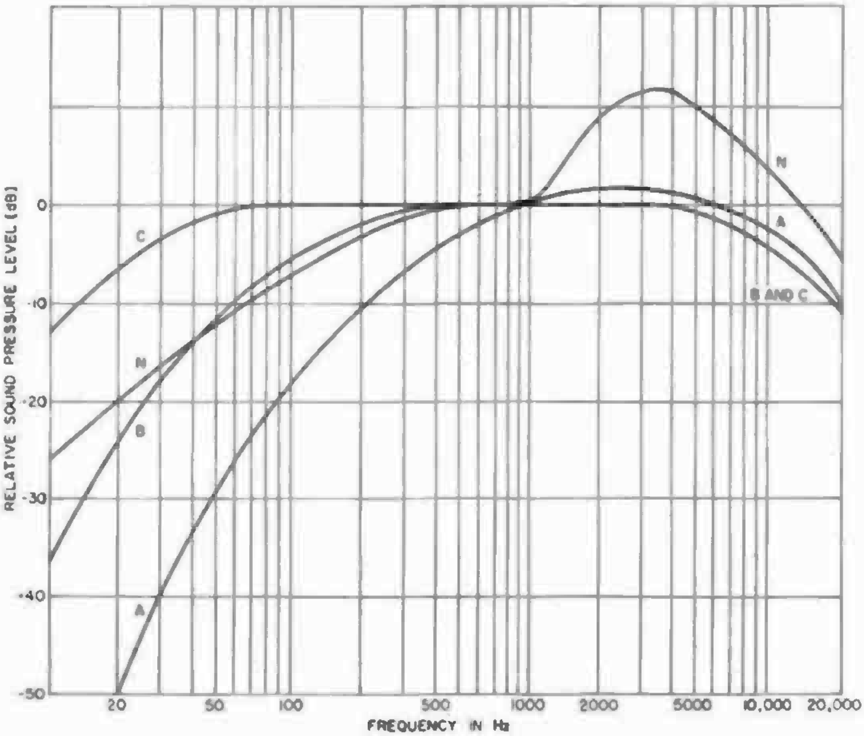

Fig. 2-93C. USASI (ASA) Standard for sound-level meters plotted to indicate the frequency response from 10 to 20,000 hertz.

sound track is played back from loudspeakers on the stage and the action is synchronized with the sound, and photographed. This method is generally used in the production of large scenes in musical motion pictures or scenes where it is impractical to record the action and sound simultaneously.

2.96 ***What are the sentences used for testing sibilance?***—Sister Susie is sewing shirts for soldiers. Or, she sells seashells by the seashore.

2.97 ***How should reverberation be added to program material to produce the illusion of a large auditorium?***—The reverberated signal is fed to the recording channel in such a manner that it is not too apparent, but is the result of natural reverberation in the studio. At appropriate times, such as during rests or other silent spots in the program, the reverberation is increased slightly to give the effect of a large hall. Equalization must be included to correct for the frequency distortion of the echo chamber. (See Questions 2.80, 2.81, and 2.82.)

2.98 ***What does the term truck-in mean?***—It is an expression used in the motion picture industry to indicate that the camera moves in on the subject to be photographed or being photographed.

2.99 ***What is a reverberant chamber, and what is its purpose?***—The reverberant chamber is the direct opposite of an anechoic chamber discussed in Question 2.83. It is constructed of highly reflective surfaces. An important factor of any noise-producing device is the total radiated sound-power level (SPL) or the acoustical energy generated by the device. Sound-power level is proportional to mechanical energy. If the PWL of a device is known before its installation, it is possible to predict the noise level it will produce in its final environment.

Reverberant chambers are generally constructed using sheet-metal walls, and having solid surfaces without any acoustical backing. Stiffening braces are used to hold the surface flat. Other surfaces such as *Masonite*, *Transite*, and similar materials may be used instead of metal. Figs. 2-99A and B are plans and cross-sectional views of typical reverberant rooms. A table of absorption coefficients for a reverberant room using octave-band filters is given in Fig. 2-99C.

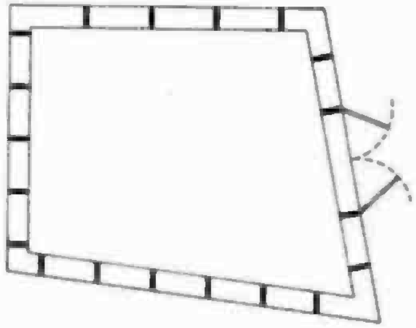

Fig. 2-99A. Plan view of reverberant chamber.

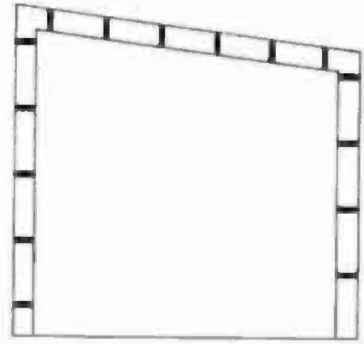

Fig. 2-99B. Cross-sectional view. Cubic volume approximately 2100 cubic feet.

Octave Band Freq.-Hz	*Meas. Abs. Coeff. ( )*
75- 150	0.056
150- 300	0.056
300- 600	0.040
600-1200	0.031
1200-2400	0.029
2400-4800	0.040
4800-9600	0.045

Fig. 2-99C. Typical reverberant room absorption coefficients for structures as shown in Fig. 2-99A and 2-99B.

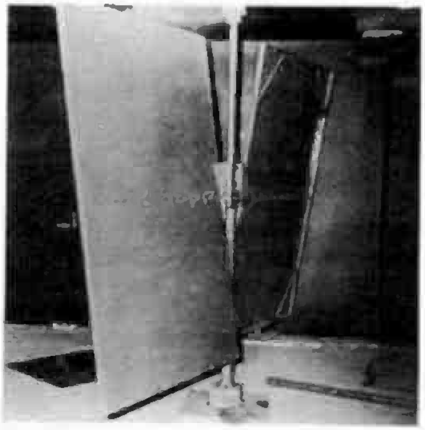

Fig. 2-99D. Interior of reverberant chamber manufactured by Industrial Acoustics Co., Inc. The panels shown in the center of the room can be revolved simultaneously both horizontally and vertically.

If the device being tested has a relatively constant acoustic output, the sound-pressure level will be essentially uniform and constant in a reverberant room. The sound-power level of such items as a fan, motor, pump, or similar device which distributes acoustical energy over a wide band of frequencies can be determined by a few simple measurements in a reverberant chamber. The sound power level may be computed:

$$PWL = SPL + 10 \log_{10} V - 10 \log_{10} T - 19 \text{ dB—re: } 10^{-13} \text{ watt}$$

where,

V is the room volume in cubic feet,
T is the reverberation time of the room in seconds,
SPL is the sound pressure level measured in decibels.

Reverberation time is calculated using the formula given in Question 2.34, or measured with a high-speed level recorder. The terms T and V are essentially constant for a given room, therefore the PWL can be obtained from the SPL measurements, once the room has been measured.

A reverberant chamber should be constructed to meet the following conditions: The walls must not be parallel, and its dimensions must not be similar. The device to be tested is placed near a wall on the floor, never in the exact center of the room. The microphone or device under test should be rotated to avoid measurement of standing waveforms. In Fig. 2-99D, the interior of a reverberant room is shown, with revolving panels that revolve simultaneously, both horizontally and vertically.

Other measurements made using a reverberant room are absorption coefficients and random calibration of microphones, loudspeakers, and other devices. Industry is making extensive use of the so-called quiet room for measurement of household appliances. The room is constructed with panels composed of sheet metal, with an interior perforated metal surface and an acoustical fill sandwiched between the metal sheets. Such rooms are used for the checking of production runs and isolating areas where it would be impractical to install extensive acoustical treatment.

Reverberant rooms may be constructed using 8-inch concrete block. A typical room constructed for microphone and loudspeaker research may have the following inside dimensions: 19.5-feet long, 13.67-feet wide, and a ceiling height of 11.4 feet. The ceiling consists of 2-inch hollow-core prestressed concrete beams. The inside is sealed with *Butyl* rubber-asphalt compound. The walls are coated with ¾-inch gypsum plaster finished with ¼-inch hard lime plaster, and the floor is covered with clear alkyd enamel. The door is a five-inch acoustical door, hung in a double frame, and carefully sealed to the concrete block structure. The inside volume is 3070 cubic feet, with a boundary of 1290 square feet. The measured response for such a room is shown in Fig. 2-99C.

2.100 What is a* Sono-Vox *sound-effects machine?—A device for producing the effect of a train whistle, an automobile horn, or a musical instrument talking. It is used quite extensively in radio commercials. The *Sono-Vox* makes the production of articulated sounds possible although the human voice is not actually used. The human speech mechanism is employed, replacing the normal vocal cord output by the sound it is desired to articulate.

Referring to Fig. 2-100, it will be noted that on the left is shown the dialogue sound track to be articulated and the person who is to act as the articulator. The sound-effects track is connected to the *Sono-Vox* amplifier which drives the *Sono-Vox* articulators. These devices are placed against the throat of the articulator who is seated in front of a microphone connected to a recording channel.

The articulator mouths the desired words but is careful to make no sound. The *Sono-Vox* articulators at the throat will transmit the sound from the sound effects track into the larynx. The sounds emerging from the throat of the human articulator are picked up by the microphone and recorded. Thus, a whistle or bell is made to talk.

2.101 How is a microphone boom constructed?—A microphone boom is a mechanical device consisting of a retractable metal tnbe or rod mounted on a dolly, as shown in Fig. 2-101A. The microphone is suspended at the end of the rectractable tube on a turret head, which may be rotated by the boom operator towards the actors for the best sound pickup.

A typical microphone boom, manufactured by the Fisher Boom Company, is shown in Fig. 2-101A and is used extensively in the production of motion pictures and television shows. The retractable tube may be extended to 16 feet and the microphone turret head rotated through 360 degrees. The bearings are *Neoprene*-sealed ball bearings, for quietness and ease of operation. The boom is demountable and, packed for shipment, weighs approximately 210 pounds.

A large studio-type microphone boom manufactured by The Mole-Richardson Co. is illustrated in Fig. 2-101B. This boom weighs about 500 pounds and is constructed similarly to the one previously described, except it is not readily portable.

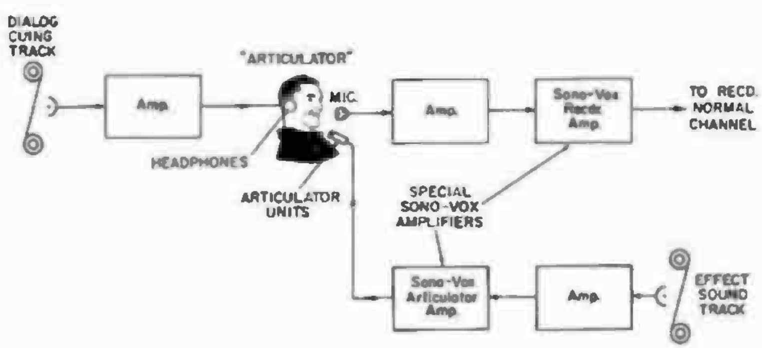

Fig. 2-100. *Sono-Vox* sound effects recording channel.

Fig. 2-101A. A light-weight portable microphone boom manufactured by the Fisher Boom Co.

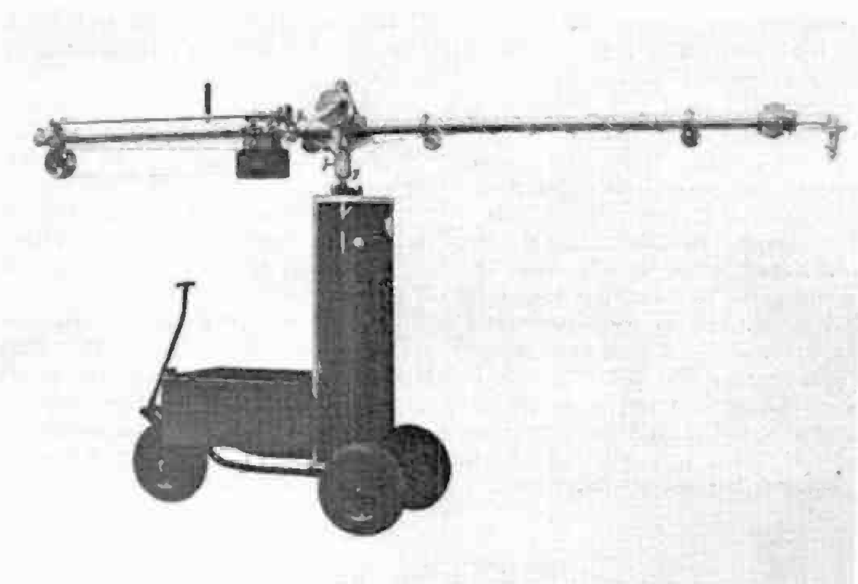

Fig. 2-101B. A heavy studio microphone boom manufactured by the Mole-Richardson Co.

2.102 *What is a turret head?*—A rotatable microphone support on the end of the retractable tube of a microphone boom, as shown in Figs. 2-101A and B. Tilting mechanisms are often mounted on the turret head for the purpose of moving the microphone in a vertical plane.

2.103 *What is the advantage of supplying the microphone boom operator with headphones?*—To aid him in maintaining uniform sound quality. The boom operator's headphones are connected to the same monitor circuit as the mixer's. Thus, the boom-man hears as the mixer hears. As a rule, the boom operator's headphones are also connected to a communication circuit so that the mixer may give him orders when necessary.

2.104 *Describe a radio playback system used for motion picture and television production.*—A system that has been used to good advantage is the radio playback system which utilizes a transmitter and individual receivers. Each principal and others concerned carries a miniature radio receiver connected to a hearing aid earpiece. A low powered radio transmitter is set up on the stage and connected to a loop antenna which surrounds the set. Working within the loop, the actors listen to the music picked up by the miniature receivers and sing and dance as the case may be. Provision is also made for the director to give orders to the actors while they are in action. (See Questions 2.95 and 4.72.)

2.105 *What is a review room?*—A small theater, on a motion picture lot, for reviewing the finished product or daily work. These rooms are maintained to a standard for uniform quality of picture and sound reproduction.

2.106 *What is the peak intensity of sound that may be expected in a review room?*—At least 80 dB above the threshold of hearing; therefore, the walls should be well insulated and diffused.

2.107 *Define the term noise criteria.* —It is the *maximum* permissible noise level for a given enclosure, measured in

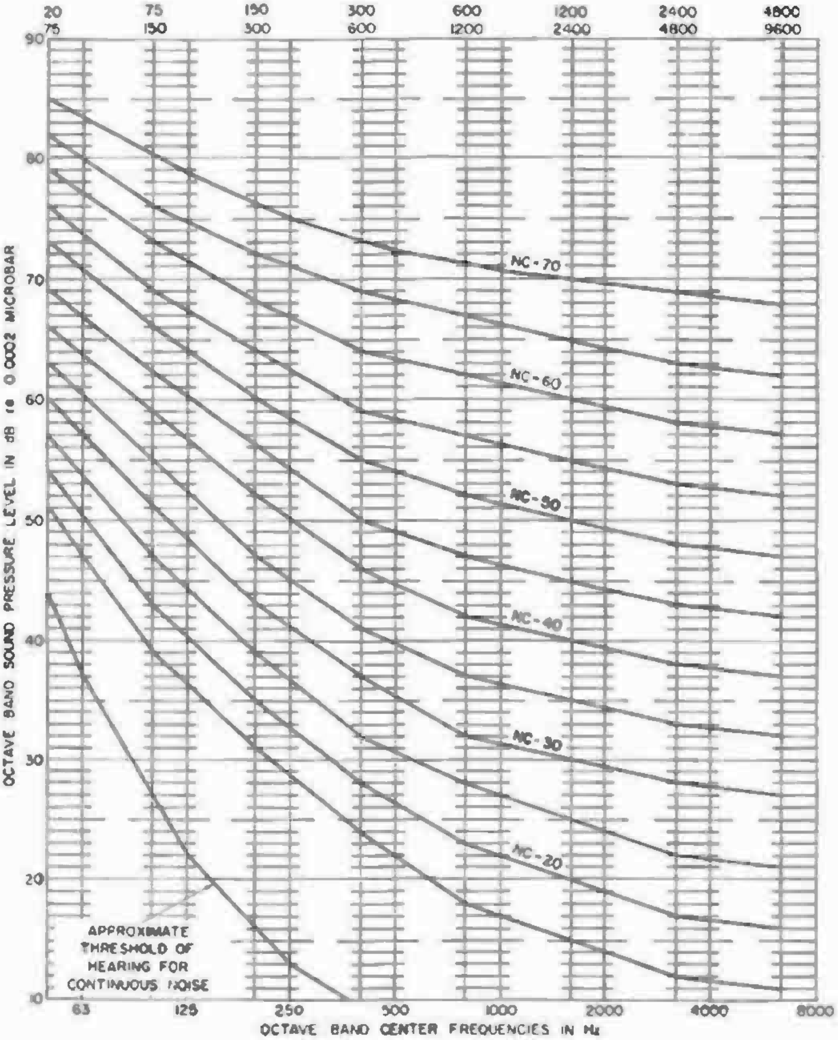

Fig. 2-107. Noise criteria, or NC, curves indicating the maximum permissible noise level for given type enclosures.

octave-bands as a function of the octave-band center frequency. A group of such curves is shown in Fig. 2-107. Curve NC-20 is the one most commonly employed for theaters, concert halls, studios, and similar structures. However, the NC-15 curve should be used whenever possible. The curves include all noises, such as external noises and leak-through from other studios, internal noises, and the ventilating system. The curve at the lower left, is the approximate threshold of hearing for continuous noise.

2.108 ***What is the average ambient noise level of a theater?***—Approximately 40 dB above the threshold of hearing.

2.109 ***Why is it desirable to reduce the low frequencies when recording dialogue?***—To remove tubbyness from the voice caused by playing too close to the microphone; also, to reduce the low-frequency accentuation caused by sets and reflected sounds. (See Question 18.81.)

2.110 ***Describe the construction of a looping stage.***—Looping stages are generally used for post-synchronization of a new sound track for an existing picture. The recording of a new sound track may be necessary for several reasons—the dialogue had noise in the background; it could not be shot on location because of expense or area; certain characters are to be revoiced; a sound track must be replaced due to technical difficulties; or the picture is to be revoiced for another language.

The enclosure of a looping stage is designed so that the walls can be readily altered to change the acoustic conditions to match a particular scene or effect. By moving a group of acoustic panels on the walls, the reverberation of the stage may be changed from a dead stage to a live stage. A group of movable wall panels on a looping stage at Universal Studios, Universal City, California, is shown in Fig. 2-110B, with a floor plan shown in Fig. 2-110A.

Referring to the floor plan, on the north wall is a total of 9 panels, with 5 panels that are movable approximately 180 degrees. Each panel is 4 ft × 16 ft, hinged at the top and bottom, supported by a pipe running parallel to one edge. The inner and outer surface of each panel is treated to be acoustically hard and soft. It will be noted the wall surfaces behind the fixed panels are treated acoustically and are quite dead. The panels are set at an angle of approximately 10 degrees. The fixed panels act somewhat as a flat splay as discussed in Question 2.77. The panels on the west wall are also movable, but are only 2 ft × 13 ft in height, as they must clear the projection room ports above. The treatment of the west and south wall panels is similar to the north wall panels. The stage volume is

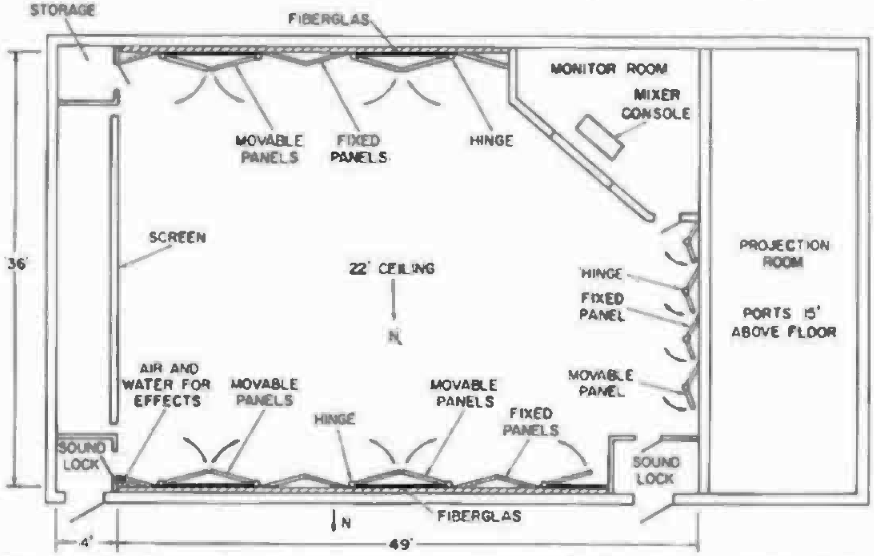

Fig. 2-110A. Floor plan of looping stage at Universal Studios, Universal City, California. In addition to looping operations, this stage is used for making sound effects and can, if necessary, be used as a dubbing stage.

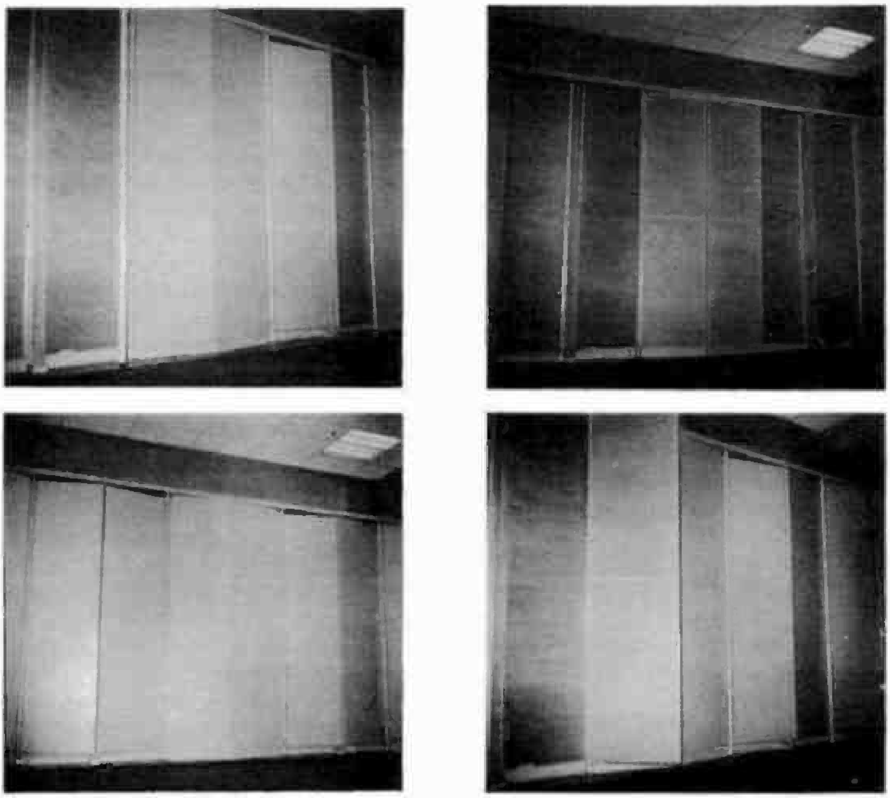

Fig. 2-110B. Interior views of a looping stage at Universal Studios, Universal City, California.

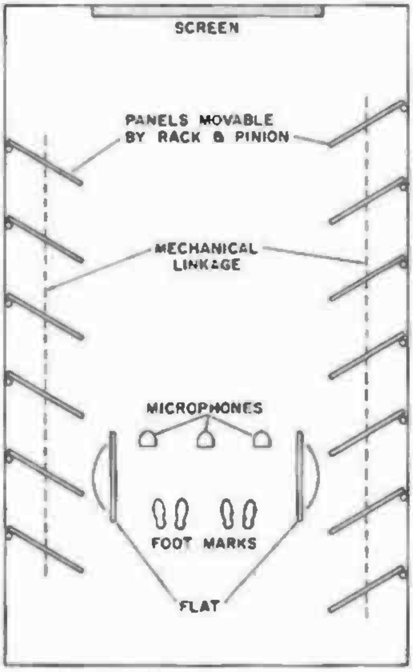

Fig. 2-110C. Looping setup using two flat panels, with movable panels on the walls.

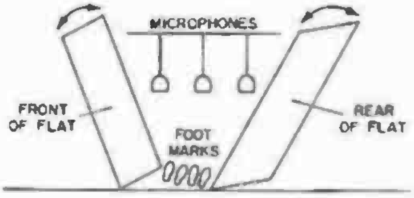

Fig. 2-110D. Looping setup using two panels set at an angle of 10 to 15 degrees.

on the order of 39,000 cu. ft. Wall areas above the panels are set at an angle for diffusion. At the east end of the stage, is a standard motion-picture screen and loudspeaker for projecting the scenes to be looped. The monitor room in the southwest corner of the stage is equipped with a mixing console and other necessary equipment. Outlets are provided on the south wall for microphones, headphones, and intercommunication. Four views of the north wall panels are given in Fig. 2-110B, showing placement for different combinations of the acoustic panel surfaces.

In Figs. 2-110C and D are shown two setups that may be used for looping.

Two flats of the proper acoustical properties are set one on each side of the actors, at an angle of 10 to 15 degrees, with a slight angle with reference to each other. The microphones are hung from the overhead at a fixed distance. Footmarks are placed on the floor for the best position of pickup and future reference. The side wall panels are acoustically hard and soft with the walls behind the panels treated for a high absorption. The panels are linked together, mechanically, and operated by a motor-driven rack and pin arrangement. Thus, for matching the original environment the mixer may adjust the acoustics of the room as he listens to the original sound track.

Looping equipment is discussed in Questions 17.223 to 17.227.

2.111 What Is 3-D sound?—The The term applied to three-dimensional or stereophonic sound.

2.112 What is a monaural sound system?—A sound system consisting of one source of sound, such as a radio, magnetic tape, a record, and the like, using a single loudspeaker for reproduction. It is also termed monophonic.

2.113 What Is a binaural sound system?—A system consisting of two microphones at the point of pickup, placed in the same relationship to each other as the ears of a listener. The microphones are connected to separate amplifier systems and transmit the program material to the listener through headphones. Each headphone is connected to its own amplifier. Such systems are considered to be as nearly perfect as present-day knowledge will permit. The block diagram of such a system is shown in Fig. 2-113. True binaural sound cannot be achieved with loudspeakers, only headphones. When

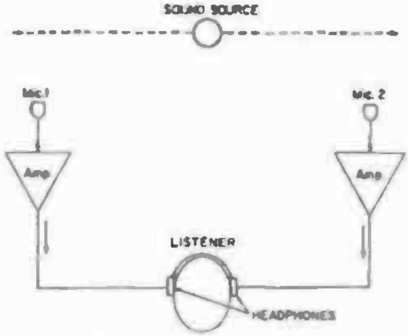

Fig. 2-113. A true binaural sound-reproducing system.

loudspeakers are used it is two channel stereophonic sound.

2.114 What is a stereophonic sound system?—A system using two or more microphones with a separate amplifier and loudspeaker for each microphone channel. This system is also referred to as an auditory perspective system. With such an arrangement of equipment, the sound travels from one speaker to the other as the principals move across the stage. Such a system permits an orchestra to be reproduced closer to its proper perspective.

2.115 Describe how a production stage may be altered acoustically for recording music.—Stages designed for production shooting generally have a reverberation period on the order of 0.5 to 0.6 second, and are not suitable for the recording of music unless the reverberation period is lengthened. Such stages may be used for music recording by the use of a group of polycylindrical diffusers (Fig. 2-76D). These diffusers are placed about 1 foot apart in a semicircle at the rear of the orchestra and slightly forward at the sides, as illustrated in the floor plan drawing of Fig. 2-115.

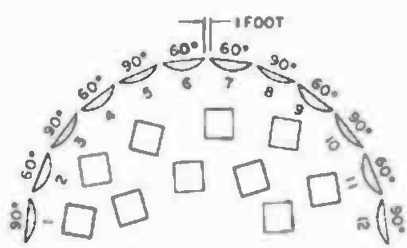

Fig. 2-115. Floor plan for setting up polycylindrical diffusers to record an orchestra. The squares represent the acoustic clouds hung from the overhead.

After construction, the diffusers are backed by hanging a Fiberglas blanket from top to bottom; however, in some instances they may be left open. They are held in vertical position by sand bags at the bottom and stage braces at the sides. Assume a total of 12 panels are to be used, they are made up in pairs having 60- and 90-degree convex surfaces and placed alternately as shown in the illustration.

In addition to the diffusers, acoustic clouds about four feet square made of plywood are hung from the overhead 12 to 15 feet above the orchestra. No two are to have the same exact angle relative to the floor. The surface of these clouds is sprayed with lacquer as are the diffusers. In the construction of such diffusers and clouds, *Masonite* should not be used as it is not stiff enough. Two by four wooden braces set edgewise are used across the upper surface of the flat panels used for clouds which, if not braced, may vibrate, causing rattles and buzzes. (See Question 2.76.)

2.116 What is acoustic gain?—It is the difference in sound pressure level between an unassisted performers acoustic output, and the same performance using a sound reinforced system as measured at a given observation location.

auditorium?—This is an expression used to indicate that the acoustic peaks in an enclosure have been reduced by the use of a series of tuned circuits in the electronics of the sound system feeding the enclosure. Although this method of correcting the acoustic characteristics of an enclosure was used by Volkman of RCA in motion-picture theaters in a simplified form as early as 1930, it has remained for Dr. C. P. Boner and his son, C. R. Boner, with the aid of modern test equipment to realize the full capabilities of such a method.

To tune an auditorium, a small loudspeaker is placed about two feet behind a microphone placed in the position where it will be normally employed. The loudspeaker is directed into the microphone and the auditorium, and is energized using a random-noise generator. The output of the speaker is adjusted for an SPL of about 80 dB, as measured on a sound-level meter close to the microphone. The initial adjustments are accomplished with the sound system off. A reading of the speaker level fed by the random-noise generator is then taken at the extreme rear of the auditorium with a sound-level meter. The sound system is now energized with the random-noise generator and the gain adjusted just below the point of feedback.

At this point of gain, the SPL of the sound system is noted. Subtracting the first measurement from the second gives the acoustical gain of the auditorium. As a rule the average auditorium will yield about 1 to 2 dB of acoustic gain. With the auditorium properly tuned, up to 30 dB of acoustical gain is possible. This will permit the full frequency range to be realized, thus increasing the intelligibility of speech and also enhancing the reproduction of music.

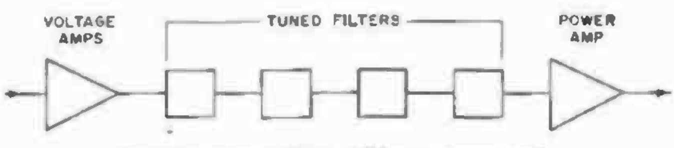

Fig. 2-117A. Voltage and power amplifier with tuned filters.

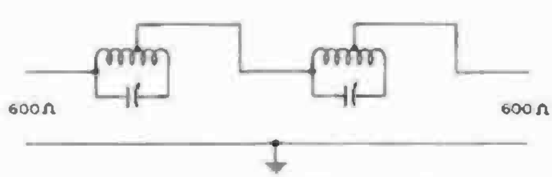

Fig. 2-117B. Tuned-filter circuitry.

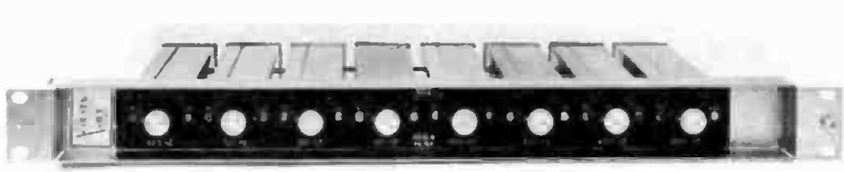

Fig. 2-117C. Front view of Altec-Lansing Acousta-Voice variable filters. Range 625 to 8000 Hz, variable in steps of 1 dB for a total range of 14 dB.

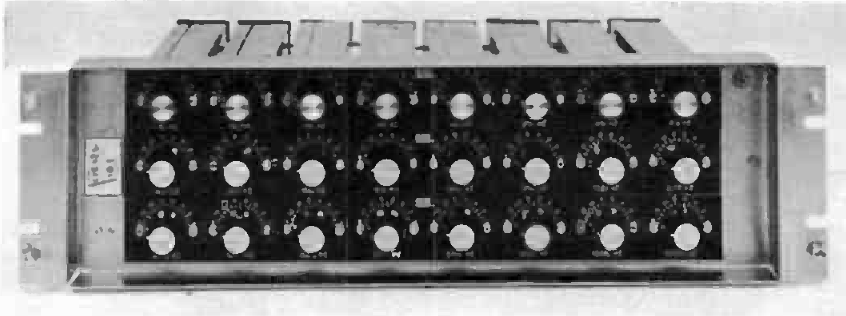

Fig. 2-117D. Front view of Altec-Lansing Acousta-Voice variable filters. Range 60 to 12,500 Hz, variable in steps of 1 dB for a total range of 14 dB.

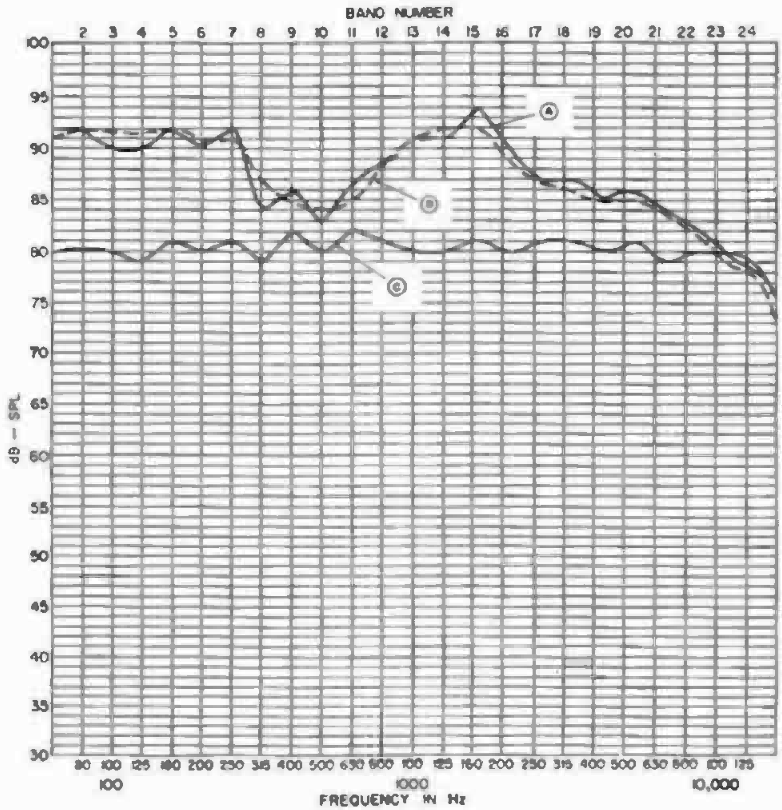

Fig. 2-117E. Response curve of the Acousto-Voicing system.

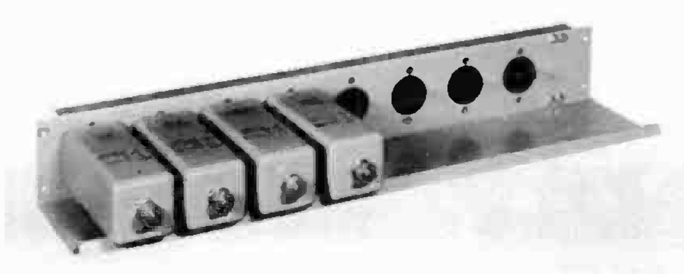

Fig. 2-117F. Tuned filter sections manufactured by the DuKane Corp. used in their *Varacoustic* sound control system. (Courtesy, Pacific Communications Ltd.)

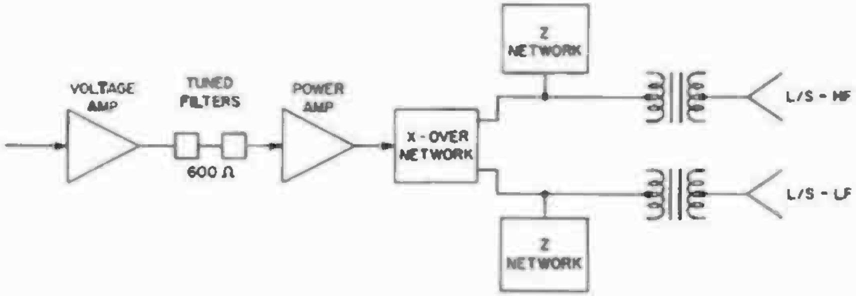

Fig. 2-117G. Impedance-correcting networks used with the DuKane *Varacoustic* sound control system.

Dr. Boner's method of tuning an enclosure is to employ a group of tuned filters between a separate voltage and power amplifier, transformer coupled by a 600-ohm impedance line (Figs. 2-117A and B). It is in this line that the tuned filters are connected.

Through the use of such filter sections, the gain of the amplifier system is reduced at the *peak frequencies* of the auditorium, resulting in an *inverse electrical correction that simulates a uniform (or approximately uniform) acoustic response.* It is not uncommon in some auditoriums to encounter peaks on the order of 20 dB or more. It should be understood the acoustics of the auditorium are not corrected, but the frequency characteristics of the amplifier system are *altered inversely* to the auditorium characteristics to compensate for the acoustical peaks. The amplifier system is of *uniform frequency* characteristics without the filters.

This system of correcting for acoustic peaks is generally used where one or more microphones are employed. *This method of acoustic correction for motion picture theater reproduction is not recommended at this time, since the sound system would not be compatible with the frequency characteristics used for recording motion pictures.*

After inserting the first peak filter, the sound system is again energized and the next largest peak noted. A tuned filter of the correct frequency is inserted in the link circuit and adjusted for the peak amplitude. This same procedure is followed until all the major peaks are removed. Small peaks may now be removed in order of their importance. It is possible in a large auditorium to employ 20 to 100 tuned filters. In a well-tuned auditorium, an acoustic gain of 15 to 20 dB can be expected. Notch filters suitable for this work are discussed in Question 7.12. (See Question 2.116.)

A second method of tuning an auditorium, developed by Arthur C. Davis and Don Davis of Altec Lansing Corp., accomplishes similar results, except the tuned-circuits are constant-impedance bridged-T filters spaced ⅓ octave apart. It is claimed for this method of tuning that *hangover and ringing* effects in the tuned circuits are avoided; also, the impedance remains constant, whereas tuned circuits are not of constant-impedance. In setting up this latter system, a method of determining the peak frequency of the auditorium and adjusting requires only the use of an oscilloscope. This system has been termed *Acousta-Voicing* by Altec Lansing.

Pictured in Figs. 2-117C and 2-117D are front-panel views of the Altec-Lansing system, showing the filters and their controls. The small group is used where a small number of filters is required. The larger group will supply corrections for 24 different frequencies.

A response curve (A) taken from the left hand loudspeaker of a two-channel stereophonic reproducing system in a given enclosure is indicated in the graph of Fig. 2-117E. Frequencies of constant amplitude were applied to the amplifier system, and the acoustic response of the enclosure was measured by means of a sound-level meter. The frequency response of the amplifiers without the filters was uniform over the indicated frequency range given at the lower edge of the graph.

Superimposed curve "B" is the *inverse* electrical response of the bridged-T filter sections alone. Each filter section is capable of being attenuated 14 dB. Greater attenuation for a given frequency can be achieved by adding a second filter of the same frequency. The lower curve "C" represents the *acoustical* response of the enclosure after the introduction of the various filter sections, again read with a sound-level meter. It will be observed that the rising characteristic at the low-frequency end and the midhigh frequency portion of the spectrum has been reduced, resulting in the listener hearing a more uniform acoustical response in the auditorium. The right-hand amplifier system was treated in a similar manner.

If the problem is one involving only two or three peaks, peak frequencies may be reduced by connecting tuned circuits or other type filters in the output of the power-amplifier feed line to the speakers. (See Fig. 20-137.) Or, the amplifier system may be altered as discussed in Question 6.96 under Equalizers, a method still used in some theater equipment. When using filters in the output line of a power amplifier, the insertion loss of the filter must be quite low; if it is not, considerable power will be lost.

In evaluating the two systems, the first system described employs filters tuned to the *exact frequency* of the auditorium peak frequency. In the latter system, the filter frequencies are *fixed*, but separated in ⅓ octave intervals. Therefore, it is possible that a filter frequency may be almost on the peak but not exactly. Also, the adjustment range is in 1-dB steps, whereas the tuned-filter type is adjusted to the *exact amplitude* of the peak.

Using either of the two systems in an auditorum results in greater acoustic gain by the reduction of acoustic feedback and the removal of objectionable peaks in the auditorium. Listening to reproduction through either of the above two systems, a noticeable improvement is observed.

A third system, manufactured by the DuKane Corp., termed *Varacoustic* sound control is pictured in Fig. 2-117F. Although only four filters are shown, any number may be employed, and they are similar to those employed by Dr. Boner. In addition to the filters (connected in a 600-ohm link between the voltage and power amplifier) are impedance-correcting networks in parallel with the lines from the crossover network feeding the loudspeaker sections. These networks tend to hold the speaker impedances fairly close to their design value which normally varies over a considerable range. (See Questions 20.89 to 20.91.)

2.118 Show a typical microphone placement plan for monophonically recording an orchestra and vocalist.—Such a plan is shown in Fig. 2-118, using a split recording channel. The final microphone placement will be governed

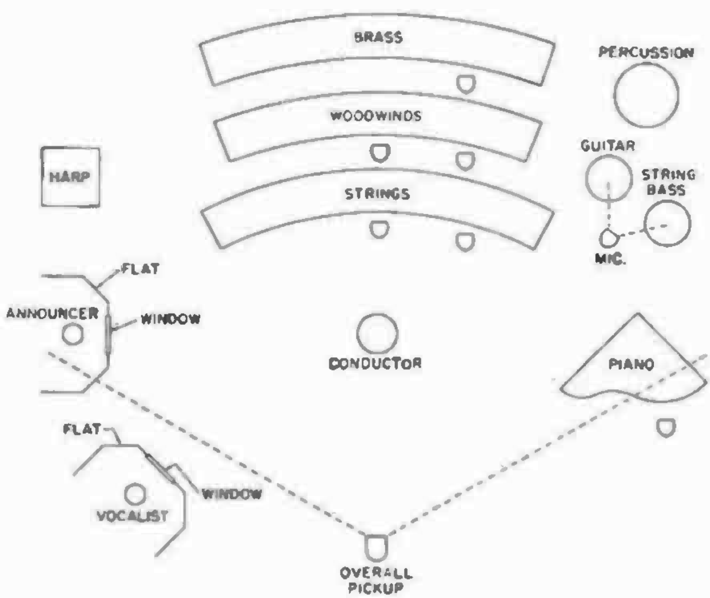

Fig. 2-118. Suggested microphone floor plan for monophonic recording of a large orchestra and a vocalist.

by the solo instruments, number of musicians, and the type of music to be recorded.

2.119 What is meant by presence? —A quality in the reproduction that gives the impression to the listener that the person speaking is present in the room.

2.120 How is presence obtained in recording?—By microphone placement, proper equalization in the midfrequency range, the acoustics of the recording studio, and the signal-to-noise ratio of the studio and recording system.

2.121 In what part of the frequency band is presence equalization introduced?—In the frequency range from 1500 to 5000 Hz. Recording characteristics are discussed in Question 18.81. These same recording characteristics are also used with magnetic tape and disc recording. Presence equalization may also be added during rerecording or when transferring sound tracks.

2.122 How are microphones set up on a music-scoring stage?—There are no fixed rules regarding microphone placement on a music stage; however, the usual method is to employ individual microphones for the solo instruments and vocalist. Separate microphones may be used to build up the woodwinds and strings, depending on the orchestration, and if the pickup is monophonic or stereo. (See Question 2.74.)

2.123 What is the recommended placement for a single microphone orchestra pickup?—Approximately 20 to 30 feet in front of the orchestra and snspended at a height of 12 to 18 feet above the floor. A capacitor or amplitude-modulated microphone is often used for this type pickup. If too much reflected sound is heard, move the microphone closer to the orchestra. This will also improve the definition. For dance bands a higher order of definition is required than for a large orchestra. One or more microphones may be required for solo instruments and a vocalist. Velocity microphones with directional characteristics should be used for the latter positions.

2.124 Show the microphone placement for a small dance band.—Such a setup is shown in Fig. 2-124, and is typical of that used for broadcast and recording. Each microphone covers a group or an individual instrument. The microphones should be placed close to the instruments to obtain a high order of definition.

2.125 Show the microphone placement for a small television show.—A typical setup for a small television studio, presenting action on a set, intermixed with commercials, and using a musical background, is shown in Fig. 2-125. A unidirectional microphone should be used on a boom to reduce reverberation and background noise of the stage.

2.126 How are microphones placed for a radio broadcast show?—The presentation of a radio broadcast show is slightly different from that of other type microphone pickups. The actors

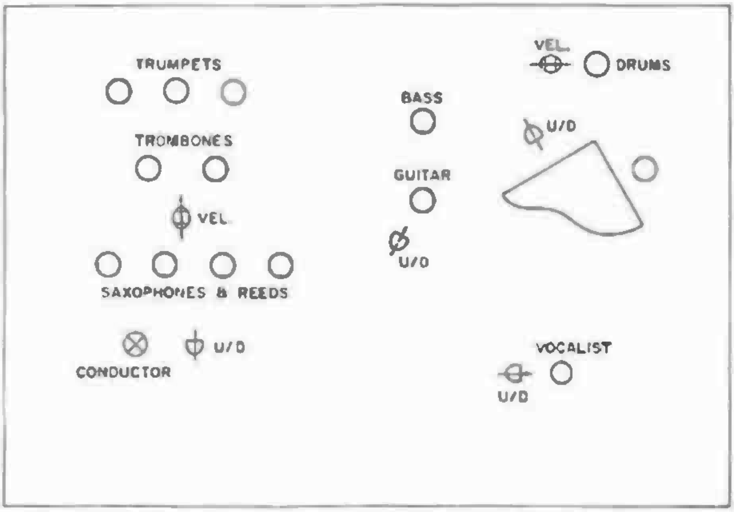

Fig. 2-124. Typical instrument and microphone setup for a small dance band.

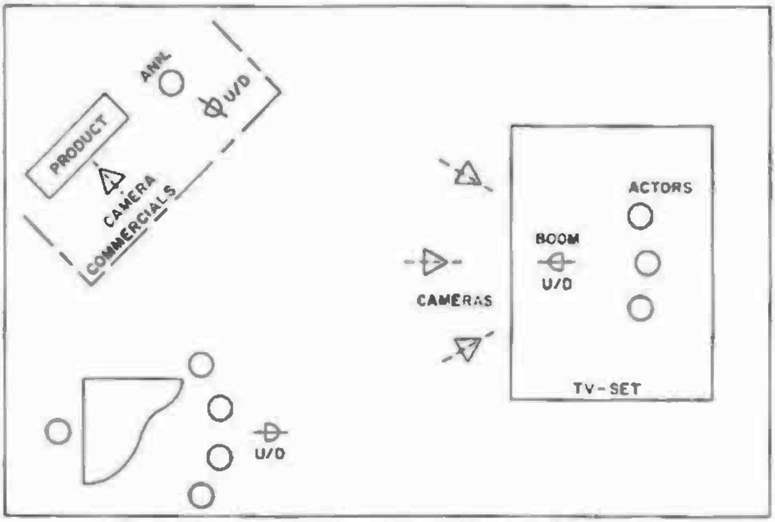

Fig. 2-125. Typical setup for small television studio.

are generally placed on each side of a ribbon-velocity microphone (figure-8 polar pattern) to eliminate the need for stepping up to the microphone. The orchestra and announcer use unidirectional microphones to reduce noise and pickup from other parts of the production. The sound effects man is placed behind flats, for isolation (Fig. 2-126). A unidirectional microphone is also employed here. (See Question 4.47.)

2.127 ***Show the basic placement for recording or broadcasting a large symphony orchestra.***—The basic plan for microphone placement, when recording or broadcasting a large symphony orchestra, is shown in Fig. 2-127. Although only a few microphones are shown, actually in such a large operation microphones would be placed in several other positions to pickup a soloist or a group of instruments. This would be particularly true for television. An overall microphone is suspended about 15 feet above the orchestra, and about 50 to 25 feet in front of the first row of instruments. Considerable care must be taken to balance the orchestra for a uniform pickup. If the recording or broadcast is to be released

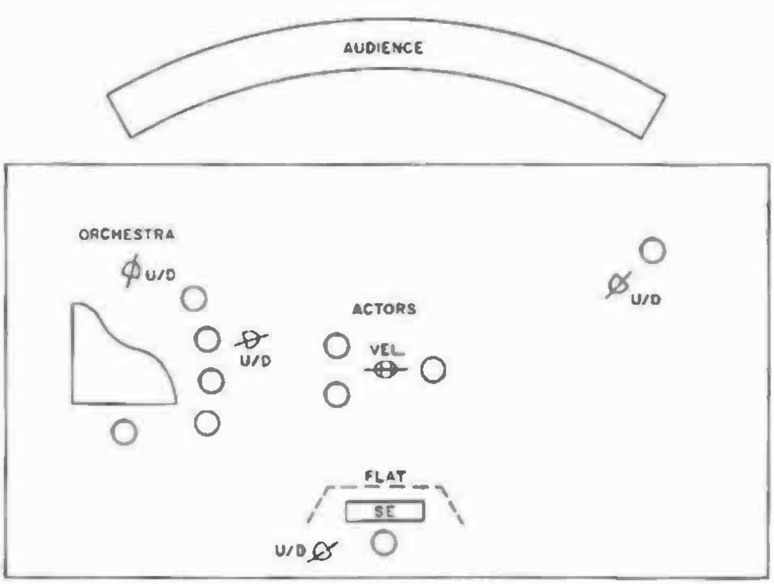

Fig. 2-126. Typical microphone setup for radio broadcast show.

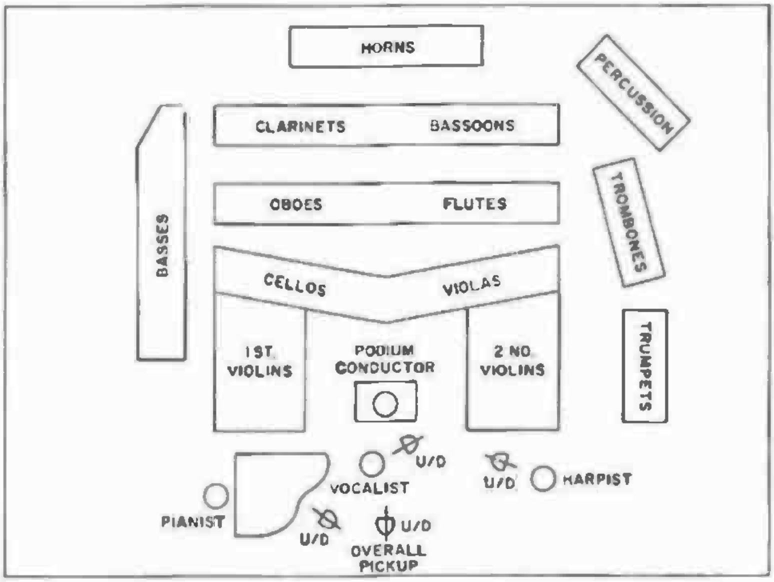

Fig. 2-127. Basic setup for large symphony orchestra.

stereophonically, dual microphones and special placement techniques are employed.

2.128 Describe the different methods used to produce artificial reverberation.—Over the years there have been many different methods of generating artificial reverberation. Among these methods are magnetic recorder-reproducers employing several reproducer heads spaced at intervals along the tape to induce delay. Another method uses long metal pipes laid in a straight line or coiled, with a loudspeaker unit at one end and a microphone at the other. Additional microphones are placed at intervals along the pipe to induce delay. Another design, the echo chamber, propagates the sound by means of a loudspeaker in a room with hard walls. The sound is fed to a loudspeaker and picked up by a microphone at the opposite end of the room. Another method employs the transmission of sound waves through a group of steel springs or liquid. In this design, vibrations are picked up from the sides of the tank, thus inducing delay.

The most commonly employed instrument is one using a large metal plate. This method is discussed at length in Question 2.130. (See Question 2.80.)

2.129 Describe the construction of a reverberation unit employing steel springs.—Reverberation units employing steel springs are often used with electronic organs and in some type units designed to be used with radio receivers. Reverberation is the prolongation of sound waves by repeated reflections, and if the reflection is long enough to be delayed 1/20 second, it is then classed as an echo.

The device shown in Fig. 2-129A is electromechanical, and introduces multiple echoes by means of reflections set up in a network of coiled springs about 3 feet in length. The audio signal, when fed into the reverberation unit, is converted into mechanical energy by a moving-coil driver unit at the top of the spring assembly. The driver unit is similar in design to a dynamic loudspeaker unit, but without a diaphragm. Waveforms are transmitted through the coiled springs, which have the property of conducting the sound waves more slowly than the speed of sound through air. In this manner, a spring of sufficient length can be made to produce a delay comparable to that in a large auditorium.

The driver unit (A) induces vibrations into the stirrup (B) by its vertical motion. Springs 3 and 4 hold the stirrup in position, but permit it to move with freedom in the vertical plane. Spring 1, at the far left, balances the tension of the other springs, through lever (F) above the driver unit. Springs 1, 3, and 4 are almost entirely immersed in oil, as they act largely as dampers, to stabilize the response of the driver unit and to prevent undesired reflections.

Sound waves from the stirrup travel downward through the open spring 5 at the far right, to a ceramic pickup (C), where the mechanical signal is converted to an electrical signal and fed to a preamplifier. The first reflected signal is delayed about 1/5 second from the original signal, which goes directly to the output amplifier. The same waveform also travels downward through spring 2 which enters a small oil damping tube. At the bottom of this spring, the wave is reflected back at a reduced intensity caused by the damping action of the oil. At the stirrup, the horizontal lever (D) transfers the wave to the right hand spring 5, then to the pickup (C), producing a second reflected sound, which is about 1/18 second behind the direct signal. Little energy is absorbed by the pickup unit and the wave again is reflected back through spring 5. The first reflected signal transverses spring 5 and is transferred by lever (D) and travels downward through spring 2, to the short oil tube. Here again the wave is reflected and reduced in intensity. It then retraces the same path to the pickup, and produces a third reflection which is about 1/15 second behind the direct signal. The second reflected signal is similarly repeated, and this process continues over and over, resulting in a series of signals about 2/15 second apart, until they are dissipated by the friction of the oil in the short tube.

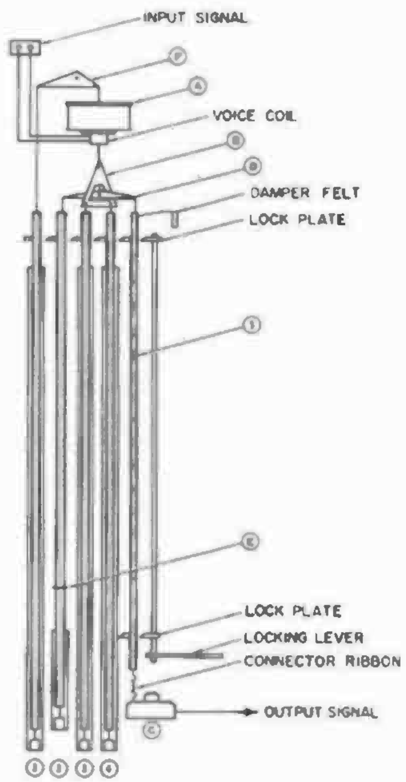

Fig. 2-129A. Interior construction of an electromechanical reverberation unit employing coil springs.

Above the short oil tube is placed a reflecting pin (E) attached to spring 2 which causes a partial reflection, thus smoothing out the overall response. The higher the level of the oil in the short tube, the less the number of reflections. Therefore, adjusting the oil level changes the reverberation time. Generally when the oil level has been established once, it is not changed, and only the electrical controls are used to control the amount of reverberation. A block diagram of the external connections is shown in Fig. 2-129B. The use and connections for reverberation units and controls in conjunction with recording consoles, are discussed in Section 9. (See Question 2.128.)

2.130 ***Describe the construction of a reverberation unit employing a steel plate.***—Because of the limitations of the

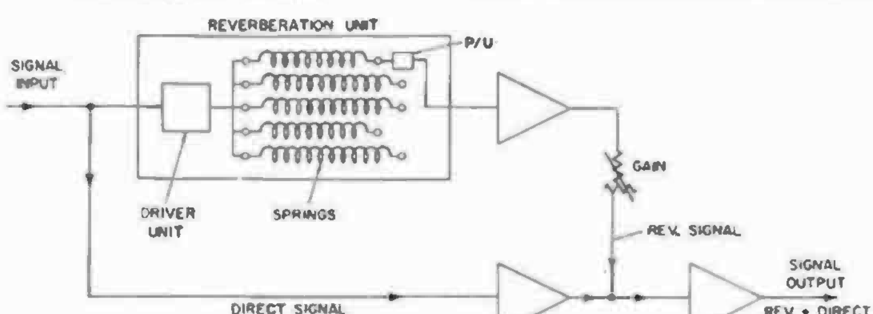

Fig. 2-129B. External connection for an electromechanical reverberation unit.

various methods of producing artificial reverberation, discussed in Question 2.128, and the space required for the construction of an echo chamber (Question 2.81), other means had to be developed for the generation of controlled synthetic reverberation. In Figs. 2-130A and 2-130B is shown a device invented by Dr. W. K. Kuhl, of Hamburg, West Germany and manufactured by EMT of Germany. The reverberation unit is obtainable for either monophonic or stereophonic reproduction. However, the stereophonic type may be used for either monophonic or stereophonic reproduction. As the basic principles are the same for either type, only the stereophonic type will be discussed.

The EMT reverberation unit utilizes the physical properties of metal to achieve its effect, by the use of a specially selected annealed steel plate. The plate is excited by an impulse which sets up within the plate, bending

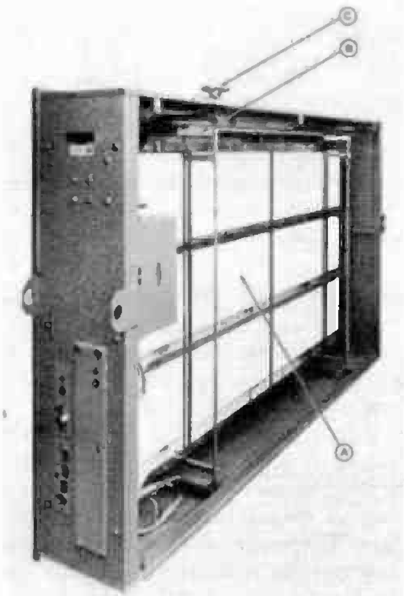

Fig. 2-130A. Front view of EMT Model 140st stereophonic reverberation unit showing the damping-plate assembly and external connections to the amplifiers. (Courtesy, Gotham Audio Corp.)

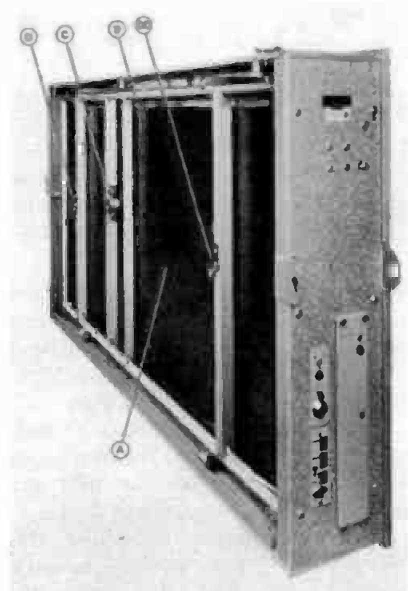

Fig. 2-130B. Rear view of EMT Model 140st stereophonic reverberation unit showing the two pickup units at the ends and the drive unit in the center. (Courtesy, Gotham Audio Corp.)

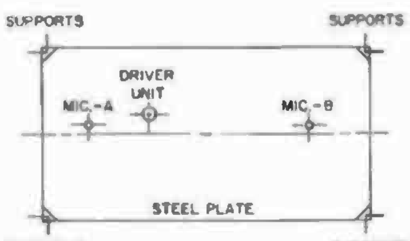

Fig. 2-130C. Placement of the microphones and driving unit for an EMT Model 140st stereophonic reverberation unit.

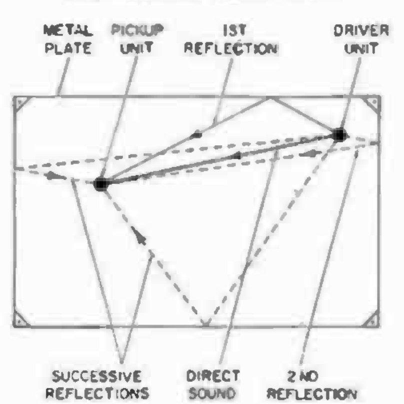

Fig. 2-130D. Reflection paths in the steel plate of an EMT 140 monophonic reverberation unit.

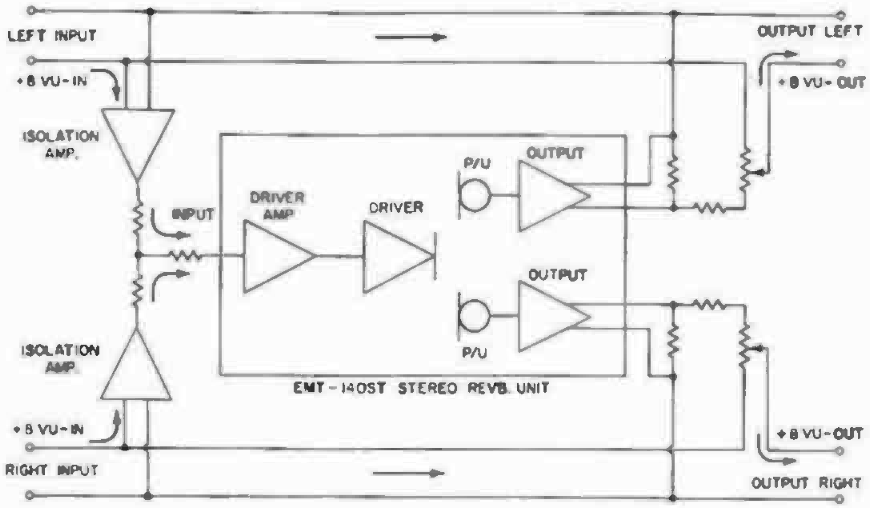

Fig. 2-130E. Block diagram of the internal and external associated equipment for EMT 140st stereophonic reverberation unit. The arrows indicate the two paths of signal travel. (Courtesy, Gotham Audio Corp.)

oscillations that will produce reflections which increase in density with time. Reflections in a three-dimensional room, such as an echo chamber (Question 2.81), become more dense as a function of the square of time. In comparing the two systems, the human ear is unable to recognize the difference between these two operating modes.

Through appropriate steel and critically chosen dimensions, it is possible to produce a plate that will have an adequate number of oscillations. The length and frequency response of the decay time produce an artificial reverberation effect which is comparable to that obtained using a conventional echo chamber. Referring to Fig. 2-130B, the principal component of the unit is a 1/32-inch steel plate (A), suspended at its four corners by steel wires under tension, in a tubnlar steel frame (B). At the front (Fig. 2-130A), parallel to the steel plate is a Fiberglas plate (A) suspended on a pantograph mechanism to permit its movement toward or away from the steel plate. At a position of 1/8 inch from the plate, the reverberation time is reduced to 1 second at 500 Hz. *At no time does the absorbent material come in actual contact with the steel plate.* The movement of the damping plate may be accomplished by the use of a manual control (C), or by the use of a remote control from the mixer console. The reverberation time is indicated in fractions of a second on a remote meter at the console.

In selecting the material for the plate, its internal damping and the resulting reverberation must be taken into consideration. Losses in the plate are additively formed by the nonfrequency-dependent and the frequency-dependent

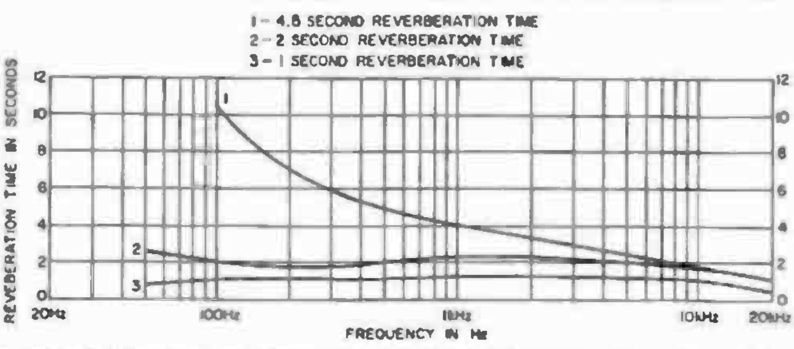

Fig. 2-130F. Reverberation times as a function of frequency for various distances between reverberation and damping plate.

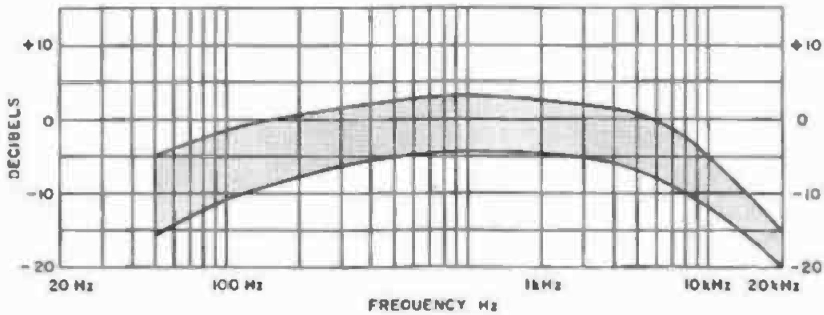

Fig. 2-130G. Tolerance range of the frequency response for an EMT 140st reverberation unit, using white noise with ⅓-octave filter on 25-Hz warble-tone generator. A 2-second reverberation time.

parts which are caused by the heat conductivity loss of the bending modes. For the high frequencies the nonfrequency-dependent parts predominate. For the midrange and low frequencies, the frequency-dependent parts predominate. Since phase velocity of the bending mode of the plate in the entire frequency range is smaller than the velocity of air, damping of the plate through the greatly reduced radiation of airborne sound may be neglected, when compared to other damping causes. Damping is approximately directly proportional to the frequency, and inversely proportional to the thickness of the plate.

The plate is of a high quality cold-drawn steel, approximately 3 feet × 6 feet × ⅟₆₄ inch in thickness, and will cause a time delay of about 5 seconds at a frequency of 500 Hz. The plate must be completely undamped, and extremely flat. Referring again to Fig. 2-130B, the tubular frame supporting the plate has three transverse bridges. One carries the moving-coil driver unit (C), while the other two carry two crystal contact microphones or pickup units, (D and DD). The frame also supports the damping mechanism for controlling the length of the reverberation time. (See Fig. 2-130C).

The steel plate is excited by an audio signal applied to the driver unit, which is similar in construction to a dynamic speaker unit, but without a diaphragm. The reverberated signal is picked up by the contact microphones, (D and DD). Since these microphones are acceleration-sensitive, their output voltage rises at the low frequencies around 250 Hz, stays constant up to 900 Hz, and falls off at a rate inversely proportional to the frequency. The resonant frequency of the microphones is well beyond 20,000 Hz. The internal capacitance of the crystal microphone is approximately 500 pF. The vibrational patterns set up in a plate for a monophonic unit are shown in Fig. 2-130D.

At a frequency of 1000 Hz, the travel time between the driving unit and the microphone is about 6×10^{-3} seconds, which equals the time a sound wave in air travels 6 feet. Because of this short time of travel of the flexing waveforms, the successive repetitions follow in rapid sequence, with the number of reverberations growing with time.

When adding reverberation to stereophonic recordings, two separate conditions must be satisfied. First, it must extract from the stereo signal its direct componeut, and second, it may not as a result adversely affect the significant information contained therein. To achieve this eud, use is made of a so-called "M" channel, which is formed by the addition of two stereo signals, according to the formula, $A + B = M$ (See Question 4.62). This is accomplished by feeding a part of the unreverberated signal output to microphones D and DD through isolation networks, to a common bus. For compatibly recorded signals, this addition of signals D and DD into an "M" channel produces a monophonic signal containing all the information of the two stereophonic signals. The two reverberated signals from the microphones are incoherent and bear no relationship. This is an important factor, for the two resulting stereo signals with echo must have between them a statistically distributed directional and informational content. The frequency response of the reverberation time, without additional damping, corresponds approximately to that of an empty stone-walled church, or about 5 seconds at 500 Hz. At the low-frequency end there is a rise, and toward the high frequencies a decline of about 1.5 seconds, out to 10,000 Hz.

The driving amplifier is so designed that the third order harmonics will not exceed 0.6 percent for a peak level of 1.55 volts at the input, using white noise through a ⅓-octave filter. This takes into consideration the statistical power distribution of sound modulation which normally drops toward the higher frequencies. Signal-to-noise ratio, measured at the output of one contact microphone preamplifier, using a reference frequency of 300 Hz with a reverberation time of 2 seconds, is greater than 60 dB.

Several of these units may be installed in a single room, without fear of interference, as there is practically no sound radiation. The ambient noise level of the room should not exceed 40 dB SPL. To eliminate low-frequency noises transmitted through the structure, the units should be set on resilient mountings, spaced about 1 foot apart.

A block diagram of the internal and external connections for the stereophonic unit and its associated equipment, is given in Fig. 2-130E. The reverberation characteristics as a function of frequency for various distances of the damping plate from the steel plate, is shown in Fig. 2-130F, and the frequency tolerance, using a white-noise generator and ⅓ octave filter, at a reverberation time of 2 seconds, is given in Fig. 2-130G. The same information may be obtained, using a 25 Hz warble oscillator. The internal input impedance is 1000 ohms, and the internal output impedance is 25 ohms, and is designed to operate into a 200-ohm load impedance.

The complete unit is mounted in a wooden case, and weighs approximately 400 pounds. The external and internal appearance of the monophonic unit is the same, except for the additional amplifiers and contact microphone.

The installation of the described reverberation unit is quite critical, if the maximum capabilities are to be realized. The manufacturer's instructions must be followed closely, as the adjustment of the steel-plate tension is all important. The use of reverberation equipment is discussed in Section 9.

2.131 What is a tempo regulator?—A device that may be attached to a standard magnetic-tape recorder, for changing the program length without changing the pitch, or changing the pitch without changing the program length.

In Fig. 2-131A is shown a tempo regulator, manufactured by ELTRO GmbH & Co., Heidleberg, West Germany, threaded for tempo regulation. For pitch control the machine is threaded as shown in Fig. 2-131B. For pitch control the most critical component is the rotating-head assembly, which consists of a single magnetic head-coil, with four separate playback headgaps 0.0002-inch in height, spaced 90 degrees apart at the perimeter. It is essential that the four gaps match perfectly with respect to three important details. First, the angular spacing must be exactly 90 degrees. Second, the four gaps must be absolutely parallel to each other. Third, the output from all four heads must

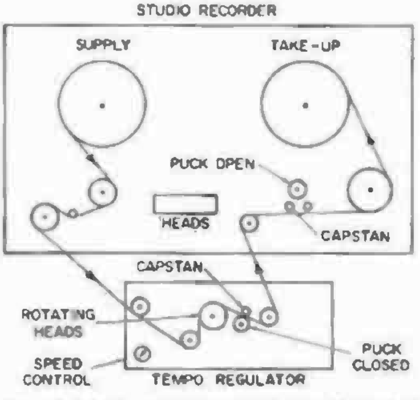

Fig. 2-131A. Tempo regulator threaded for tempo regulation.

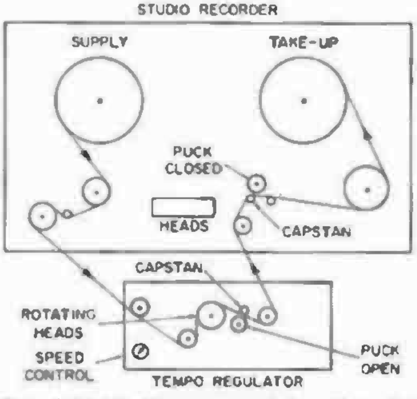

Fig. 2-131B. Tempo regulator threaded for pitch regulation.

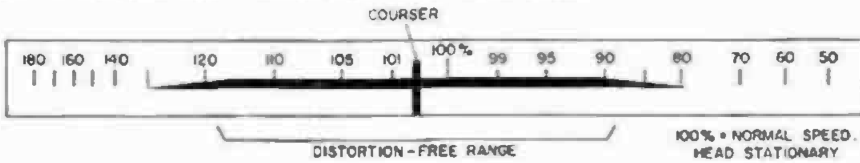

Fig. 2-131C. Expansion-compression control scale on an Eltro tempo regulator unit.

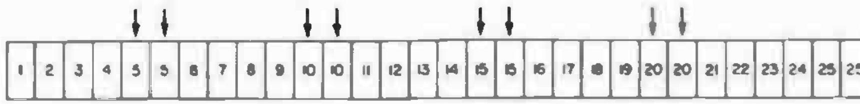

Fig. 2-131D. Playback with an expansion of 20 percent. Every 5th section is repeated as indicated by the arrows.

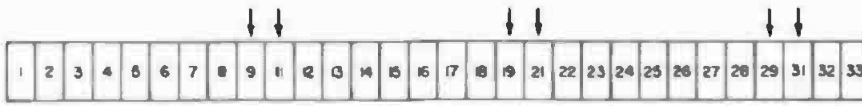

Fig. 2-131E. Playback with a compression ratio of 10 percent. Every 10th section is omitted.

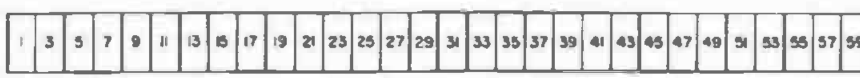

Fig. 2-131F. Playback with a compression ratio of 50 percent. Every 2nd section is omitted.

correspond within 1 dB, over a frequency range of 30 to 15,000 Hz.

The head-coil is comparable to those used on a standard tape recorder. The output of the rotating-head assembly is connected in place of the playback head on the studio recorder. The magnetic tape is threaded around the rotating-head assembly, to make a 90-degree wrap-around covering two head-gaps. During playback, the rotating head scans the tape by one gap, leaving the tape as the next gap starts to scan. This action provides continuous reproduction regardless of the tape speed or the rotational speed of the head.

The capstan of the regulator unit is fixed to a rotating stator of a drive motor. This stator is driven by an auxilliary motor over a constantly variable transmission. The playback-head assembly is driven by the rotor of the drive motor through a gear transmission. Both motors are synchronous; therefore, the speed of the capstan, in relation to the speed of the playback head, remains constant.

Consider now, what actually determines the pitch. It depends on the speed of the head gap relative to that of the tape. If the tape is threaded in the standard recorder, as shown in Fig. 2-131B and the speed indicator dial set to 100 percent, the rotating head will be at rest, and the tape will playback normally with constant speed and constant pitch. If the head assembly is rotated in a direction opposite to the tape travel, the result will be an increase in the relative head gap to the tape speed with an increase in pitch. Rotating the head in the direction of the tape travel, the relative speed of the gap-to-tape speed is decreased and the pitch is *lowered*. As an example, if the rotational speed of the head-gap is increased to twice the gap-to-tape speed (30 ips), the pitch will be *raised* one full octave. To illustrate tempo change, it will be assumed that the pitch, as explained above, has been raised one octave. Now if the speed of the playback is changed to 7.5 ips, the pitch is brought back to normal, while the tempo has been cut in half. This the regulator accomplishes by linking the rotational speed of the head with the speed of the studio recorder capstan in such a manner as to maintain the pitch constant over a wide range of speeds.

To better understand the tempo regulator functions, expansion of the reproduction time takes place when the head is rotating in the opposite direction of the tape travel. When compressing the program material, the head rotates in the direction of the tape travel. The tape speed during reproduction, in relation to the original recording speed of 15 ips, is adjustable by a control on the regulator. For example, with the control at 120 percent, an increase in tape speed by 20 percent results in a reduction of reproducing time to $^{100}/_{120} \times 1\% = 83.33$ percent of the original recording time. A decrease of tape speed to 95 percent represents an increase of reproducing time to $^{100}/_{95} \times 1\% = 105.26$ percent of the original recording time. The tape speed may be varied from approximately 50 to 200 percent of the original recording time. Accurate adjustments to within 0.5 percent of the recording speeds are possible.

In Fig. 2-131C is shown the scale on a regulator unit for setting the speed of the rotating head for a given compression or expansion. The area between 80 to 120 percent is the normal operating band, where the minimum amount of distortion may be expected. Outside of these limits, the distortion will increase, depending largely on the type of recording.

The increase of reproducing time is basically an expansion of the recording time, obtained by repeating the reproduction time of certain individual sections. The decrease of time is obtained by omitting certain sections of the recording, and combining the remaining sections to obtain a continuous reproduction. At the same time, the speed of the recording medium in relation to the playback media (relative speed) is constant. This is achieved by the special motor assembly discussed previously.

Playback of a tape with an expansion ratio of 20 percent means the reproduction of every 5th section of tape is repeated again, as shown in Fig. 2-131D. The arrows above the numerals indicate the repeated sections. Using the compression ratio of 10 percent (Fig. 2-131E), every 10th section is omitted, and the 11th section follows immediately after the 9th section, the 21st after the 19th, etc. Using a compression ratio of 50 percent, every 2nd section is omitted (Fig. 2-131F).

To prevent the occurrence of audible distortion, it is necessary that during expansion no complete sounds are omitted, or during compression, no complete sounds are repeated. The additions or deletions are on the order of 30 milliseconds, and are always shorter than the shortest sound, in either music or speech. In use, the regulator unit may be installed permanently on a studio recorder, or set on a stand that is parallel with the transport system of the recorder. The output of the rotating head is connected in place of the normal playback head in the studio recorder. The regular unit contains no amplifiers or electronic components, except the special driving motor assembly.

Tempo regulators have been used quite successfully in the motion picture industry for correcting sound tracks where the sound and picture are not synchronized. The sound track is rerecorded at a speed that matches the action. If the out of sync conditions are such that the rate of change varies between action and sound, short sections of the sound track are rerecorded and during the editing are cut in at the proper positions.

Other uses for tempo regulators include training courses, slowed down

Fig. 2-132A. Universal Audio Inc., Model 962 Digital Metronome.

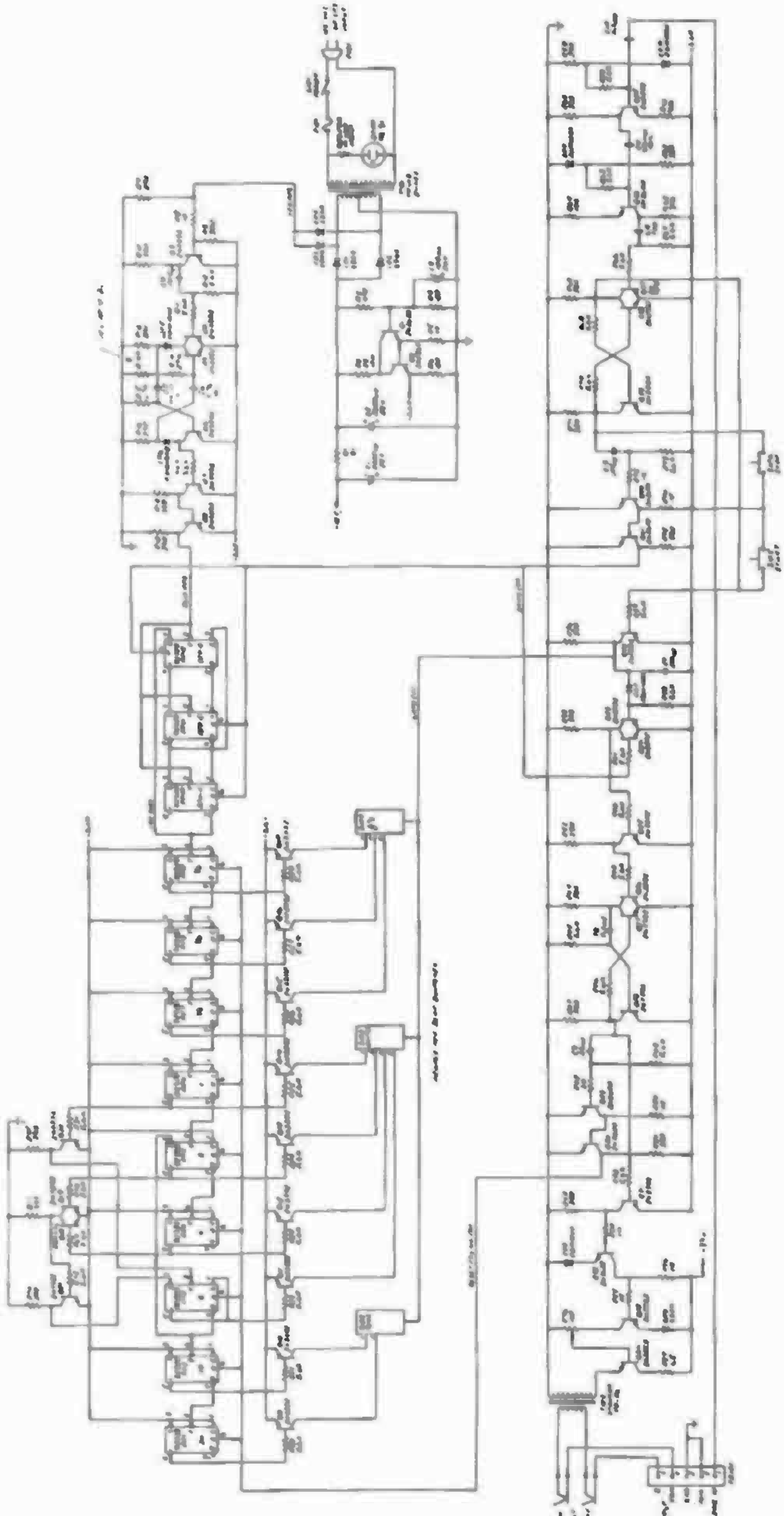

Fig. 2-132B. Schematic diagram for Universal Audio Inc. Model 962 digital metronome.

speech for language study, decreasing the time of commercials for radio, dictation speed reproduction, and many other uses. Speech compression is also possible using computer techniques. For details of this type compression the reader is referred to the appropriate literature.

2.132 ***Outline the principles of metronome design and operation.***—Metronomes are either a spring-driven or motor-driven mechanical device, with a moving pendulum, which produces a tick as it moves back and forth, for beating time. The rate of the beat is controlled by a movable bob that may be raised or lowered on the pendulum, to adjust for a fixed amount of beats per minute.

The Universal Audio Inc., Model 962 digital metronome (Fig. 2-132A) is a precision instrument capable of 312 different tempo beats with maximum deviation (nonaccumulative) on the order of ± 250 microseconds, and is used by motion picture studios for animation, sound effects, and with music scoring. The volume is adjustable, and low output impedance permits the use of a large number of headphones. The beat has a sharp and distinct wavefront, so that it may be easily heard over background noise usually present in film loops. Based on film speeds of 24 fps, tempo beats of 1 frame per beat to 40 frames per beat in ⅛ steps are available by selector switches.

An oscillator, synchronized to the 60-Hz line frequency generates the timing pulses. When the proper count of pulses is reached, an output click is produced, and the counter is reset. Two voltages, —3.6 and —16 Vdc, and a 120-Hz signal for oscillator synchronization, are generated in the power supply. The natural frequency of the oscillator is slightly slower than the required 960 Hz since it is a free-running (astable) *unsymmetrical multivibrator* so designed for ease in synchronization, which is accomplished by the 120 Hz derived from the transformer secondary and diodes CR3 and CR4. The schematic diagram is given in Fig. 2-132B.

2.133 ***How are film spool-banks and continuous film-feed mechanisms constructed?***—A film spool-bank is a group of 16-mm or 35-mm *ball-bearing* plastic rollers used in film processing machines, mounted on a metal pole with

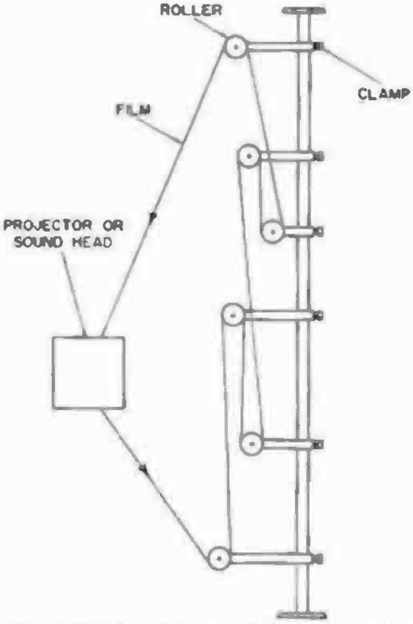

Fig. 2-133A. Film spool-bank with adjustable rollers.

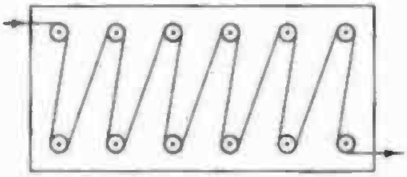

Fig. 2-133B. Film spool-bank using a plywood support.

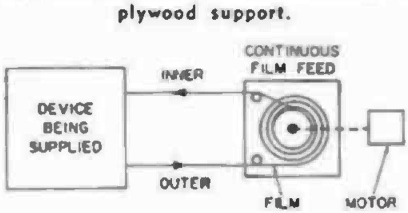

Fig. 2-133C. Continuous film-feed mechanism, for audio-visual devices

clamps for holding and separation of the rollers, as pictured in Fig. 2-133A. A portion of the rollers are mounted on short supporting arms to permit the film loops to clear each other as shown. Using several rollers will permit a rather large amount of film footage to be stored with little wear to the film surfaces.

A second method (Fig. 2-133B) employs a group of rollers mounted on a plywood board. This will also store considerable film. Both these methods of loop storage are used extensively in rerecording installations.

A third method is given in Fig. 2-133C, and consists of a *loosely wound* roll of film mounted on a motor-driven flange. The *inner end* of the film supplies the projector or sound head, while the *outer end* acts as a take-up. The driving-motor speed must be *constant* and compatible with the device being supplied. This system if not cleaned frequently may *abrade* the film surfaces.

The latter system is used quite frequently with audio-visual film display units, while the other two described may be used when looping dialogue and picture. (See Questions 17.208, 17.224 to 17.227, and 17.230.)

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Section 3

Constant-Speed Devices, Motors, and Generators

The search for obtaining a constant linear speed is one that has been going on for years. The paper by Dr. E. W. Kellogg, "A Review of the Quest for Constant Speed," *Journal SMPTE*, April 1937, is a classic on this subject. For the recording and reproduction of sound, many drive systems have been developed, and ingenious new devices are still appearing. This section discusses the fundamental principles of constant-speed driving systems, single- and three-phase interlock systems, power-factor correction, motors, inverters, and general synchronization for both studio and location. Synchronization of camera and ¼-inch tape recorders by means of line frequency sync pulses are discussed as well as several different types of resolving systems.

3.1 What does the term constant speed mean?—A device which moves in a given direction without a change in velocity. The term constant speed is used in the recording industry to indicate devices used for driving recording and reproducing equipment.

3.2 What is the smallest change in the speed of a reproducing device which will be perceptible to a critical listener?—Approximately 0.30 percent at a frequency of 3000 Hz. The average person can detect about one percent.

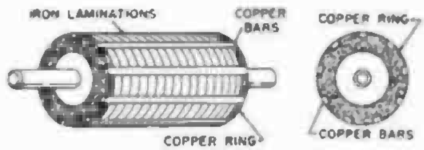
3.3 If the speed of a recording device is increased above normal, what is the effect on reproduction?—When the recorder material is played back at the correct speed, the pitch will be lowered and the playing time increased. The reverse is true if the recording speed is below normal.

3.4 What types of motors are used for driving studio recording and reproducing equipment?—Synchronous, multiduty (dual purpose) and selsyn-interlock motors, which may be used with either single- or three-phase power. The selsyn interlock motors are driven from a selsyn generator.

3.5 What controls the speed of a synchronous motor?—The frequency of the applied power source.

3.6 What is an induction motor?—Of all the motors used for alternating current operation, the induction motor enjoys the greatest use. Induction motors are a form of squirrel-cage motor, the armature consisting of a group of heavy copper bars welded to heavy copper rings at each end of the armature laminations. (See Fig. 3-6A.) The copper bars are short-circuited at each end by the copper rings, thus circulating currents will be induced in them by the field windings. The induced current reacts on the field, causing the armature to turn. The armature does not turn at the same speed as the rotating field, but runs somewhat slower. The difference between these two speeds is called "slip." Thus, if the rotor turns at 1750 rpm and the synchronous speed is 1800 rpm, the slip is 2.78 percent.

To understand how an induction motor functions, the action of the rotating field must be understood. Fig. 3-6B shows the field construction of a typical induction motor having six field poles. Each pole and its opposite member are energized by one of the 3-phase windings labeled phase A, B, and C. It will be noted that Pole 1 is connected in series with Pole 4, Pole 2 is in series



(a) *Side view.* (b) *End view.*

Fig. 3-6A. Rotor for a squirrel-cage motor.

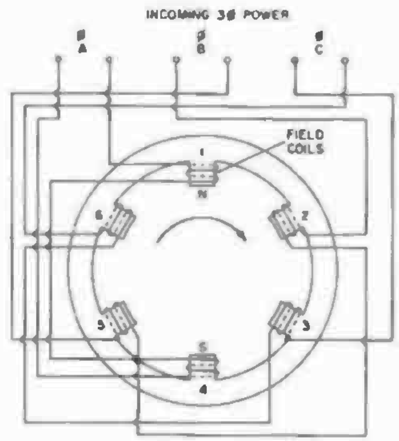


Fig. 3-6B. Field-coil connection of a three-phase induction motor.

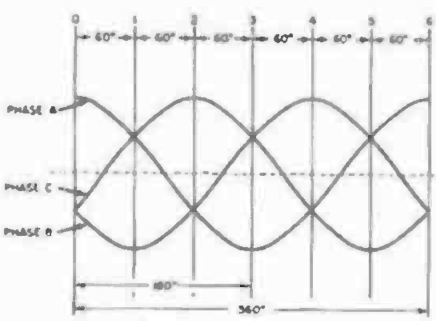


Fig. 3-6C. Rotating magnetic field of a three-phase induction motor.

with Pole 5, and Pole 3 is in series with Pole 6. The voltages applied to these poles are of equal magnitude but differ in phase, as shown in Fig. 3-6C.

Following the voltages as they are applied to the pole coils for a given instant, at zero degrees the magnetic field produced by the three voltages is at a maximum and extends from Poles 1 to 4. Under these conditions, it will be assumed that Pole 1 is North and Pole 4 South. At Position 1, or 60 degrees, the magnetic field has its greatest intensity extending from Pole 2 to Pole 5. At this instant, Pole 2 is North and Pole 5 South. At 120 degrees of rotation, at Position 2, the magnetic field is at maximum from Pole 3 to Pole 6 and it is apparent the field is rotating clockwise. At Position 3 or 180 degrees, Poles 4 and 1 are North and South, respectively, and the field has continued to rotate. This action is continued for each cycle. If the field current has a frequency of 60 Hz, the magnetic field rotates 60 times per second, or 3600 rpm. This is called the synchronous speed. However, the rotor does not rotate exactly at synchronous speed but turns slightly slower, inducing slip. Slip is discussed in Question 3.12.

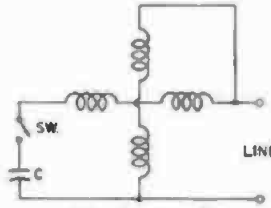
3.7 ***What controls the speed of an induction motor?***—The voltage of the applied power; however, induction motors are not absolute constant-speed devices and should not be used when a constant speed is essential.

3.8 ***What determines the speed of a direct-current motor, series connected?*** —The load and the applied voltage.

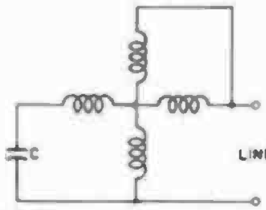
3.9 ***What is a capacitor-start motor and how does it function?***—A capacitor-start motor is similar in its construction to an induction motor, except for an added starting winding, a capacitor, and a centrifugally operated switch. The motor is started with the capacitor and the start winding in the circuit, which causes a phase shift between the windings, thus creating torque. When the motor obtains approximately 80 percent of its rated speed, the centrifugally operated switch opens and cuts out the capacitor and start winding. Typical circuits are shown in Fig. 3-9.

In some types of motors, the capacitor is left in the circuit after starting, as in (b) of Fig. 3-9. Others use a large capacitance when starting, then reduce its value, as in (c) of Fig. 3-9. A fourth method, part (d) Fig. 3-9, employs a transformer which transforms a small capacitor for starting, then either drops out of the circuit or is left in, depending on the service required of the motor. (See Question 8.34.)

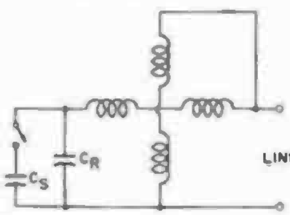
Capacitor-start motors are used quite extensively in magnetic tape recorders, record turntables, and similar devices. As a rule, the centrifugally operated switch is not used with sound equipment because there is a possibility of inducing noise when the switch is operated.



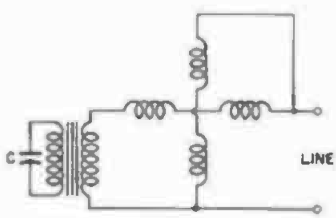
(a) *Start only.*



(B) *Permanently connected.*



(C) *Large capacitor start, small run.*



(D) *Transformer coupled.*

Fig. 3-9. Capacitor-start motor circuits.

3.10 How can an induction motor be altered to operate as a synchronous motor?—By cutting slots in the armature which are equal in number to the pole pieces, or a multiple thereof, and parallel to the shaft.

3.11 What is an induction generator?—A generator designed similar to an induction motor. To generate, it must be driven above synchronous speed and excited from the same ac supply system to which it is to deliver power.

3.12 What is slip in a synchronous motor?—The loss in rpm of a motor when the load is applied. Upon excitation of the primary windings, a rotating field is set up, causing the rotor to rotate. This is called synchronous speed. When the load is applied, the speed drops. The difference in the two speeds is called the slip and is stated in percentage of the synchronous speed. Thus, if the rotor turns 1750 rpm and the synchronous speed is 1800 rpm, the slip is equal to 2.78 percent.

3.13 How is the frequency of an alternator calculated?

$$F = P\left(\frac{S}{120}\right) \quad \text{or} \quad P = F\left(\frac{120}{S}\right)$$

where,

P is the number of poles,
S is the speed in rpm the machine is rotating.

The figure 120 is a constant for converting alternations per minute to hertz.

3.14 What is power-factor correction?—When a large number of electric motors are connected to a power source, the power factor drops. To correct this condition, capacitors are connected in parallel with the power source. (See Questions 3.58, 3.59, and 3.60.)

3.15 What is a shaded-pole motor?—A motor having a copper ring around a section of the pole piece in the direction of rotation, as shown in Fig. 3-15. This coil is called a shading coil and causes a phase difference between the flux from the larger portion of the pole piece and the flux emanating from the smaller portion of the pole piece. This produces a two-phase action in the armature and is sufficient to start the armature rotating.

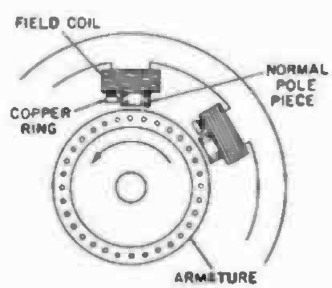


Fig. 3-15. A shaded-pole motor with a single-turn copper coil on a pole-piece.

3.16 What is torque?—A twisting motion in the operation of a machine. Torque is the product of force multiplied by the distance from the center of rotation. For example, a force of 20 pounds applied to the end of a two-foot pipe wrench would be equivalent to 40 foot pounds of torque. In motors, two types of torque must be considered, starting and pull-out. Excessive starting torque results in low efficiency, low power factor, and poor speed regulation. If the starting torque is low, the load cannot be started. High pull-out or stalling torque also means low power factor, with high starting current. If the pull-out torque is low, the motor may stall on small loads.

3.17 *What is a generator?*—A device for converting mechanical energy into electrical energy. The electrical energy may be either direct or alternating current.

3.18 *What is an alternator?*—An alternating-current generator.

3.19 *What is a dynamotor?*—A combination motor and generator, each having its own winding. Very often the motor section is of low-voltage direct current design and the generator section is of high-voltage direct current design.

3.20 *What is a dynamo?*—A generator.

3.21 *What is a genemotor?*—It is similar to a dynamotor.

3.22 *What is a rotary converter?*—A motor-generator consisting of a dc motor direct-coupled to an ac generator mounted on a common base. The motor is battery driven from a storage battery of 6, 12, or 32 volts. The ac generator section consists of a 250- or 500-watt, 115-volt unit. Both the dc and ac sides are filtered to prevent the brushes on the dc side and the slip rings on the ac side from transmitting noise to the equipment being driven by the converter.

A typical machine of this type, manufactured by the Carter Motor Co., is pictured in Fig. 3-22. It is especially designed for driving sound equipment, and will handle loads up to 250 watts having a power factor of 70 percent. The frequency of the ac voltage is controlled manually by a field rheostat in the case above the machine, and indicated on a vibrating-reed type frequency meter.

Fig. 3-22. Carter Motor Co., Model BR-1021CP converter, with manual frequency control and frequency meter.

Rotary converters are also manufactured using a common winding and armature for both the dc and ac sides of the machine. Such machines are not satisfactory for use with sound equipment because the common connection between the dc and ac sides of the circuit results in the transmission of brush noise to the sound equipment.

3.23 *What is an inverter?*—A device for changing direct current to alternating current. This is generally accomplished by means of vacuum tubes. (See Question 21.107.)

3.24 *What does the term electrical degrees mean when associated with a motor or generator?*—Electrical degrees are equivalent to mechanical degrees times the number of pairs of poles in the machine.

3.25 *What is ambient temperature?*—The temperature of the surrounding air or medium in which a device is operating. Motors are rated for a given rise in temperature. (See Question 25.136.)

3.26 *What is a delta connection?*—A winding connected in the form of the Greek letter delta, as shown in Fig. 3-26.

3.27 *What is a wye connection?*—A winding connected in the form of the letter Y, as shown in Fig. 3-27.

3.28 *What is a star connection?*—Windings connected as shown in Fig. 3-27.

3.29 *How may the direction of rotation of a three-phase motor be reversed?*—By reversing any two wires of the 3-phase power source.

3.30 *How may a single-phase motor be operated from a three-phase source?*

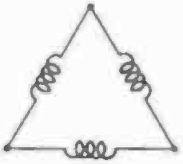


Fig. 3-26. Delta connection for motors, generators, and transformers.

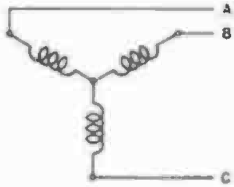


Fig. 3-27. Wye- or star-connected motor.

—By connecting a transformer of the correct voltage ratio between the motor and one pair of the 3-phase power source leads, as shown in Fig. 3-30. It is common practice in the motion picture industry to operate single-phase motors in conjunction with 3-phase motors using the described method.

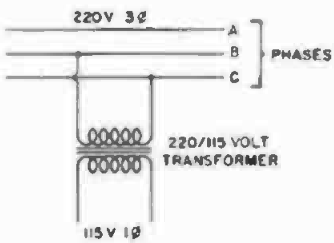


Fig. 3-30. Transformer connections for single-phase operation from a three-phase source.

3.31 ***Can a single-phase synchronous motor be operated in conjunction with a three-phase motor system and still maintain synchronization?***—Yes, provided it is operated from the same power source. Many times sound equipment using single-phase synchronous motors is operated with cameras driven by three-phase motors, during the shooting of motion pictures. The above system is only used when shooting the original picture and sound tracks. For rerecording, a selsyn interlock system is used. Fig. 3-30 shows how the single-phase motor is connected to the 3-phase power source. If the 3-phase and single-phase power are fed from the same source, the single-phase motor may be fed from any single-phase source of power.

3.32 ***Will a three-phase camera motor operate satisfactorily on either a three- or four-wire power system?***—Yes, provided the power characteristics of the motor are such that it will develop sufficient power when operating on the four-wire system. Three-phase, three-wire systems are generally of the order of 230 volts between phases. Four-wire systems are approximately 208 volts between phases. This means that when the motor is operating on a four-wire system, the power developed by the motor is considerably less than normal when operating from a three-wire system. A motion picture camera when cold requires considerably more power to operate; therefore, tests should be made to determine whether the camera can be brought up to speed before using it on production.

Generally, it is necessary to rewind a 3-phase motor designed for 230-volt operation, if it is to be used on a four-wire system. A four-wire, 3-phase, 208-volt power distribution system is shown in Fig. 3-32A. The voltage between any two phases at the load side is 208 volts. 120-volt single-phase power may be taken between ground and any one of the three phases.

The characteristics of a 220-volt, three-phase camera motor operating from a power source of 208 volts is shown in Fig. 3-32B, part (a). It will be noted that as the voltage is reduced below 208 volts, the motor speed starts to drop off because of the lack of power, which reduces the frames per second at a voltage of 200 volts. Part (b) is the power curve after rewinding the motor to operate at 208 volts. Here it may be seen the motor runs at its normal speed down to 200 volts and maintaining a speed of 24 frames per second.

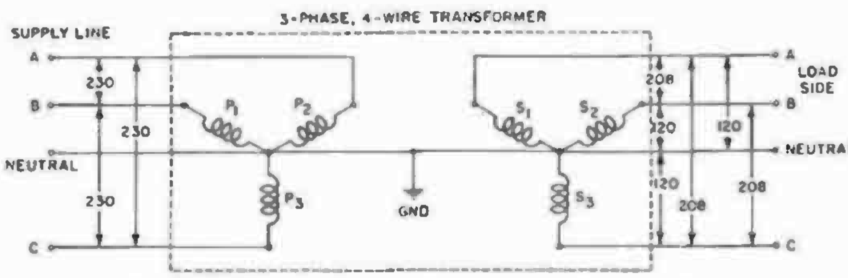
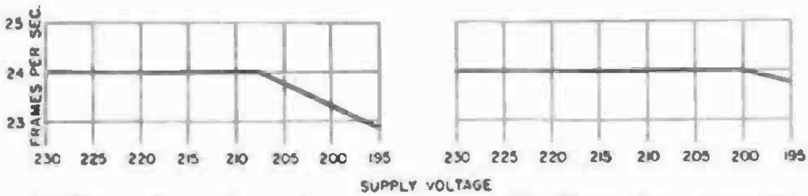


Fig. 3-32A. A three-phase, four-wire distribution system.



(a) *Three-phase, three-wire.* (b) *Three-phase, four-wire.*

Fig. 3-32B. Supply voltage versus sync drop-out of a three-phase synchronous camera motor.

The curves are actual curves made on a motor that is standard for 35-mm cameras and were made after running 3000 feet of film through the camera to warm up the motor. (See Question 8.22.)

3.33 ***What is an asynchronous motor?***—One in which the speed is not proportional to the frequency of the power source.

3.34 ***How many electrical watts are equal to one horsepower?***—746 watts.

3.35 ***How is the speed of a single-phase camera motor affected when the voltage source is lowered?***—Fig. 3-35 shows a test made on a typical single-phase camera motor plotted as line voltage versus frames per second for both warm and cold conditions of the motor. It will be observed that when the motor is warm the voltage may be dropped from 120 volts to 114 volts before the frames per second start to fall off. The condition is considerably worse when the motor is cold, as shown by the solid line.

This clearly indicates that the motor must be warmed up before using it, or that the voltage source must be maintained sufficiently high to maintain the correct speed. This may be accomplished by the use of a continuously variable autotransformer. (See Question 8.8.)

3.36 ***What is a prony brake?***—A friction device used for measuring the brake horsepower of a motor. The brake consists of an arm (Fig. 3-36) and a friction band clamped around a pulley mounted on the end of the motor shaft. The tension of the friction band may be adjusted by the wing nuts at the top of the band. A spring balance scale is connected to the end of the arm and secured overhead. The motor, in rotating, causes the friction band to move the arm downward and pull on the scale which is calibrated in either ounces or pounds. Knowing the speed of the motor, the length of the arm from the center of the motor shaft, and the pull on the scale, the brake horsepower may be calculated as follows:

$$B_{HP} = \frac{2\pi L \times rpm \times lbs}{33{,}000}$$

where,

L = Length of the brake arm in feet,
rpm = speed of motor,
lbs = the pull on scales, in lbs.

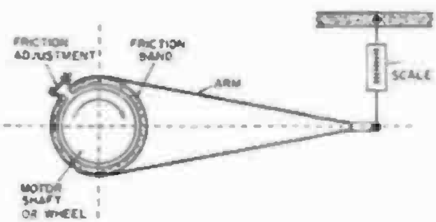


Fig. 3-36. Prony brake for measuring brake horsepower of a motor.

3.37 ***What is the formula for calculating the required size capacitor for a capacitor-start motor?***

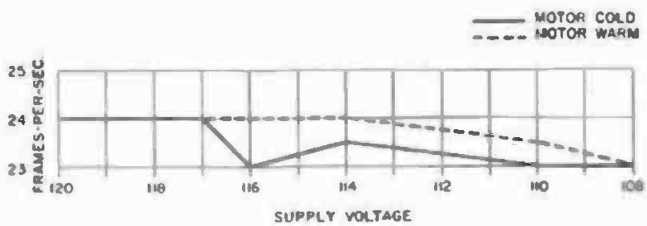


Fig. 3-35. Frames-per-second versus line voltage for a 120-volt synchronous, single-phase camera motor.

$$\mu F = \frac{2650 \times \text{amperes}}{\text{applied voltage}}$$

where,

amperes is the current drawn by the motor,
2650 is a constant,
the applied voltage is equal to the line voltage.

3.38 How is the efficiency of a motor calculated using a prony brake?

$$\text{Efficiency} = \frac{B_{HP}}{746}$$

where,

B_{HP} = brake horsepower,
746 = one horsepower in watts.

3.39 How can the torque of a motor be calculated using a prony brake?

$$\text{Torque (in ounce-inches)} = \frac{1352 \times \text{watts}}{\text{rpm}}$$

where,

watts is the power consumed by the motor,
rpm is the speed of the motor shaft,
1352 is a constant.

3.40 How is torque (in ounce-inches) converted to watts?

$$\text{Watts} = \frac{\text{oz-in} \times \text{rpm}}{1352}$$

3.41 How is torque (in ounce-inches) converted to horsepower?

$$\text{Hp} = \frac{\text{oz-in} \times \text{rpm}}{1{,}008{,}000}$$

3.42 State the equation for calculating the horsepower of a motor in watts.—The relationship of electrical horsepower to watts may be defined as follows:

One electrical watt is equivalent to one joule per second, or 60 joules per minute. Since one joule is equivalent to 0.7374 ft-lb, 60 joules will equal 44.244 ft-lb. Also, one horsepower is equivalent to 33,000 ft-lb per min. Therefore, the electrical equivalent of one horsepower is:

$$\frac{33{,}000}{44.244} \text{ or 746 electrical watts.}$$

The horsepower of a fractional horsepower motor may be expressed as a function of the torque and the revolutions per minute. For fractional horsepower motors, the approximate formula is:

$$\text{Hp} = (\text{in-oz}) \times 9.3 \times N \times 10^{-7}$$

where,

in-oz, is the torque in ounce-inches,
N is revolutions per minute.

3.43 How is the efficiency of an electric motor calculated?

$$\text{Efficiency} = \frac{\text{watts output}}{\text{watts input}}$$

3.44 How is synchronous speed calculated?

$$\text{Sync speed} = \frac{120 \times \text{line frequency}}{\text{number of poles}}$$

3.45 Define the term ounce-inches.—It is the weight in ounces that a motor will lift at a distance of one inch from the centerline of the motor shaft.

3.46 How can commutator noise be eliminated?—By the use of capacitors and chokes, as shown in Fig. 3-46. The most satisfactory method is by cut-and-try, as each source of noise presents an individual problem. As a rule, chokes will not be required unless radio-frequency interference is noted. The frequency of the interference will determine the value of the inductance. For large dc generators supplying several hundred amperes, such as used on motion picture sets with arc lights, a bank of 25 to 50,000 μF electrolytic capacitors is connected directly across the generator brushes. The capacitors are housed in a metal box, grounded to the frame of the generator. No smaller than a No. 10 gauge wire should be used for the capacitor connections. The voltage rating of the capacitors should be 25 to 50 volts greater than the maximum output voltage of the generator. (See Question 25.117.)

3.47 What is an electrical gearing system?—A name sometimes applied to a selsyn-interlock distributor system used for synchronous operation of recording equipment. (See Question 3.49.)

3.48 What is a selsyn-interlock distributor system?—A synchronous distributor system used for interlocking sound recording equipment with motion

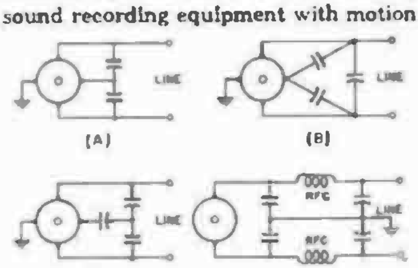


Fig. 3-46. Methods of filtering commutator ripple. The capacitor sizes from A to C vary 0.10 to 0.50 microfarad.

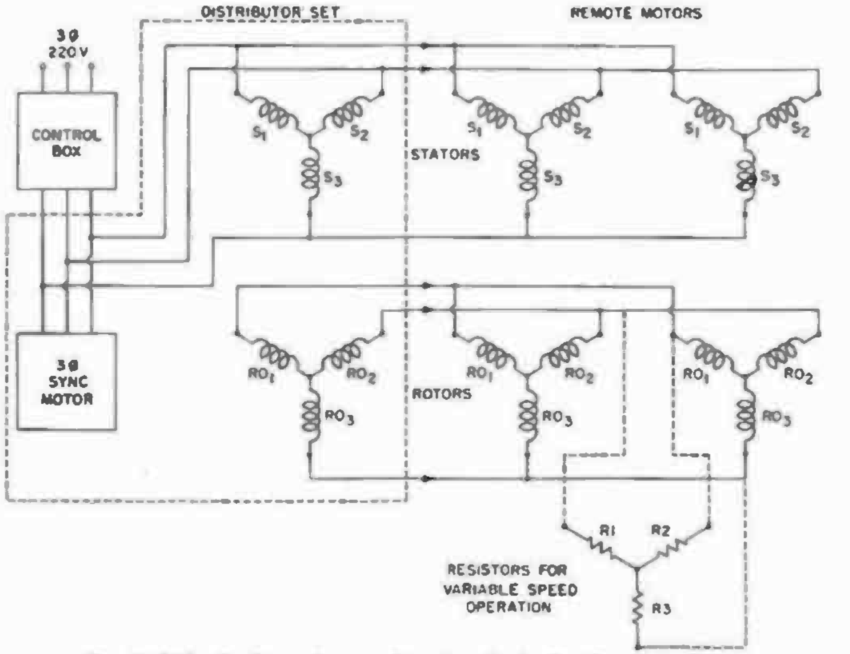


Fig. 3-49A. A three-phase selsyn-interlock distributor system.

picture cameras. Selsyn distributor systems are described in Question 3.49.

3.49 ***How does a three-phase selsyn-interlock distributor system function?***—The diagram for such a device appears in Fig. 3-49A. Starting at the left is a control box consisting of a group of relays, resistors, and three solenoids, one connected in each phase of the incoming power source. The solenoids have normally a low impedance; however, at the instant current flows through the coil, a soft iron plunger is pulled upward into the coil increasing its impedance. As the driving motor starts to rotate, the initial current drops and, as it does, the plunger falls out of the coil lowering its impedance to the original value. The pulling up of the plungers into the starting coils prevents the system from making a sudden start which might break film or damage equipment. When adjusted properly, a slow smooth start is obtained, even with 10 to 20 machines being driven simultaneously. The driving motor is a 3-phase, 220-volt, synchronous type turning 1200 rpm and is mechanically coupled

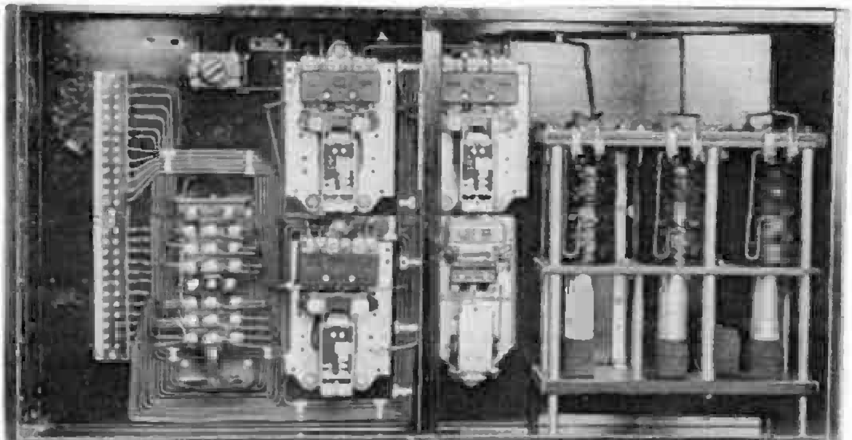


Fig. 3-49B. Control cabinet RCA selsyn distributor unit.

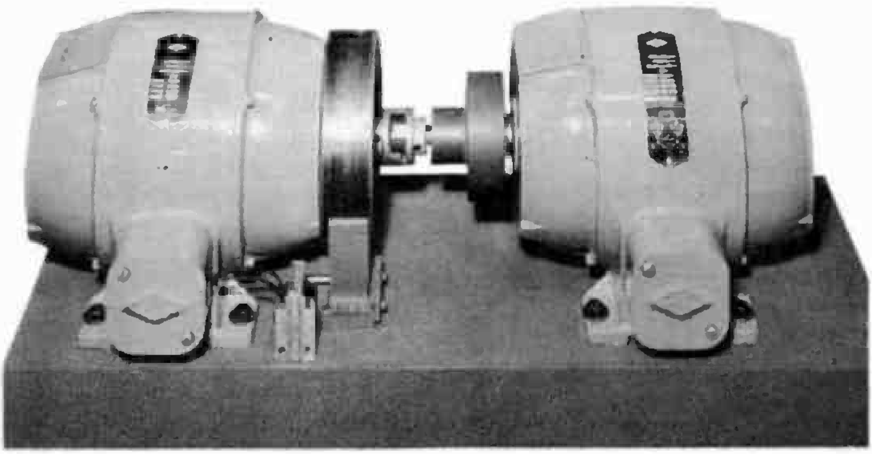


Fig. 3-49C. Selsyn interlock distributor set manufactured by Stancil-Hoffman. The electrically controlled brake may be seen at the front of the flywheel.

to the selsyn distributor unit. A heavy flywheel is mounted on the shaft to provide a large amount of inertia for a soft start and gradual slowdown. When running, it also helps to iron out small irregularities in the speed. An electrically controlled brake may be applied to the outer rim of the flywheel to bring the system to a gentle stop when the circuit to the drive motor is opened. When the circuit for running is closed, the brake is lifted automatically.

The sole purpose of an interlock system for rerecording and other purposes is that all the machines in the system, including the projection machine (which is the heaviest to start), must be brought up to speed in synchronism, from a standing start, in about 6 seconds. This must be accomplished in a smooth manner to avoid breaking film and sound track.

Selsyn distributors are sometimes referred to as rotary transformers, because of the 1:1 ratio of the two windings. As a rule, the internal windings of the distributor unit are star-connected. It will be noted the rotor windings of the distributor unit are connected in parallel, but not to the power source. The stator winding of the distributor and all other units being driven are also connected in parallel, but in this instance they are connected to the 3-phase power. Phasing of each unit is accomplished by connecting similar numbers of the windings together and to the same phase.

Because the distributor unit is always the largest unit in the system, it becomes the master and all other units become slaves. When properly phased, absolute synchronism is maintained as long as the frequency of the power source remains constant. Under these conditions the system may be termed as a constant-speed device.

In some distributor systems, the three solenoids shown in the control cabinet (Fig. 3-49B) are replaced by a group of resistors connected in the supply lines to the synchronous drive motor. The voltage drop across the resistors during the high inrush of current when the motor is at rest, holds the cutout relay open. When the system approaches its normal running speed the current through the resistors drops, and the relay closes, cutting out the resistors. A typical selsyn interlock distributor set is shown in Fig. 3-49C.

3.50 How may the selsyn distributor unit in Question 3.49 be made to operate at variable speed?—At times during the rerecording of a motion picture, a variable-speed distributor system is required to obtain a particular sound effect. If a separate distributor system is available, it may be made to run at variable speed by connecting three variable resistors, R1, R2, and R3, across the rotor windings as shown in Fig. 3-49A. If the values of the three resistors are varied the same amount and simultaneously, the speed of the system may be controlled. If the three resistors are shorted, the machine will run at 1800 rpm.

3.51 Is it possible to synchronize camera and sound with direct current

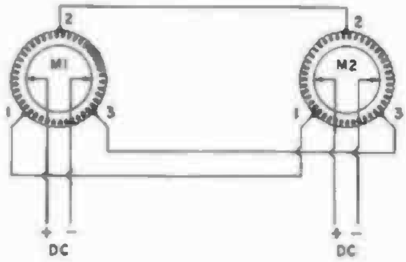


Fig. 3-51. A dc synchronous drive system for sound recorder and camera. The motor speed is set by means of a tachometer.

motors?—Yes. Special motors are available for this purpose. The connections for such a synchronizing system are shown in Fig. 3-51. It will be noted that the dc windings of the motors are tapped in three different places, 120 degrees apart. These taps are carried to each motor to be synchronized. Local batteries at each motor supply the necessary power for rotating the motors. A tachometer is used to set the initial speed.

3.52 ***How is frequency control obtained in other types of distributor systems?***—Other systems are available, similar to the one described in Question 3.49, except that vacuum-tube control circuits have been added to compensate for changes in the power frequency. Because present day power systems are controlled to hold the frequency within a fraction of one cycle, the system described in Question 3.49 is used in preference to the electronically controlled type.

3.53 ***Can single-phase selsyn motors be used to drive sound equipment?***—Yes, except that single-phase systems are not as efficient as 3-phase systems and are more subject to slipping out of lock. A diagram of a typical single-phase, 115-volt, ac interlock system is shown in Fig. 3-53. As a rule, single-phase selsyn motors are internally connected in delta. It will be noted the rotors have only one winding and are connected to the source of power and driven by a synchronous motor mechanically connected to the shaft.

3.54 ***Describe a simple interlock system.***—A simple but effective single three-phase interlock system can be devised by mounting an interlock motor on a motion picture projector and driving the interlock motor from the synchronous motor on the projector by means of a Gilmer belt and gears. Although this system is not suitable for rerecording, it may be used for running picture and sound track in synchronism. As a rule, the motor on the projector is around ⅙ to ⅓ horsepower, so no difficulties should be encountered. However, for rerecording, a selsyn interlock system is required. (See Question 3.49.)

3.55 ***Describe an interlock distributor system, using a differential generator for correcting out of phase conditions.***—At times, during the screening of a picture it will be discovered that the picture and sound are out of sync, because of improper splicing or threading of a machine. This necessitates stopping the system and correcting the fault and re-

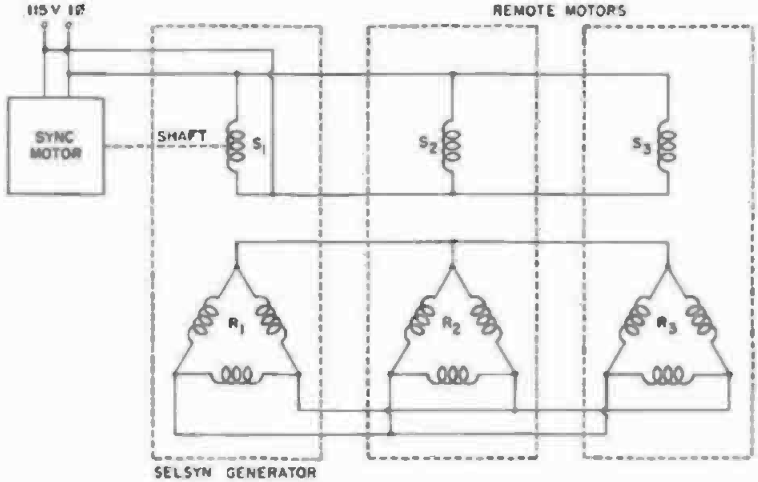


Fig. 3-53. A single-phase selsyn-interlock distributor system.

threading all machines. If several sound tracks are involved, this can become quite costly, plus the time lost. One of the major studios provided for this situation in its review rooms by installing differential generators in the interlock system feeding three sound reproducers and two projectors. If it becomes necessary to correct for an out of sync condition, the machine at fault,

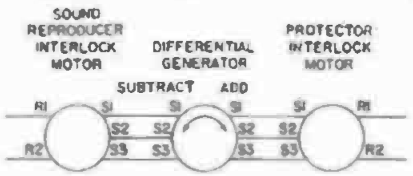


Fig. 3-55. Differential generator connected in the line feeding a sound reproducer and projection machine, using single-phase motors.

while running, can be advanced or retarded to correct the number of frames to bring it back into sync. The control dials for the differential generators are calibrated in frames per second to aid the editorial department in correcting the situation later. A footage counter indicates the exact footage of the out of sync condition.

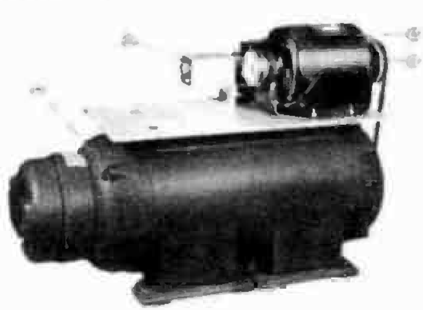


Fig. 3-56. Dual-speed selsyn generator made by Magna-Tech Electronic Co.

3.56 ***Describe a dual-speed selsyn-interlock distributor system.***—A dual-speed selsyn-interlock distributor system generator, manufactured by Magna-Tech Electronic Co., is shown in Fig. 3-56. This device consists of two selsyn generators. The smaller of the two (A) is mounted above the larger generator (B). The larger generator is driven by a 208/230-volt three-phase, 2½ horsepower synchronous speed motor at 1200 rpm. This generator is used to interlock and drive 35-mm machines running at 90 fpm (1200 rpm).

The smaller generator is driven from the larger generator by a Gilmer belt and gears (C), and is employed for interlocking and driving 16-mm machines at a speed of 36 fpm (600 rpm). A third generator (D) is provided for driving a remote footage counter in the dubbing stage. (See Question 18.339.) The larger generator also houses a braking system for bringing the system to a smooth stop. External cabinets house the control equipment.

If desired, the 16-mm generator (B) may be substituted for one turning at 45 fpm for driving machines using 17.5-mm magnetic film. Dual-interlock systems are very convenient in plants where both 16- and 35-mm recording equipment is in use, as this permits a 35-mm picture to be run with 16-mm recording and reproducing equipment or vice versa. Distributor systems of this design are often referred to as a *piggy-back* distributor system.

3.57 ***What is a pm generator?***—An alternator which uses permanent magnets rather than coils in the rotating member. In this respect, it differs from the conventional type alternator. The rotor consists of a nonmagnetic ring in which Alnico V magnets are embedded and equally spaced in such a manner that the magnetic axis of each magnet is perpendicular of the axis of the rotor shaft. The inner end of the magnet rests on an iron ring which acts as a return magnetic circuit. The number of magnets will depend on the speed and frequency of the machine. This type of construction eliminates arcing caused by slip rings.

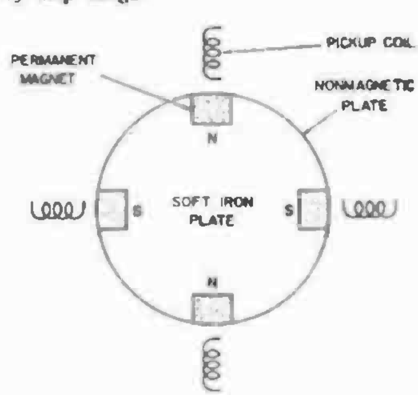


Fig. 3-57. Basic design of a permanent-magnet generator. Such design is used for sync-pulse generators mounted on the end of a camera motor shaft.

Small generators based on the above design are often used for generating sync pulses for one quarter inch magnetic recorders. Two types are in use, two-pole and four-pole. The construction of the rotating member is shown in Fig. 3-57. The number of magnets will depend on the desired frequency and the rotational speed of the motor shaft. The frequency used for sync-pulse recording is 60 Hz; however, in some instances where the generator is mounted on a camera shaft, turning at 1440 rpm, the frequency is 48 Hz. The frequency may be computed:

$$f = \text{rpm}/60 \times \text{the number of poles.}$$

Thus, for a motor turning at 1800 rpm, using two-poles:

$$f = 1800/60 = 30 \times 2 = 60 \text{ Hz.}$$

3.58 ***What is power factor and how is it measured?***—The power factor of an alternating current is the number by which the apparent power in the circuit (volts times current) must be multiplied to ascertain the true power. When an alternating-current circuit contains inductance, the current will lag behind the voltage. When the circuit contains capacitance, the current will lead the voltage. In each instance, the current and the voltage reach their maximum values at different instants and the product of the voltage and the current at any given time is less than it would be if the two were in phase. If the voltage and current are measured separately, the voltmeter and the ammeter will indicate the mean effective values. If the power in the circuit is measured using a wattmeter, the instrument indicates the combined efforts of the voltage and current synchronously, not the product of their effective values, which occur at different instants. Consequently, a wattmeter indication will be less than the product of separate voltmeter and ammeter readings.

The ratio of the power read by the wattmeter and the power read by the voltmeter and ammeter is the power factor of the circuit.

$$1\,\phi\,\text{PF} = \frac{\text{watts}}{\text{V} \times \text{A}}$$

where,

watts is the reading of the wattmeter,
volts and amperes are the readings on the individual meters.

The power factor for a two-phase circuit is expressed:

$$2\,\phi\,\text{PF} = \frac{\text{watts}}{2(\text{V} \times \text{A})}$$

and for a 3-phase circuit:

$$3\,\phi\,\text{PF} = \frac{\text{watts}}{\sqrt{3} \times \text{line to line volts} \times \text{line amps}}$$

A single-phase motor drawing 5 amperes at 220 volts, as shown by the voltmeter and ammeter, has a power factor of 80 percent. The true power is $5 \times 220 \times 0.80$, or 880 watts.

3.59 ***How is a synchronous motor used for power-factor correction?***—If the synchronous motor is driven from an external source of power and a variable dc voltage is applied to the rotor windings and set to a value which will be called, for the sake of illustration, 100 percent, no current will flow from the stator windings to the rotor windings. Under these conditions the voltage generated in the stator windings (counter emf) exactly balances the voltage applied to the stator from the external voltage source.

If the dc exciter voltage is now reduced to a value less than 100 percent, a reactive component is produced which will lag the applied voltage. The machine will now act as a capacitor. Thus, a synchronous motor may be used for correcting the power factor and, when so used, is called a synchronous capacitor or a rotary condenser.

When the motor is used as a capacitor, the motor is connected in parallel with the line to be corrected, and the dc excitation voltage is adjusted to produce a leading current which offsets the lagging line current. The result is unity power factor.

3.60 ***How may the power factor be corrected by the use of capacitors across the line, and how are their values calculated?***—Power factor may also be corrected by connecting capacitors in parallel with the line to be corrected. The approximate value of capacitance required may be calculated:

$$C_{\mu F} = \frac{I \times \text{Sin}\,\phi \times 10^6}{2\pi f E}$$

where,

E is the line voltage,
I is the current drawn by the load,
f is the frequency of the power source.

The value of the capacitance required for power-factor correction may also be determined experimentally by connecting a dynamometer-type voltmeter across the line and adding sufficient capacitance in parallel with the line until the voltage is brought up to normal.

If the value of capacitance is large, ac electrolytic capacitors may be used. Dc electrolytic capacitors must never be used as they will explode within a few minutes of being connected across the line. If practical, paper or oil-filled capacitors should be employed.

3.61 How can the power factor be estimated if measurements cannot be taken?—For regular lighting loads and no motors, 0.95; with lighting and motors, 0.85; and for motors only 0.80.

3.62 How may the current-per-phase of an electrical system be calculated?—For a single-phase system:

$$I = \frac{W}{E \times PF}.$$

For a two-phase system:

$$I = \frac{W}{E \times PF} \times 0.5.$$

For a three-phase system:

$$I = \frac{W}{E \times PF} \times 0.58.$$

where,

I is the line current,
W is the power delivered in watts,
E is the potential existing between the mains,
PF is the power factor.

3.63 What is the procedure for testing motor-starting capacitors?—For the proper starting and operation of a motor using a capacitor permanently or for starting only, the capacitor must maintain its power factor and capacitance, within fairly close limits. A high power factor is manifest by reduced starting torque and a prolonged starting period. To test a starting capacitor properly, it should be tested under the exact conditions in which it operates. A circuit for this purpose is shown in Fig. 3-63A. The power-factor wattmeter must be capable of reading quite low values, since the losses of a good capacitor are quite low, although the product of current and voltage as read on the voltmeter and ammeter is quite high. Therefore, it is desirable to have a power-factor wattmeter that will read about one fifth the watt volt-ampere capacity of the test circuit.

To find the capacitance and power factor of a capacitor, the voltage is adjusted to the rated value of the capacitance, and readings of the voltmeter and ammeter are taken. The capacitance may then be calculated:

$$C = \frac{1 \times 2650}{115}$$

The percent power factor is then found by dividing 100 times the power in watts, as read on the power-factor wattmeter, by the product of the current times the voltage. This is true only for 60 Hz. The problem may be somewhat simplified by the use of the nomograph in Fig. 3-63B.

Using the test circuit shown, assuming the capacitor draws 5.2 amperes, at 120 volts at 60 Hz, and dissipates 40 watts, the percentage power factor is found by

$$\text{Percent PF} = \frac{100 \times \text{Power in watts}}{\text{Current} \times \text{Voltage}}$$

or

$$\frac{100 \times 40}{5.2 \times 120} = 6.4 \text{ Percent}$$

To read this value on the nomograph connect 120 volts from the left of "A" to 5.2 amperes on the left of "C," with a straightedge. Note where the straightedge intersects "B." From this point on "B" connect the straightedge with 40 watts on the left edge of "A" and read

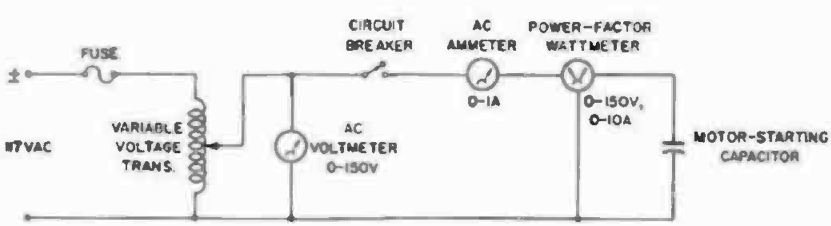
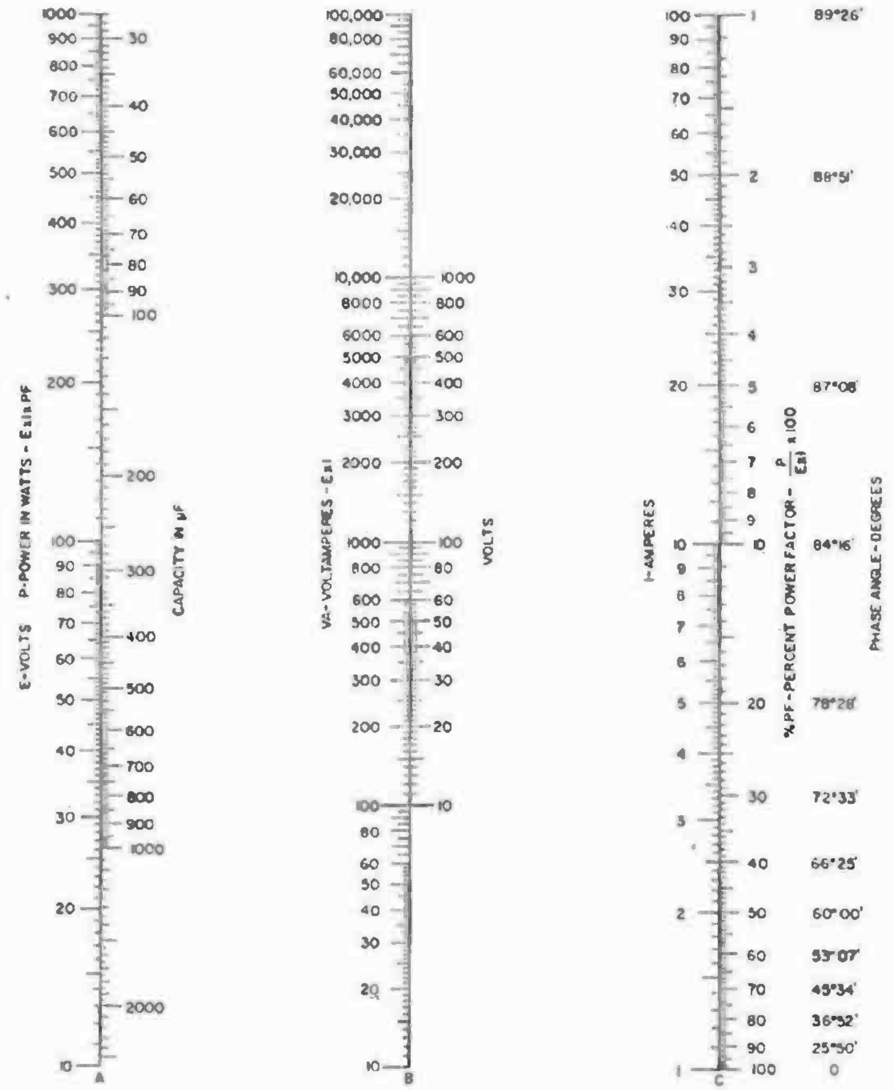


Fig. 3-63A. Circuit for testing motor-starting capacitors.



POWER-POWER FACTOR CHART

Fig. 3-63B. Nomograph for determining the power factor of motor-starting capacitors.

the power factor (6.5 percent) on the right edge of "C."

Capacitance may be determined from the nomograph by placing the straightedge at 5.2 amperes on the left of "C" to connect with 120-volts on the right edge of "B" and reading the capacitance on the right edge of "A."

Motor-starting capacitors may be of three different types, depending on the voltage and the required capacitance. If the required capacitance is small, either oil or paper dielectric may be used. If the needed capacitors are large, (10 microfarads or greater) they are generally of the ac or nonpolarized electrolytic type. The characteristics of motor-starting capacitors must remain stable, that is, they must hold their capacitance and power factor within fairly close limits. If they do not, the motor will have low starting torque and prolonged starting, resulting in the motor and capacitor overheating. *Dc polarized electrolytic capacitors must never be used as motor-starting capacitors.* (See Question 3.60.)

3.64 ***What is meant when it is said a motor-generator is hunting?***—A motor-generator is said to be hunting when it oscillates during the starting period or increases or decreases its speed when the load is changed.

3.65 ***Where are torque motors used?***—Manufacturers of professional (and some nonprofessional) magnetic recording and reproducing equipment make use of torque motors mounted on the spindles for the feed and take-up reels, to provide a constant tension on the film or tape, regardless of the diameter of the recording media on either reel. The motors may be of the induction type, or of the single-phase shaded-pole design. Resistors connected in the ac supply line are used to control the speed, and to provide an adjustment for the desired tension. The power drops off proportionally to the increase in speed, and develops maximum power at stall speeds. Therefore, this characteristic is used to an advantage to control the tension for the feed and take-up reels.

One advantage of using torque motors in the feed and take-up positions is that the tape or film may be rewound in either direction. During the rewind cycle, the circuitry of the motor system is such if the tape breaks or runs out the motors stop automatically. The motor on the feed reel always acts as a holdback. This subject is further discussed in Section 17.

3.66 ***How can the power factor of a dynamotor be corrected to operate with a low power factor motor?***—If the low power factor motor is connected across the ac side of a dynamotor, the dynamotor will speed up. Power correction may be applied by adding capacitance across the power source at the ac generator terminals in parallel with the load. A typical example of how the power factor may be corrected is given below.

A 0.5 kW dynamotor driven by a 12-volt battery, develops 115 volts ac, which is to be used to drive a portable recording channel and supply sufficient power to drive a 35-mm motion picture camera. When the camera was applied to the ac side of the dynamotor, the power factor, as measured with an ac ammeter and voltmeter, indicated a power factor of 32 percent and a power consumption of 190 watts.

Connecting a 20 μF oil-filled capacitor across the ac terminals increased the power factor to 40 percent, with 190 watts of power consumed. Increasing the capacitance to 70 μF increased the power consumption to 280 watts with a power factor of 92.5 percent.

Power factor may be calculated:

$$PF = \frac{W}{AC_v \times AC_i}$$

where,

AC_v is the voltage across the load,
AC_i is the current drawn by the load,
W is the watts consumed.

Capacitance is applied in parallel with the load until a power factor of at least 90 percent is obtained. The proper amount of capacitance is the point where the maximum power factor correction is obtained with a minimum of capacitance.

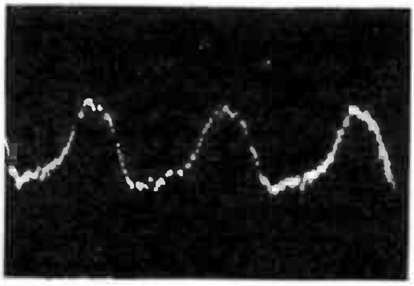


Fig. 3-67A. Waveform of dc generator with shorted turns in armature.

3.67 ***What is the appearance of the current waveform of a maladjusted dc generator?***—Malfunctioning of the generator because of bad brushes and commutator bars or shorted turns will produce a waveform at the ontput of the generator similar to the one shown in Fig. 3-67A. A well-adjusted and operating dc generator will, after the current has passed through a filter to remove commutator ripple, show practically a pure dc current when observed

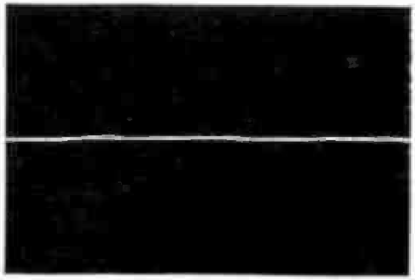


Fig. 3-67B. Waveform of dc voltage from a well adjusted dc generator.

Fig. 3-67C. Waveform of a properly adjusted dc generator before filtering.

on an oscilloscope (Fig. 3-67B). The waveform before passing through the filter is shown in Fig. 3-67C.

3.68 ***How are filter circuits connected in the input and output of a dc dynamotor?***—In the manner shown in Fig. 3-68. Large capacitors are connected in parallel with the battery side to remove the effects of motor commutator ripple and brush noise. An iron-core choke and two filter capacitors are connected to the load side to remove commutator ripple and radio-frequency interference. The capacitors and choke must be placed in a metal can and grounded to the frame of the machine.

3.69 ***How are filters connected in the input and output of a converter?***—In the manner shown in Fig. 3-69. Capacitors C1 and C2 at the ac output remove any noise due to sparking at the slip rings, and the small capacitor C3 at the output terminals removes radio-frequency interference. The frame of the machine should have a physical ground, if possible.

3.70 ***Describe the construction of an inside-out hysteresis synchronous motor.***—One such type motor is shown in Fig. 3-70A. This motor is often referred to as an inside-out motor, because the stator revolves rather than the armature. This type of motor is used extensively for the driving of ¼-inch magnetic-tape recorders, at speeds of 1⅞ to 30 inches per second. Such motors are available in dual- and triple-speed designs. Common speeds for the dual type are; 300/600, 360/720, 450/900, and 600/1200 rpm; and 300/600/900 for the triple speeds. A nonmagnetic puck or capstan is pressed on the motor spindle to obtain the desired linear speed for magnetic tape.

Hysteresis synchronous motors provide the exact constant-speed required for magnetic recorders, with high torque and constant angular velocity. The rotor (stator) is a dynamically balanced flywheel, which assures a constant speed under changing load conditions. A Mumetal shield is placed over the stator winding and rotor, to eliminate the high flux-torque from the region of the motor shaft. Impellers on the rotor circulate air to provide forced ventilation for the motor and associated equipment.

Fig. 3-70B shows the internal and external connections for a dual-speed motor, and Fig. 3-70C shows the connections for the triple-speed unit. The average power consumed hy this type motor is approximately 30 to 40 watts, at 117 volts, with a rotor torque of 7 to 10 ounces. The bearings may be ball-bearing or oilless sleeve type. The motor is single hole mounting, and uses a capacitor ranging from 1.5 to 3.0 μF.

3.71 ***How is speed reduction obtained for recording and reproducing equipment?***—Several methods used to obtain speed reduction in recording and

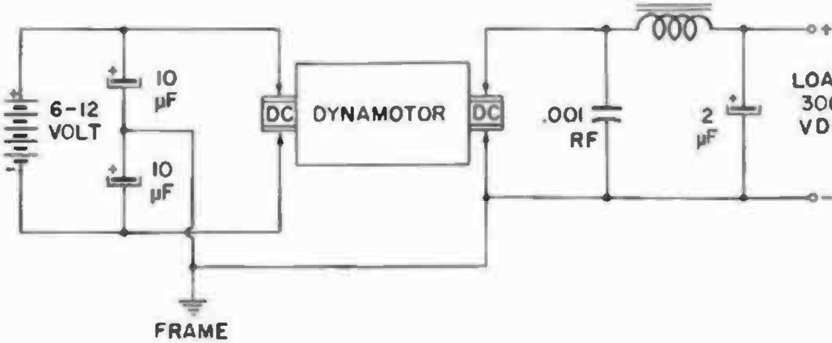


Fig. 3-68. Filter system used with a 6/12 volt dc to 3000 volt dc dynamotor.

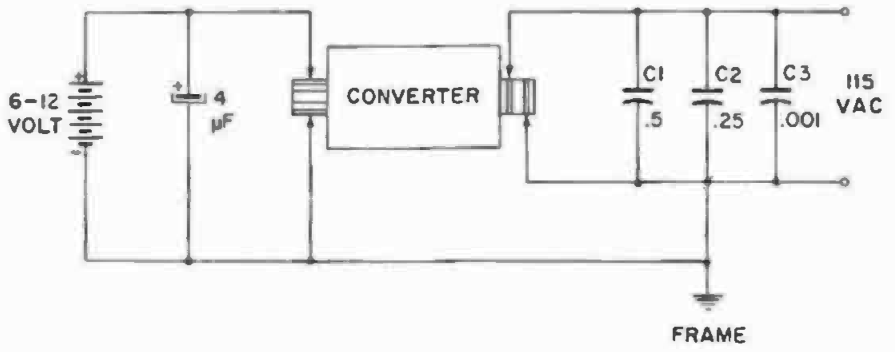


Fig. 3-69. Filter system used with a 6/12 volt dc to 115 volt ac converter.

reproducing equipment are shown in Fig. 3-71. Among these various systems will be found gear reducers, intermediate idler rollers, and belt drives. Fig. 3-71(a) shows three types of puck or idler drive systems. Fig. 3-71(b) illustrates two types of gear reducers, Fig. 3-71(c) shows two types of belt-speed reducers, and Fig. 3-71(d) shows a "Gilmer Timing Belt" drive system. For the sake of clarity in the drawings, decoupling devices (compliance) generally connected between the motor and the driven member have been left out.

Decoupling devices are generally of a loose coupling design; that is, the motor shaft is connected to the driven member through a piece of rubber, felt, leather, or adhesive tape, to reduce the transmission of vibration from the motor to the speed-reduction system. In some systems, a second decoupler is used to isolate the gearbox vibration from the driven system. Pucks, idlers, and belts serve as their own decoupling devices.

Intermediate rubber-covered idler roller drives are nsed in magnetic recorders and other types of recording equipment because of their low cost and smoothness of operation. Also, they serve as decoupling devices and prevent hunting in the drive system. The principal objection to their use is slippage, due to the deformation of the rubber at the point of surface contact. This slippage may be reduced to a minimum by the use of multiple pucks with narrow faces. Wide faces are not always best because the wide contact surface does not bear uniformly over the entire surface.

Rim drive or drives on the outer surface of a flywheel damp out the action the flywheel is intended to impart to the drive system, that of smoothing

Fig. 3-70A. A two-speed hysteresis synchronous motor used for driving magnetic-tape recorders.

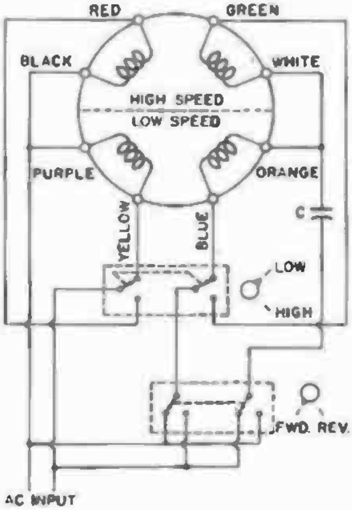


Fig. 3-70B. Connections for the Technical Development Co. dual-speed, hysteresis synchronous motor.

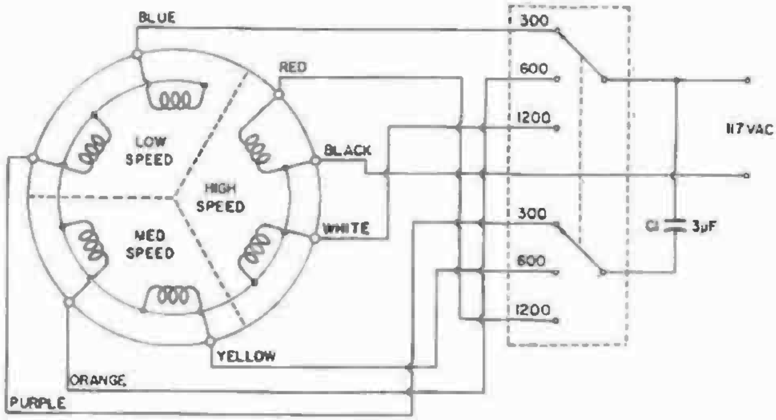
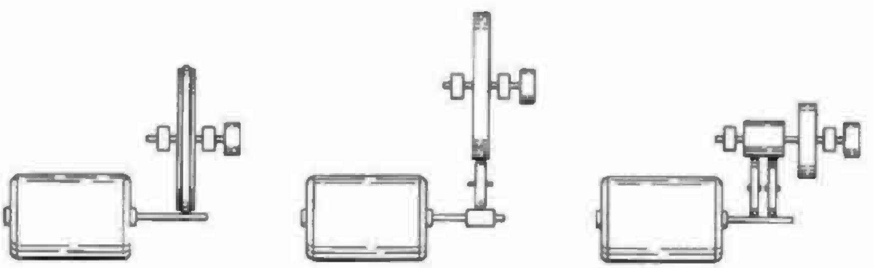
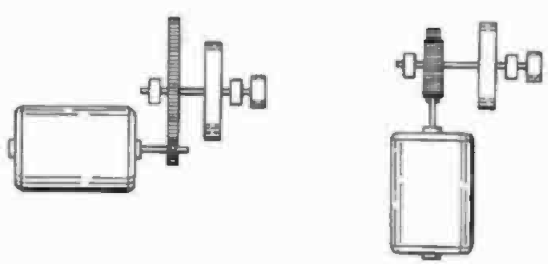


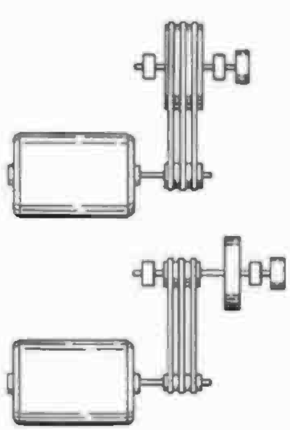
Fig. 3-70C. Internal and external connections for a 3-speed hysteresis motor. The speeds for this particular motor are; 300/600/1200 rpm.

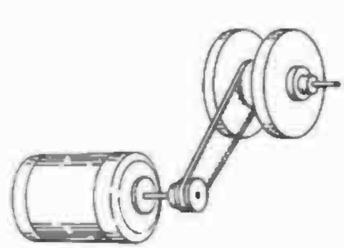


(a) *Puck or idler roller reducers.*



(b) *Gear reducers.*





(c) *Belt reducers.* (d) *Gilmer drive belt.*

Fig. 3-71. Drive systems used in recording and reproducing equipment.

out the rotational irregularities and surface imperfections of the idler or puck surfaces.

Gear drives have the advantage of positive nonslip drive between the motor and the driven members. However, rather elaborate decoupling devices are necessary between the gearbox and the driven member to eliminate vibrations from the gearbox, being transmitted to the driven system.

Single- and double-belt drives are quiet and require no lubrication, and also act as decoupling devices. Flat- or round-type belts may be used, but are subject to slippage. Belt-driven rim-drive devices, such as a recording or reproducing turntable, lose their flywheel effect when belt-driven, as the belt applies a damping effect to the drive system.

The best method of obtaining a positive drive with a belt is the "Gilmer Timing Belt Drive" shown in Fig. 3-71 (d), which incorporates a unique system of securing a positive speed reduction. Basically, it is a combination of a belt and a gear drive system, having the positive drive of a gear and the advantages of a belt.

The Gilmer belt drive is made of molded neoprene on a base of non-stretch cotton material. Precision teeth are molded into the belt which make contact with gear-type pulleys. Because of the characteristics of neoprene, there is a certain amount of natural compliance which eliminates the noise and ripple generally associated with gear driven systems. Also, no lubrication of any kind is required. Gilmer belt systems are used in many different types of recording and reproducing equipment where absolute synchronous speed is required. (See Sections 17 and 18.)

3.72 What is the relationship of frames per second to hertz?—Frequency meters on camera supply units are generally calibrated to read in both frames per second and hertz. This enables the recordist to increase or decrease the speed of the camera motor for special effects. If the camera is run overspeed, when the picture is projected later the action will be slowed-down; if the camera is run underspeed, the action will be speeded-up. In these instances, the camera is run wild, without sound. The relationship of frequency to frames per-second is:

40 Hz	16 frames per second
45 Hz	18 frames per second
50 Hz	20 frames per second
55 Hz	22 frames per second
60 Hz	24 frames per second
65 Hz	26 frames per second

When a change of frames per second is required and a calibrated scale is not readily available, the corresponding frequency may be determined by multiplying the desired frames per second by a factor of 2.5, which is derived by:

$$60/24 = 2.5$$

where,

24 is the frames per second for a standard frequency of 60 Hz. (See Question 22.37.)

3.73 Describe the different types of insulation used in motor windings.—Progressive advances in motor design, manufacturing techniques, and newly developed insulating materials have advanced to where a given size motor of a few years ago, is now considerably smaller and more efficient. The minimum physical size of a motor and its life expectancy are limited and determined by the destructive effects of internal operating temperature and winding insulation. The materials for a motor winding are divided into groups, and standardized by the Institute of Electrical and Electronic Engineers, (IEEE).

Class	*Maximum Spot Temperature*
O	90°C.
A	105°C.
B	130°C.
H	180°C.
C	No limit set

The electrical and mechanical properties of the insulated windings must not be impaired by the application of a permissible temperature for given classification. Cotton, silk, paper and similar materials may be used as a class-O insulation; however, if these materials are to be used as class-A it is necessary that they be impregnated or immersed in a liquid dielectric to afford greater insulation. Other class-A insulations are: molded and laminated materials with cellulose filler, phenolic and other similar resins, also films and sheets of cellulose acetate or other cellulose derivatives. Conductors are varnished. Class-B and class-H make use of materials such as mica, asbestos, Fiberglas, etc., with suitable binding substances. Where the temperature may be higher in class-C, materials such as mica, porcelain, quartz, and glass may be required.

3.74 Describe the operation of a centrifugal governor.—Centrifugal governors are used with both series and shunt ac or dc motors and generators. The governor consists of an insulated plate, mounted on one end of the armature shaft, with two stationary and two movable weighted contacts. The weighted contacts are adjusted to be opened by the centrifugal force acting on the movable contacts if the armature exceeds its rated speed by more than 5 percent. When the contacts are opened, a resistor is connected in the circuit, reducing the armature speed by the reduction of the line voltage. When the armature falls below the rated speed, the contacts close, shorting out the resistor. The voltage then rises as does the speed of the armature. The cutting in and out of the resistor is a continuous action as the armature speed varies. If the line voltage and the load are fairly constant, the governor is not too active, and the armature speed oscillates around the rated speed.

The disadvantage of a centrifugal governor is the high-frequency interference generated by the opening and closing of the contacts (with the possibility of generating noise), and the pitting of the contact surfaces. The contacts are normally shunted with a 0.25-μF high quality high-voltage capacitor. If this shunting capacitor opens, contact interference will be maximum. It may be desirable to connect a noise-suppression diode across the contacts and install a line-noise filter near the machine to prevent interference being fed back over the power line. This is particularly true if the device is installed near high-gain amplifying equipment. Noise-suppression devices are discussed in Question 24.67. The circuitry for connecting centrifugal governors is given in Fig. 3-74A and Fig. 3-74B.

3.75 Describe how dynamic braking is applied to a motor.—Dynamic braking can be applied to any motor circuit. However, as the motor increases in complexity, the switching circuitry becomes more involved. The braking effect to the motor is accomplished by disconnecting the power source and applying a direct current to one or more windings, thus reducing the rotation of the armature by magnetic drag until it comes to rest.

Several circuits are available for this purpose. Fig. 3-75A is the circuit for a shaded-pole motor. A dynamic braking circuit for split-phase start, capacitor start, or permanent split-capacitor motor is shown in Fig. 3-75B. Several oth-

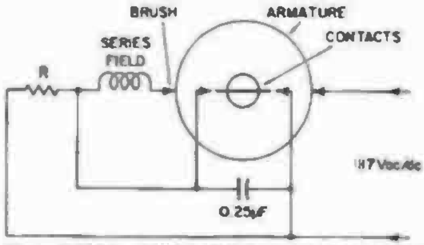


Fig. 3-74A. Series-field connection for centrifugal governor.

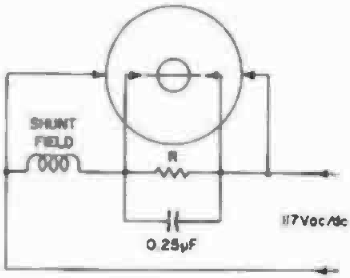


Fig. 3-74B. Shunt-field connection for centrifugal governor.

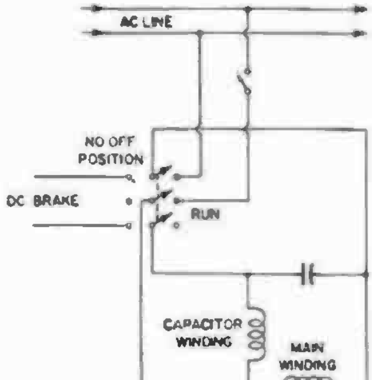


Fig. 3-75A. Dynamic braking circuit for permanent split-capacitor motor. Windings in series, when braked.

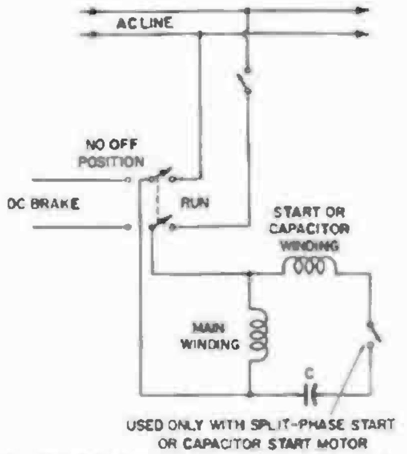


Fig. 3-75B. Dynamic braking circuit for split-phase start, capacitor start, or permanent split-capacitor motors. Main winding used for braking.

ers shown in Figs. 3-75C to E are self-explanatory. The direct current is supplied from a half-wave rectifier. Extreme care must be exercised in the braking of geared motors to avoid shocking beyond their capabilities. This is especially true for loads having considerable inertia.

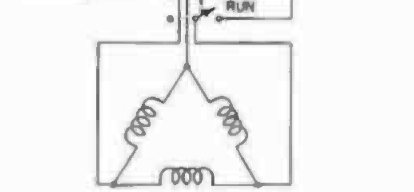


Fig. 3-75C. Dynamic braking circuit for three-phase motor. For delta connection only one phase of the stator is used. If star connected, two phases are used.

3.76 ***When certain motor data are known, how may other unknown related factors be determined?***—The nomograph in Fig. 3-76 provides a fast convenient way to determine factors of: Torque (in-oz) speed (rpm) frequency (Hz) horsepower and stator poles, as applied in the formulas:

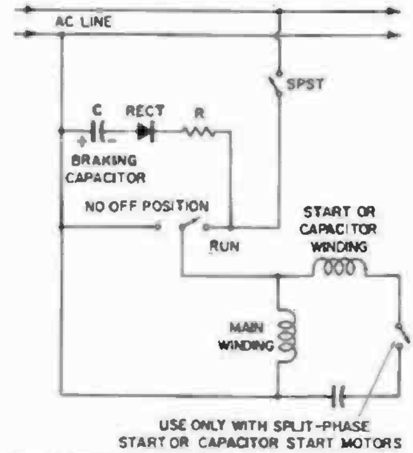


Fig. 3-75D. Dynamic braking circuit for split-phase start, capacitor start, or permanent split-capacitor motors.

$$\text{Speed} = \frac{120 \times \text{frequency}}{\text{Stator poles}}$$

Horsepower =

$$9.92 \times \text{Torque} \times \text{Speed} \times 10^{-7}$$

By connecting the points of known values on the nomograph, other values may be read. Example: To find the speed, given a frequency of 50 Hz, and 6 stator poles, the line connectiug these two values, intersects the speed scale

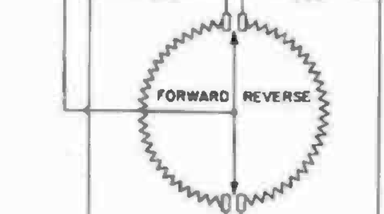


Fig. 3-75E. Dynamic brake and speed control of 3-wire reversible capacitor motor. Maximum speed will be obtained when the resistor shunted across the capacitor is minimum and will increase with an increase of resistance. Maximum speed is obtained when contact arm reaches the horizontal position and shunt circuit is open. Motor is braked when contact arm is vertical.

at 1000-rpm, line "A." When the horsepower is 0.250 and the speed is 1500-rpm what is the torque? The straight edge laid across the known values, torque is read at $1.68 \times 10^{x+2-y}$ ($x = 0$ $y = 0$) $= 1.68 \times 10^2 = 168$ in-oz, line "B." Given a horsepower of $\frac{1}{70}$ or 0.0143 or 0.143×10^{-1} then $x = -1$, with a speed of 3,600-rps or 360×10^1 then $y = 1$. The torque scale is intersected at $4 \times 10^{(x+2-y)}$ or $4 \times 10^{-1+2-1} = 4 \times 10^0$ $= 4$ inch ounces, line "C." (See Questions 25.116 and 25.136.)

3.77 ***Describe the characteristics of synchronous motors used for driving sound equipment.***—The synchronous motor employed for driving of recording and projection equipment is not the conventional synchronous motor generally found in an industrial plant. Motors for sound work are of special design and are of the variable reluctance type, using squirrel-cage construction for the armature, with salient poles milled in the armature laminations. Such armatures are constructed by using heavy copper bars running lengthwise of the armature, and shorted at each end by a heavy copper ring. Salient poles are milled the length of the armature laminations, at an angle of 7 to 10 degrees, using the armature shaft as reference (Fig. 3-77A.)

In the early days of sound, synchronous motors were not available and induction motors were made to run at synchronous speeds by cutting slots in the armature laminations. This was

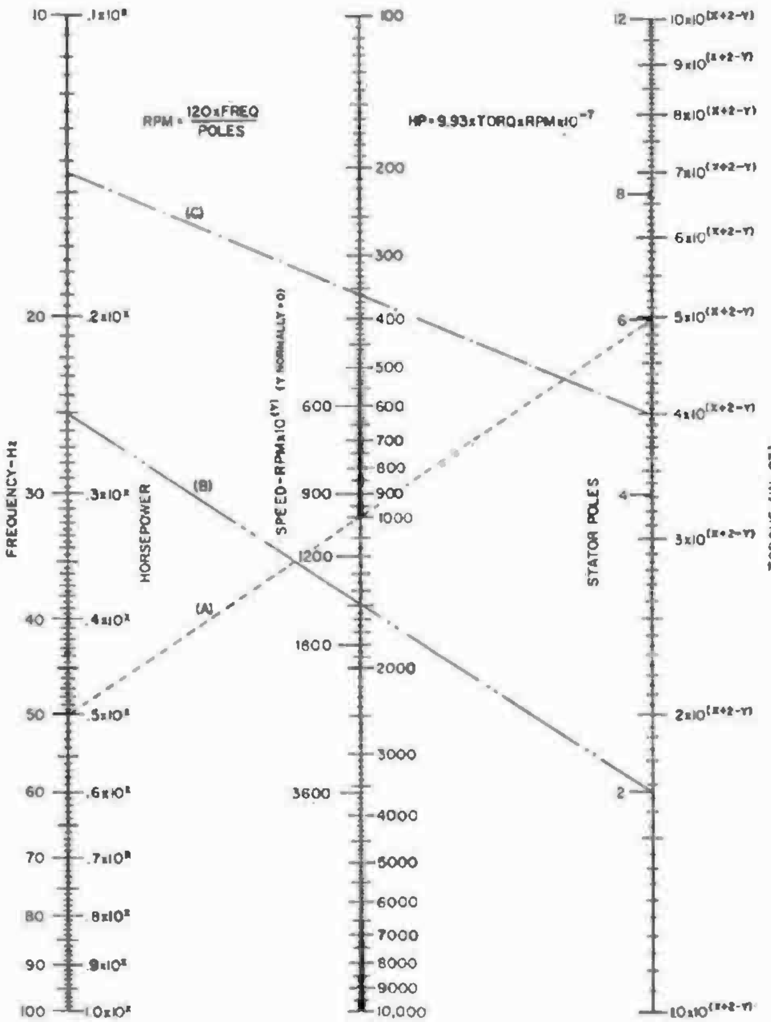


Fig. 3-76. Nomograph for determining the characteristics of motors. (Courtesy, Bodine Electric Co.)

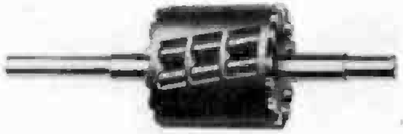


Fig. 3-77A. Squirrel-cage motor armature showing the copper rods, rings, and salient poles. (Courtesy, Bodine Electric Co.)

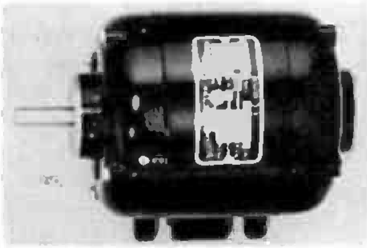


Fig. 3-77B. Fractional horsepower motor manufactured by Bodine Electric Co.

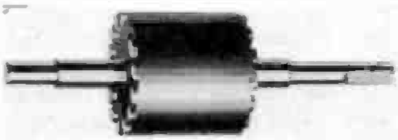


Fig. 3-77C. Armature of a standard induction motor. (Courtesy, Bodine Electric Co.)

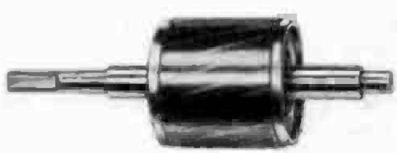


Fig. 3-77D. Armature of a permanent split-capacitor motor. (Courtesy, Bodine Electric Co.)

done by placing several hack-saw blades in a frame and cutting a slot about ⅛-inch in width, the length of the armature, at an angle of about 10 degrees; the armature was then rebalanced.

If the bars of the squirrel-cage rotor are of sufficient low resistance to bring the rotor to synchronous speed, the salient poles will pull the armature into step with the points of greatest flux density of the rotating field. As there are no windings, the armature requires no maintenance.

Synchronous motors may be designed to operate over a limited speed range by changing the frequency of the driving-power source. This usually requires a change in the supply voltage to maintain a satisfactory power input. Synchronous motors may also be provided with stator windings to produce a number of different pole combinations, and by using a suitable rotor, the motor may be made to operate at a number of different speeds. This is discussed in Question 3.70.

Because of the quick starting characteristics of squirrel-cage motors due to the copper bars, shorting rings, and salient poles, they have a high starting torque and are brought into synchronism very quickly. As this is somewhat of a disadvantage, particularly for the cameras and projection machines, some means must be taken to provide a soft start, such as putting resistors in the power source and cutting them out of the circuit as the motor comes up to speed.

Because such motors can not be interlocked at rest, a sync mark must be provided, to indicate when they reach synchronous speed. On production, this is accomplished by the use of clap-sticks held in front of the camera, and photographed while the sound is being recorded. Later, the picture of the clap-sticks and their sound is used by the editorial department to synchronize the picture with the sound track. (See Question 18.334.)

Motors of the above described type are characterized by their high starting torque, rapid acceleration, low power factor (because the exciting current is taken from the line through the stator winding), and efficiency. However, these factors are of little consequence as the power rating is low, generally on the order of 1/10 to 1/3 horsepower. Synchronous motors of this type cannot be used for background projection or rerecording purposes. In background projection, the shutter of the projector must be synchronized with the camera beforehand. If this were not done, the camera shutter might be opened while the projector shutter was closed, resulting in no exposure of the background projection. Selsyn interlock motors must be used to bring the camera and projector up to speed and in synchronization. When sync speed has been attained, clap-sticks are used to establish the synchronization of picture and sound. (See Question 19.59.)

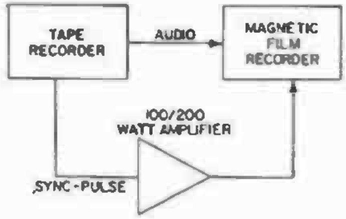


Fig. 3-78A. Basic resolver system using the sync pulse to drive the transfer machine motor.

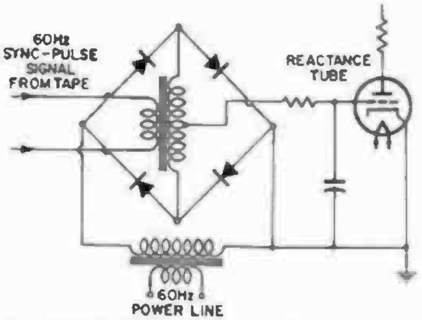


Fig. 3-78B. Bridge circuit with reactance tube for controlling the frequency of the 60-Hz oscillator driving the synchronous motor in the ¼-inch tape reproducer.

In rerecording installations, quite often certain machines are equipped with dual-purpose motors. These motors function as both selsyn interlock and a straight synchronous motor, which is quite convenient during a rerecording session, where a single track is wanted for listening purposes, and the balance of the machines are set for interlock running.

Transfers from one synchronous machine to another are permissible, provided a start mark is recorded on the new sound track when the machine has attained synchronous speed. It is quite common practice when transferring a magnetic sound track to a photographic sound track to start the photographic recorder first and then start the magnetic sound track. The machine carrying the magnetic track will settle down in a very short time, while the photographic recorder takes longer in comparison because of the transport-loop system. A fractional horsepower synchronous motor, manufactured by the Bodine Motor Co., appears in Fig. 3-77B. Figs. 3-77C and D show an armature for a standard induction motor and also one for permanent split-capacitor motor.

3.78 ***Describe the methods for recording a sync pulse on quarter-inch magnetic tape for synchronous operation with camera.***—Sync-pulse systems used for the synchronization of motion picture camera and ¼-inch tape recorders are quite common. Several different types of systems are in general usage. All these systems accomplish the same end result, but in a slightly different manner. Such systems are said to record magnetic sprocket holes, because they produce the same results as sprocket type machines. The first of these systems was developed by Col. H. B. Ranger, of the Rangertone Co. In this system, the sync pulse was the power-line frequency. A low level 60-Hz signal was recorded in the center of the tape, over the sound track, at an angle of approximately 87 degrees to the direction of the tape travel (Fig. 3-78C.) Later, when this track was transferred for editorial purposes, the sync-pulse signal on the tape was amplified by a power amplifier and used to drive the motor on the transfer machine (Fig. 3-78A). Any variation in the line frequency or camera during the original recording was transmitted to the transfer machine motor, duplicating the elec-

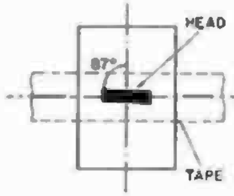


Fig. 3-78C. Rangertone sync-pulse head used on ¼-inch magnetic tape recorder.

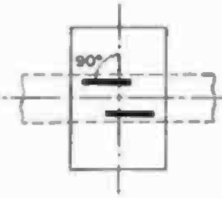


Fig. 3-78D. Echelon Perfectone sync-pulse head used on ¼-inch magnetic tape recorder.

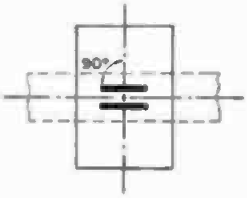


Fig. 3-78E. Nagra (Kudelski) sync-pulse head used on ¼-inch magnetic tape recorder.

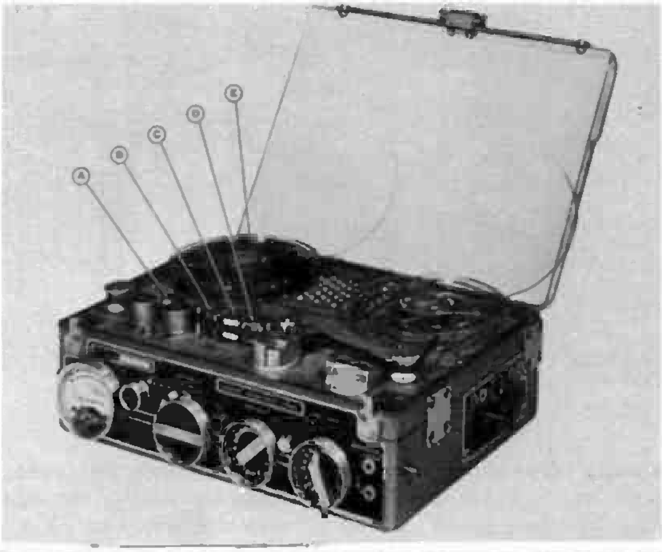


Fig. 3-78F. Nagra III ¼-inch magnetic recorder. (Courtesy, Magnetic Sales Corp.)

trical conditions prevailing at the time of the recording; thus, synchronization between the sound and picture was achieved.

A second system, developed by Ranger for transferring the original sound track, compares the sync-pulse signal on the tape to the 60-Hz power-line frequency driving the transfer machine motor. This comparison is made automatically by using a bridge circuit in which one arm is a reactance tube that controls the frequency of an oscillator, which is normally adjusted for 60 Hz (Fig. 3-78B). The oscillator drives the power amplifier which, in turn, drives the synchronous motor in the tape recorder now being used as a reproducer. If the sync-pulse signal at a given moment is exactly in step with the power frequency, the bridge circuit is in balance and the reactance tube is inactive. If the sync pulse on the tape gets ahead of the power frequency, the bridge becomes unbalanced and slows down the motor in the recorder until the sync pulse is again in step with the power-line frequency. If the sync pulse is slow, it is speeded up by the reverse process. Corrections to the speed of the tape must be made slowly, so as not to produce a noticeable change to the ear. Experience indicates that if the corrections are less than 1 cycle in two seconds, the variation in speed is not heard.

A third method of applying sync pulses to a tape recorder is termed the echelon system, used with Perfectone recorder, developed for location work by Ryder Sound Services. In the echelon method, the sync pulse is recorded

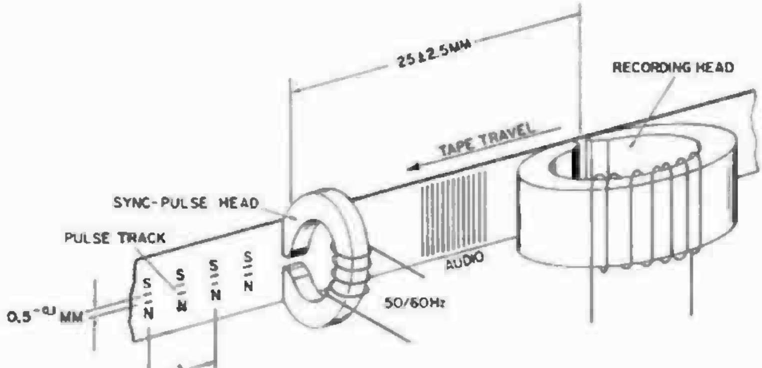


Fig. 3-78G. Rangertone transversal recording.

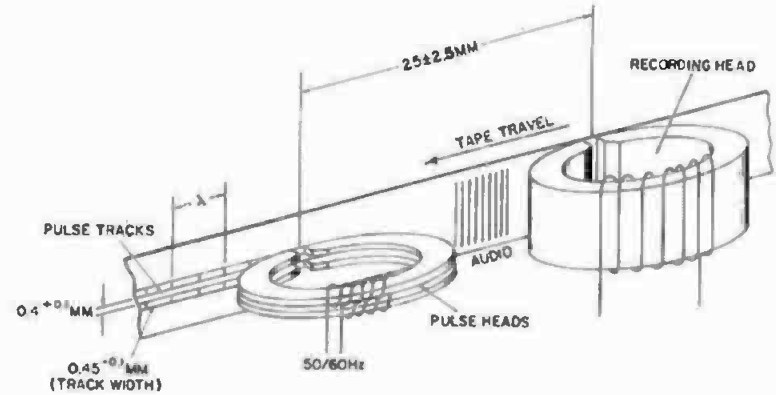


Fig. 3-78H. Nagra push-pull recording.

on the two outer edges of the tape at an angle of 90 degrees to the direction of tape travel. The pole pieces of the heads are staggered about one half wavelength of 60 Hz, running at a speed of 7.5-inches per second, or approximately 1/16 inch (Fig. 3-78D). The sync-pulse signal may be generated by a permanent-magnet generator mounted on the camera motor, or taken from the camera motor-supply voltage as discussed in Question 3.57.

A fourth method called the NeoPilot system, developed by Stefan Kudelski of Lausanne, Switzerland, is used with the well-known Nagra recorders. This system employs a sync-pulse head that records a 50- or 60-Hz push-pull signal in the center of the tape, at an angle of 90 degrees of the direction of tape travel (Fig. 3-78E). Although the construction of the NeoPilot head appears to be similar to the echelon type, it is not, since the pole pieces are in-line and staggered. The sync-pulse signal may be taken from the camera power source, or from a sync-pulse generator mounted on the camera motor. All of the systems that have been discussed employ approximately 1.25 Vac at the sync-pulse head.

A Nagra III recorder-reproducer is shown in Fig. 3-78F with the head placement indicated. At "A" is the erase head, "B" the sync-pulse head, "C" the record head, and "D" the playback head. The supply reel for the tape is shown at "E." This recorder is treated in detail in Question 17.177.

The positioning of the sync-pulse head for the Rangertone system is shown in Fig. 3-78G. Here, the sync-pulse head is mounted on the center line of the tape at a right angle to the audio signal after the tape has passed over the recording head. Some cross talk between the sync pulse and the audio signal may be expected using this system.

In Fig. 3-78H is shown the position of the sync-pulse head on a Nagra recorder using the NeoPilot system. Since in this system the sync-pulse head is of push-pull design, no cross-talk interference is observed between the sync pulse and the audio signal. All three systems are compatible except for the different methods used for taking off the sync-pulse signal. For the Ryder echelon head, it is placed in the same position as for the NeoPilot system.

In the absence of a recorder especially designed for sync-pulse operation, a stereophonic two-track recorder may be used. In this instance the sync-pulse signal is recorded on one track and the audio signal on the other. The sync-pulse signal is then reproduced and used to drive the transfer recorder, or interlock other machines as previously discussed. (See Question 3.80.)

3.79 *Can frequencies other than 60 Hz be used for sync-pulse signals?*—Yes, they may be operated at 50, 60, 100, 120, and, in some instances, 14,000 Hz. However, if the higher frequency is used, the recorder must be capable of recording a strong 14,000-Hz signal on the tape, and precautions taken to prevent leakage of the sync signal into the recording circuits and causing beats. Pictured in Fig. 3-79 is a resolver manufactured by Magna-Tech Electronic Co., Inc., Model 92B. This device is designed to operate at both 60 and 14,000 Hz. The sync-pulse head is mounted between

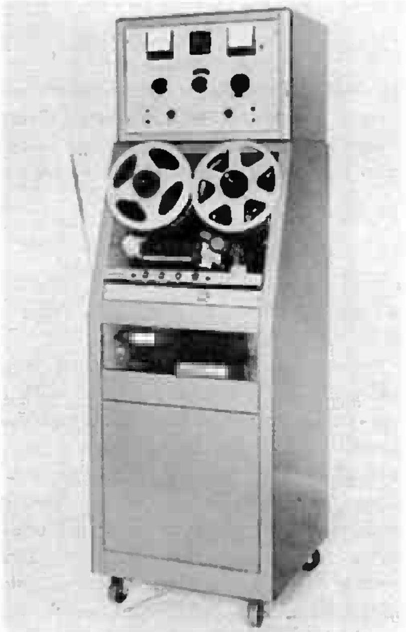


Fig. 3-79. Magna-Tech Electronic Co., Model 92B playback synchronizer operating in conjunction with an Ampex ¼-inch tape recorder.

the normal head assembly. The capstan and pinch-wheel are at the right.

3.80 Describe the methods used to connect a tape recorder and camera for sync-pulse operation.—Five different methods of connecting or operating a ¼-inch tape recorder and camera for sync-pulse operation are shown in Fig. 3-80A through E. In Fig. 3-80A the tape recorder is operated from its internal batteries and is fed a sync-pulse signal from a permanent-magnet generator (60 Hz) mounted on one end of the camera motor shaft. The power supply for the camera consists of a three-phase 220-Vac 60-Hz generator driven from batteries. With the sync-pulse voltage taken from the camera motor-driven generator, any change in camera speed is reflected in the frequency of the sync pulse and recorded on the tape, along with any small variations in the battery-driven motor in the recorder.

The system shown in Fig. 3-80B is similar, except that the sync-pulse voltage is taken from a step-down transformer connected across two phases of the three-phase voltage fed to the camera motor. Any change in the frequency of the camera-supply voltage is recorded on the tape. When shooting in a studio, the camera is generally supplied from the house mains, which also supply the sync pulse through a step-down transformer (Fig. 3-80C).

Systems are also available where the sync pulse is transmitted by a small radio transmitter to the recorder, as in Fig. 3-80D. Another system for driving the camera employs a transistorized inverter operated from batteries, and generating 117 Vac, at 60 Hz (Fig. 3-80E). The frequency of the generator is precision-controlled. The sync-pulse signal is taken from a small precision unit which generates 60 Hz with an accuracy of ±0.0005 percent. The manufacturer states the inverter and sync-pulse generator have an accuracy of 20 parts per million, over a temperature range of minus 20°C to plus 50°C (minus 40°F to plus 122°F); or 5 parts per million over a temperature range of plus 8°C to plus 34°C. Using this synchronization, several cameras can be used simultaneously without interconnecting cables. While there are other

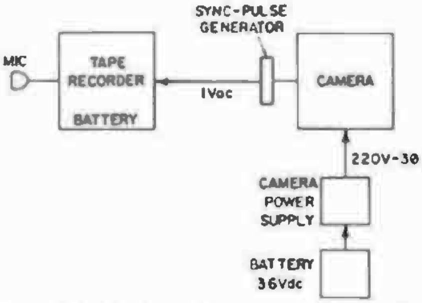


Fig. 3-80A. Recorder operated from internal batteries; the sync-pulse signal is obtained from pulse-generator on camera; camera powered from battery power supply.

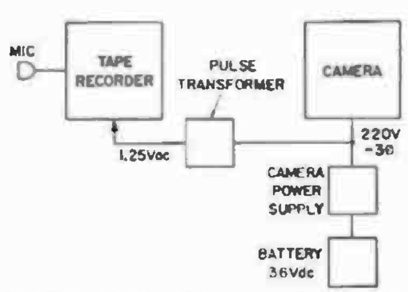


Fig. 3-80B. Recorder powered from internal batteries. camera from battery power supply; sync-pulse from step-down transformer across two-legs of camera three-phase power supply.

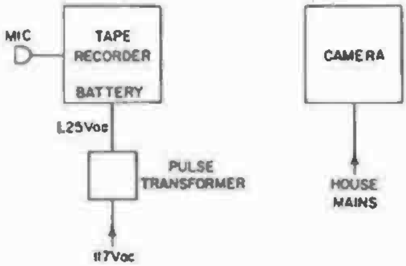


Fig. 3-80C. Recorder is battery operated; camera operates from 220-volt, three-phase mains: pulse signal from step-down transformer across 117 Vac mains.

variations of the described systems, those shown in Figs. 3-80A through C, are the most commonly employed. (See Question 3.78.)

3.81 Describe a system for interlocking nonsynchronous recorders and reproducers.—The problem of interlocking nonsynchronous sound equipment and picture has been one of the chief difficulties of the sound industry, because synchronous motors, although rated constant-speed, are only approximately so. They are subject to slight speed changes due to possible variation of the coupling angle between the rotating field and the rotor. The angular velocity may not be constant, but the average speed will be synchronous.

Nonsynchronous motors never operate at synchronous speeds, even at no load, and the difference between synchronous and actual speed will depend upon the design, power input, and torque requirements. They are also affected by the line voltage and the load. To overcome this difficulty, the Radio Corporation of America, has developed a system of interlocking synchronous and nonsynchronous machines, with a trade name of Unilock.

To explain its operation, in Fig. 3-81A is shown a tape recorder with a sync-pulse recorded on the tape (see Question 3.78), interlocked with a standard 35-mm or 16-mm projection machine. The only piece of equipment required at the projector is a rate-generator mounted in the film path. The moving film drives a drum containing holes that chop a light beam between a small lamp and photodiode, generating a 60-Hz signal at a frame rate of 24 frames per second. The output voltage of the rate generator is applied to the input of a Schmitt-trigger amplifier in channel "B" of the Unilock system. The output from the sync-pulse head on the tape recorder is applied to a second Schmitt-trigger in Channel "A." To operate, the tape and film are threaded at their respective start marks, and the system thrown to the interlock mode and the projector and tape transports started.

During the acceleration period, the tape will attain sync speed almost in-

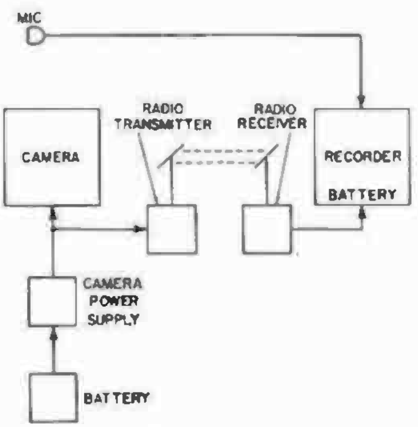


Fig. 3-80D. The camera is operated from a battery-operated generator. The sync-pulse is sent to the recorder by radio.

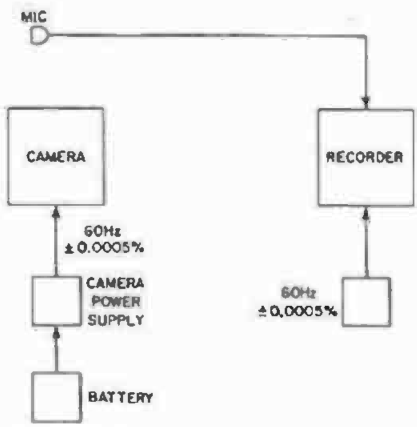


Fig. 3-80E. Camera operated from 60-Hz generator with a regulation of 0.0005%. 60-Hz signal with some percent regulation applied to recorder.

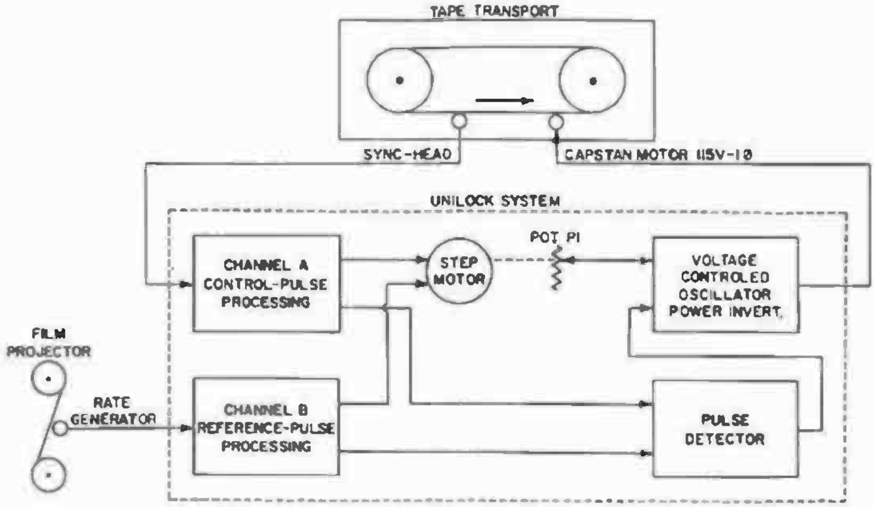


Fig. 3-81A. Film projector and tape recorder interlocked using a Unilock system.

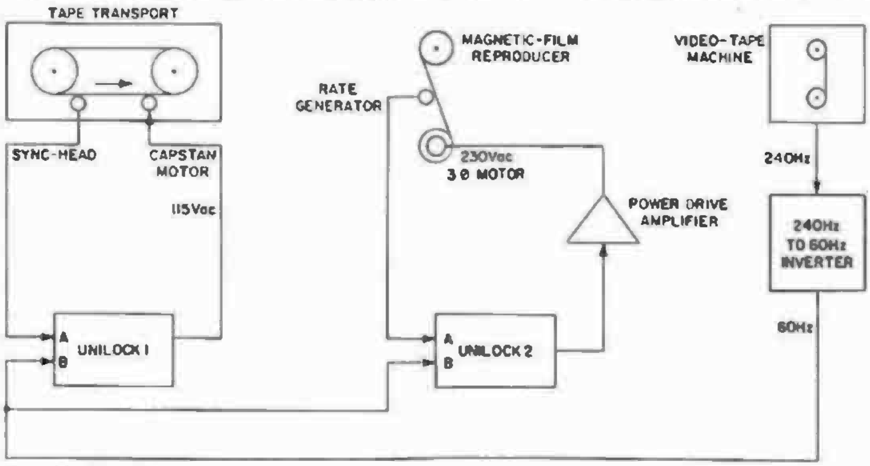


Fig. 3-81B. Tape recorder, 35- or 16-mm magnetic-film machine, and video-tape recorder interlocked using a Unilock system.

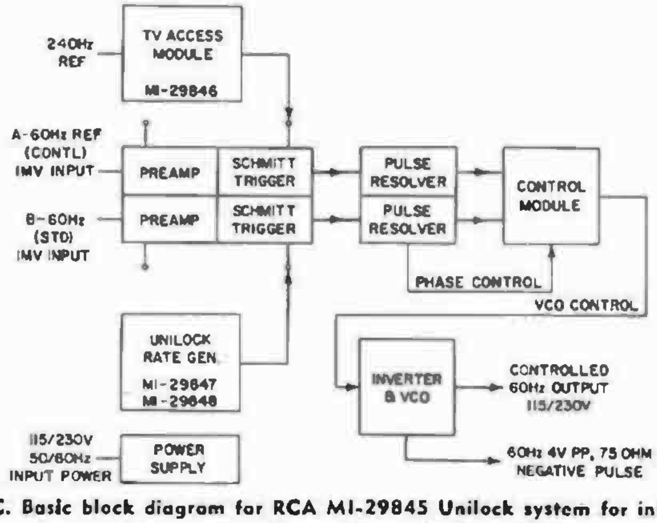


Fig. 3-81C. Basic block diagram for RCA MI-29845 Unilock system for interlocking nonsynchronous recording and reproducing equipment.

Fig. 3-81D. Unilock sync or pick-off head mounted on an RCA TP-66 16 mm projector.

stantaneously, while the projector accelerates more slowly. This produces a lower pulse rate in channel "B." As the pulse-count differential between the two channels builds up, the stepping motor turns potentiometer P1 so as to slow down the tape-transport capstan. As the projector attains full speed, the pulse rate from the rate generator approaches 60 Hz. The number of pulses in channel "B" will eventually catch up with the total pulses of Channel "A." In so doing, the stepping motor will bring potentiometer P1 back to the position where the capstan motor is driven at synchronous speed, thus synchronism is attained. The time required for the system to stabilize is approximately 5 seconds. Even if the system is stopped, it can be started again with both machines in synchronism.

To use the system with a video-tape recorder, a magnetic-film reproducer, or a projector, the equipment is connected as shown in Fig. 3-81B. The 240-Hz control-track signal from the video machine, after suitable amplification, is applied to a converter which reduces the 240 Hz to 60 Hz. The inverter output drives a 230 Vac three-phase power drive amplifier, which supplies the motive power for the magnetic-film reproducer. Many other combinations of this system are possible and any number of

Fig. 3-81E. Reference signal rate-generator mounted on RCA TP-66 16mm projector.

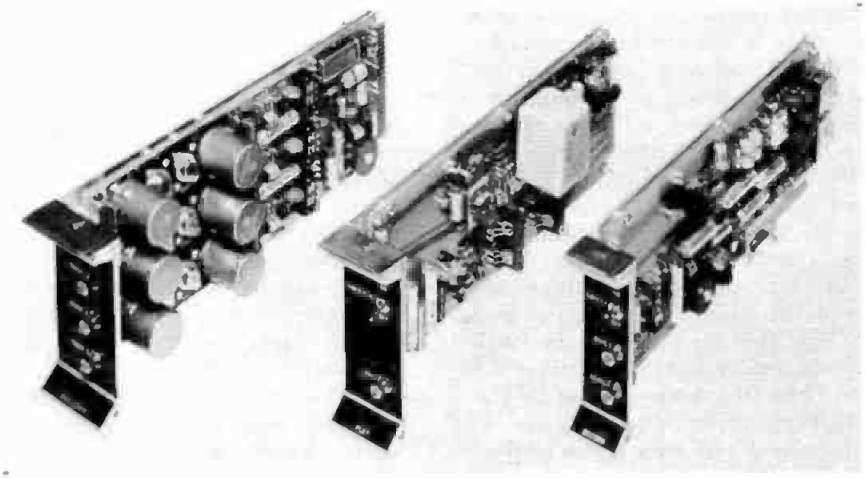


Fig. 3-81F. RCA Unilock control modules.

machines can be interlocked. Control equipment is available for either 50- or 60-Hz operation.

A basic block diagram in Fig. 3-81C is completely solid-state. The inverter unit will supply 30 watts at 117 Vac for driving the capstan motor of ¼-inch tape recorders, or 300 watts at 230 Vac for driving the motor of a magnetic-film recorder/reproducer. To operate a ¼-inch tape recorder, a sync or pick-off head is required, and is shown mounted on an RCA RT-21 tape recorder. A reference signal rate-generator is shown installed on an RCA TP-66 projector in Fig. 3-81E. Three modules associated with the control system are shown in Fig. 3-81F, with an inverter, preamplifier, pulse resolver, control, and power supply shown in Fig. 3-81G.

3.82 Describe the methods used for phasing a selsyn interlock system.—Assuming the system is designed for three-phase operation, the phasing is started at the synchronous motor driving the selsyn generator set. The driving motor is phased for the correct direction of rotation by connecting its leads to the three-phase power source, and noting the direction of the rotation, which should be clockwise when facing the shaft end driving the selsyn generator unit. If the rotation is counterclockwise, reverse two of the power leads. The generator unit will have 6 leads, numbers 1 to 3 being the stators, and 4 to 6 the rotors. These 6 leads are run to each machine to be driven, and the leads numbered similarly to the generator leads. The stators of all mo-

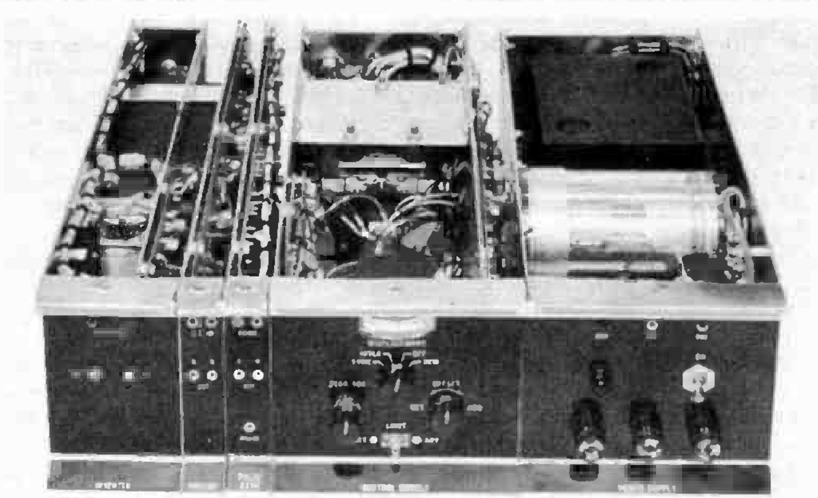


Fig. 3-81G. RCA Unilock control modules.

tors are connected to the stator lines from the selsyn generator, number for number. The selsyn generator is started and the rotors of each motor shorted together, without load, and the direction of rotation noted. If the motor rotates in the wrong direction, reverse the two stator leads. This procedure is followed through for the whole system. After phasing the motors, two phases of the power line (locking position) are closed. The rotors should then all pull into lock. Rocking the rotors by hand, the lock should feel firm. Closing the third phase, the motors should all run at 1200 rpm in the same direction.

It is possible in some selsyn systems to have what appears to be correct phasing, but one motor does not lock solid, and has a tendency to run away (1800 rpm). This condition must be corrected or the motor will overheat and run away frequently. This condition can be caused by the leads on the stator side not being properly phased, although about 90 percent of the time it operates normally.

Single-phase interlock systems are somewhat simpler to phase. The single-phase line is connected to the rotors, and two of the three stator leads are reversed for the correct direction of rotation. If the selsyn is a dual-purpose type (selsyn and synchronous) designed for three-phase operation, it will have 9 leads. Leads 7 to 9 are for synchronous operation. These leads are connected to a source of 230-volt three-phase power and reversed for the correct direction of operation. Care must be taken that the synchronous switch is never thrown when the system is operating as a selsyn motor.

3.83 ***What is the purpose of reversing a rerecording system?***—Reversible rerecording systems serve two important purposes; they save rehearsal time, and they afford the operating personnel greater convenience of operation. It is the current practice to rerecord a full reel of picture and sound track that may run from 940 to 980 feet in length. Many times a difficult cue or sound effect is encountered down in a reel 500 to 600 feet from the start. If the cue is such that it may require several rehearsals, this can become costly and time consuming, as all the sound tracks and picture must be rewound for each rehearsal. Interlock systems can be designed to hold the machines in sync, while reversing and running back to a portion to be again rehearsed. After reversing, they may be run forward again without loosing synchronization. This may be done as many times as required.

Interlock systems designed for reversal operation require special treatment. The picture gate in the projector must be suitable for running in reverse, or modified to do so. The modification will differ with each make of machine. Solenoids must be installed to lift the pad roller off the impedance drum in the sound head when running in reverse. Also, the fire shutter must be closed when the picture comes to a stop at the point of reversal. Most magnetic film machines will run reverse as they have torque motors on the feed and take-up spindles.

In some installations, the magnetic recorder is provided with an erasure system for erasing a portion of the sound track, while running in reverse, to permit recording that portion over again. Recorders are available that will permit a single word or musical note without a trace of the substitution.

Distributor systems can be reversed in several different ways. Two of the most common are the reversing of the power leads by a system of relays, and reversing the power circuits and rotating the distributor unit in reverse by means of a separate motor running at 120 feet per minute. Reversal systems are usually designed to fit a particular installation.

3.84 ***Describe a resolving system for transferring ¼-inch tape, using a sync-pulse to drive a sprocket-driven recorder.***—A resolver is a device used for transferring a ¼-inch magnetic tape sound track employing a sync-pulse signal, to a sprocket driven recorder, while maintaining absolute synchronization. Several different methods are available; they will be discussed in their proper order. The reader is referred to Question 3.78, before reading the following discussion.

Assume a motion picture is being shot, where the camera is being supplied with a three-phase 220-volt 60-Hz voltage from a battery-driven power supply, and the ¼-inch tape recorder is driven from internal batteries. The camera motor power supply does not generate a constant frequency, even with constant-frequency control, but will vary both above and below the nominal frequency. If the alternator is generating 61.0 Hz, the camera motor is turning 1.66 percent faster, and if the alternator is generating only 59 Hz; the rotor is 1.66 percent slower, with reference to 60 Hz.

In the first method of transfer that might be used, the battery-driven recorder is now used as a reproducer (Fig. 3-84A). The sync-pulse signal is amplified by a 60- to 100-watt amplifier, which drives the synchronous motor in the sprocket-driven transfer recorder. Since the power for the transfer recorder motor is the original pulse signal recorded during the shooting of the sound track now undergoing transfer, the sprocket-driven machine motor will increase and decrease its speed in

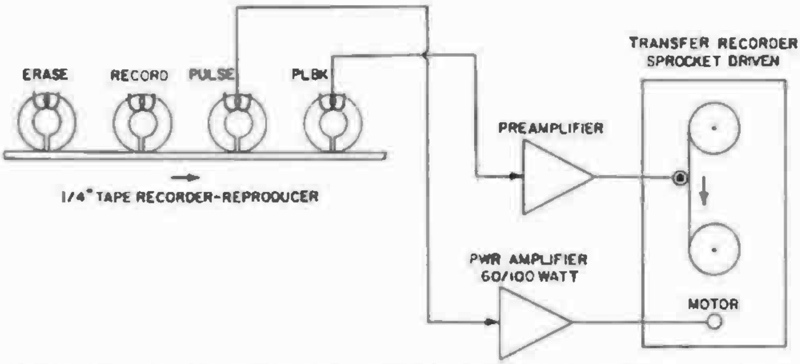


Fig. 3-84A. Direct method of transferring ¼-inch tape using a sync-pulse signal to a sprocket driven machine.

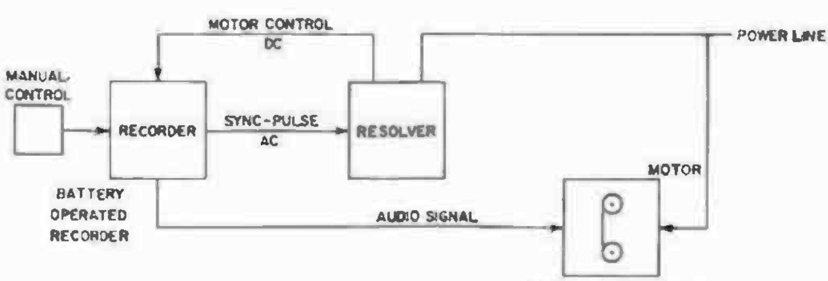


Fig. 3-84B. Kudelski Nagra III ¼-inch tape recorder connected to a sync-pulse resolver. The sync-pulse signal is fed to the resolver from the recorder. After correction, it is returned to the motor-control circuits in the recorder. The manual control extends the range normally outside the resolver range. The audio signal is fed from the recorder to the transfer machine in the usual manner.

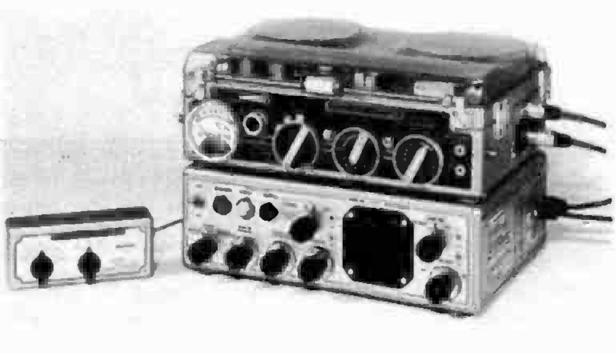


Fig. 3-84C. Kudelski Nagra III and Model SLO resolver, with Variator manual control at left.

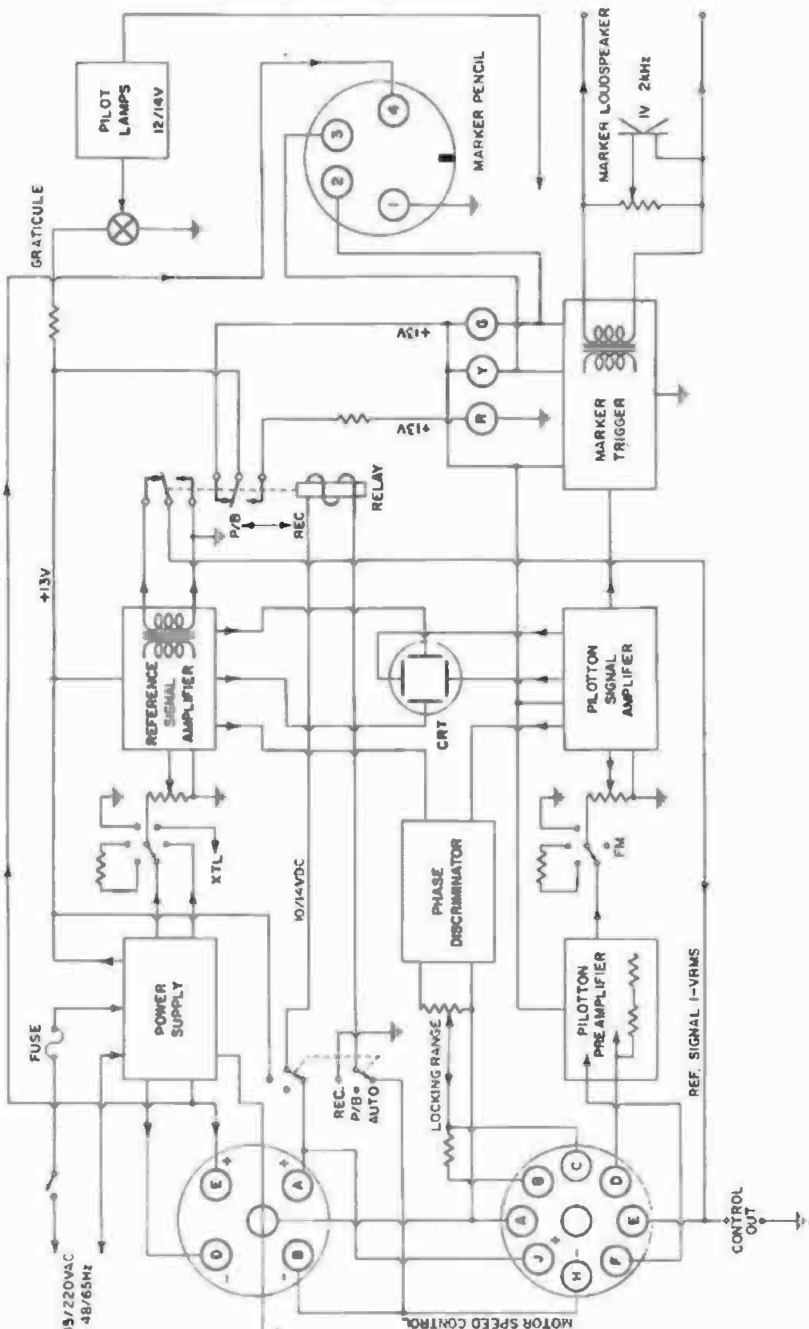


Fig. 3-84D. Block diagram for Kudelski Model SLO resolver for use with Nagra recorders or similar equipment. The two plugs at the left connect to the motor-control circuits in the recorder and the switch circuitry. The Variator control plugs into the recorder.

accordance with the record of the sync signal on the tape. Thus, the same result is obtained as if the original sound track had been recorded on the sprocket-driven machine. It will be observed that in this system and all others to be discussed, corrections are automatically made for variations in the speed of the tape recorder, elongation of the tape, and any other elements that disturbed the linear speed during the recording of the original tape. This method can be used with satisfactory results, when other methods of transfer are not available.

In the second method, the same results are obtained by reversing the procedure. Here, the sprocket-driven machine is powered from the studio power line, and the speed of the motor driving the tape recorder is controlled by a resolver to synchronize it with the studio power-line frequency. The advantage of this method over the first is that there are always slight discrepancies between the sync signal on the tape and the power line frequency, and in many instances it is next to impossible to avoid parastic oscillations, resulting in slow drifts and low frequency flutter. This second method eliminates these difficulties, and is the system used by Rangertone. (See Question 3.78.)

A third resolving system (Fig. 3-84B), developed by Kudelski of Lausanne, Switzerland for the Nagra line of recorders, is of different design. The principal component of the circuitry is a phase-discriminating circuit for comparing the frequency of the sync pulse on the tape to that of the studio line frequency. The phase difference is converted into a pseudodirect current which is used to correct the speed of the dc motor in the Nagra recorder as it reproduces the sync-pulse signal. Three types of reference signal may be employed: the internal line frequency (50 or 60 Hz), twice the line frequency (100 or 120 Hz) of any voltage up to 50 volts, or a signal generated from a quartz crystal clock for instrument use. Kudelski uses the name *NeoPilot* for the sync-pulse voltage. For normal transfer work, a cathode-ray tube displays the reference signal (line frequency) horizontally, and the sync signal from the tape vertically. When the reference and pulse signals are in absolute synchronization, the oscilloscope display is a steady oval nonmoving Lissajous pattern. The latitude of the speed control on the tape recorder motor circuit is ±2.5 percent, for a linear speed of 7.5 ips. If a greater spread is required, a manual control, called a Variator, is connected to the recorder by means of a cable, which permits the speed control operating range to be extended. There are also internal controls for extending the speed control range where a Variator is not available. If a malfunction occurred during the original recording, it may be necessary to control manually the motor speed correction. This is accomplished by slowly rotating the Variator. As a rule, most transfers will fall within the 1.5-percent range. If the power line was used for the sync signal during the recording of the original tape, the differences in speed will fall well within 0.5 percent.

Included in the resolver is a control for selecting either automatic or manual control of the tape recorder motor speed. If a tape is being transferred automatically and a dropout occurs, a marker trigger circuit becomes operative when the pulse signal drops to a predetermined level. The trigger circuit actuates an audible signal from a built-in loudspeaker, turns off a green sync-signal indicator light, turns on a yellow warning light, and actuates an external circuit which may be applied to a pencil-marking device on the transfer recorder. This latter device, however is not necessary, but is quite handy if the transfer operator cannot keep a steady watch on the operation, since it indicates dropouts in the original sound track and where they appear on the transfer track. The resolver may also be used to record a sync signal when a ¼-inch tape is being rerecorded from a sound track on a sprocket-driven machine for playback purposes.

A front view of a Kudelski type SLO resolver is shown in Fig. 3-84C, with a *Variator* manual control at the left. Above the resolver is a Nagra III ¼-inch tape recorder. A simplified block diagram for the internal circuitry of the resolver appears in Fig. 3-84D.

3.85 Show a block diagram for playing back ¼-inch tape on location using a resolver.—The Kudelski Type SLO resolver can be quite useful on location for making immediate playbacks after recording, or from tapes previously recorded in the studio. A block diagram of the connections appears in Fig. 3-85. Here is shown a Nagra recorder being used as a playback machine, in conjunction with the resolver, fed from a source of 50 or 60 Hz. The audio signal is fed to a power amplifier and loudspeaker.

If local power is not available, the equipment may be fed from the cam-

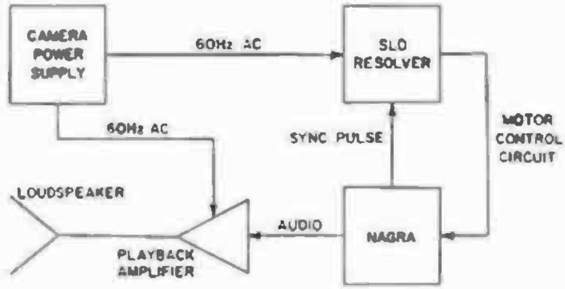


Fig. 3-85. Location playback system using a camera power supply unit for a source of 60-Hz power. The power supply must be adjusted carefully to hold the frequency as near 60 Hz as possible.

era-motor power supply system. In this latter instance, the camera power-supply frequency must be maintained as near to the line frequency of the studio as possible. A camera supply system with automatic frequency control is ideal for this purpose. Generally, for playbacks on location the tapes are prerecorded (using the studio mains) with the sync-pulse signal on the tape. This assures absolute synchronization of the picture with the playback.

3.86 Describe a synchronous resolver using a stroboscope lamp.—A unique method of obtaining synchronous sound transfer is by the use of a Strobelite resolver (Fig. 3-86A) developed and manufactured by Magnetic Sales Corp. Using this device, absolute synchronization can be obtained between any type of ¼-inch sync-pulse tape recorder and a synchronous sprocket-driven recorder. Synchronization is achieved by manual control of either machine, depending on the basic setup.

In Fig. 3-86B is shown a ¼-inch sync-pulse tape recorder reproducing an audio signal which is fed to a synchronous sprocket-driven machine, and recorded in the usual manner. The sync-pulse signal is fed to the resolver unit, amplified, then applied to a *stroboscopic* lamp in the resolver. This lamp

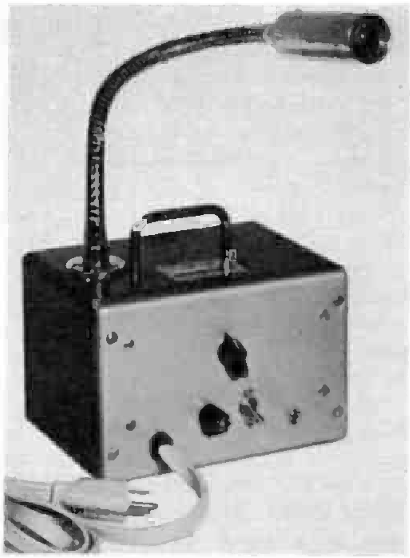


Fig. 3-86A. Strobelite resolver manufactured by Magnetic Sales Corp.

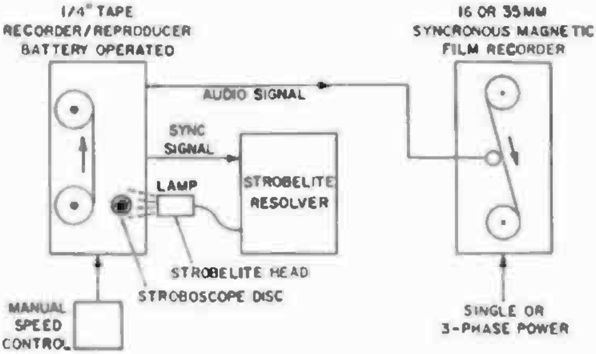


Fig. 3-86B. Block diagram for transferring ¼-inch sound track to magnetic film using a Strobelite resolver manufactured by Magnetic Sales Corp.

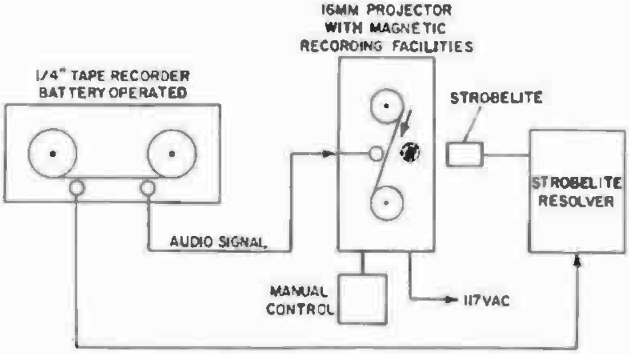


Fig. 3-86C. Basic setup for making synchronous sound transfers from a sync-pulse tape recorder to a 16-mm projector with sound recording facilities.

flashes once for each cycle of the synchronizing signal. The lamp flashes are employed to illuminate a *stroboscope* disc, mounted on the recorder (some recorders come factory equipped with these discs). When the image on the disc is brought to rest, the tape is running at the exact speed as when it was

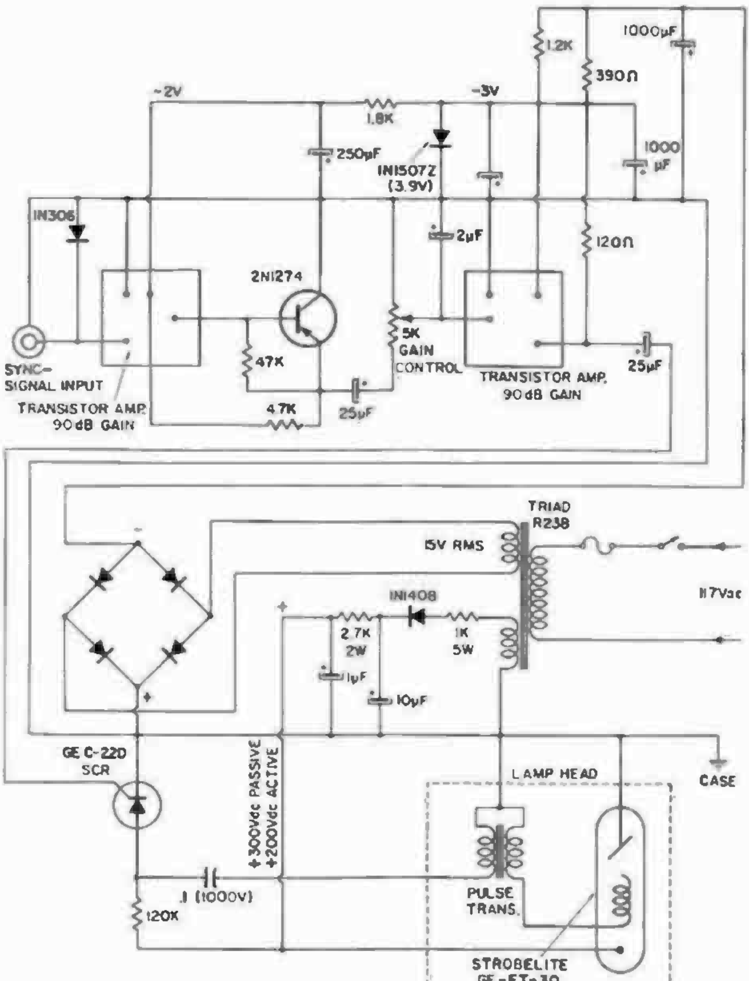


Fig. 3-86D. Schematic diagram for Magnetic Sales Corp. Strobelite resolver.

originally recorded with the picture. The external *manual* control for the tape recorder is adjusted to bring the tape recorder to the proper speed for stopping the image on the stroboscope disc.

A second method for using the Strobelite resolver is shown in block diagram in Fig. 3-86C. In this setup, the tape recorder is run at a fixed speed and the audio signal is applied to a 16-mm projector equipped with magnetic recording facilities. A stroboscope disc is mounted on one of the exposed rotating shafts. (See Question 3.87.)

The sync-pulse signal from the tape is applied to the resolver input, and the stroboscopic lamp placed in front of the stroboscope disc mounted on the projector. The speed of the projector is adjusted manually for a standing image on the disc by connecting a variable resistor (about 150 ohms, 50 watt) in place of the fixed resistor connected normally in the speed-governor control circuit (see Question 3.74). This will permit the speed of the projector/recorder to be varied above and below 24 frames per second.

To operate the system, the recorder is started and the speed of the projector/recorder adjusted for a *steady image* on the stroboscope disc, and thus maintained throughout the transfer. The sound may be recorded on a striped picture or on magnetic film.

The schematic diagram for the Strobelite resolver is shown in Fig. 3-86D. The sync-pulse signal is applied to a transistor amplifier, which feeds the signal to the gate of a silicon controlled rectifier (SCR). (See Question 11.150.) When the SCR fires, it applies a sharp pulse to a flash tube through a pulse transformer stepping up the voltage high enough to fire the stroboscopic lamp. The power supply provides 200 Vdc for the SCR tube, and low voltage for the transistors. (See Questions 3.81 and 3.84.)

3.87 Show a group of stroboscope discs suitable for use on a rotating shaft to indicate synchronous speed.—Several stroboscopic discs that may be duplicated and fastened to the end of a rotating shaft or sprocket, for checking synchronous speed, using a neon or regular lamp operated from the power mains, are shown in Fig. 3-87. For speeds other than those given they may be calculated as given in Question 13.117.

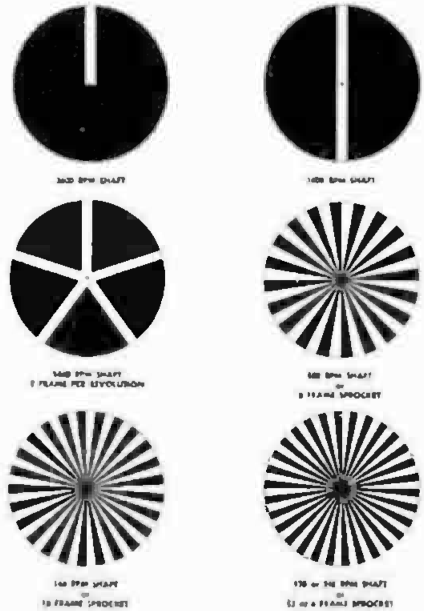


Fig. 3-87. Stroboscope discs. (Courtesy, Magnetic Sales Corp.)

For shaft speeds of 3600 rpm, only one bar is used to make it easier to distinguish. At speeds where the bars are large in number, they may be divided into even multiples of the basic number.

3.88 Describe a portable camera motor-drive unit for a constant source of 3-phase 60-Hz 230 Vac.—In the production of motion pictures, the demand for a portable synchronous motor-drive system has been going on for years, with many different systems having been proposed and developed. The basic requirement of such a system is a source of 230 Vac, 60-Hz three-phase power, that will interlock the camera with the sound equipment and hold the frequency to within a very small percent of the basic frequency (in USA it is 60 Hz), and yet be operated from batteries.

Since the advent of the battery-driven ¼-inch tape recorder for location work, the load of the sound equipment has been eliminated. Thus, the power-supply unit is now only required to supply power for the cameras and a low voltage (1.5 Vac) for the sync-pulse signal recorded on the tape recorder. (See Question 3.78.)

The schematic diagram for a camera control and generator unit, developed by F. G. Albin and manufactured by Magnetic Sales Corp., is shown in Fig. 3-88A. The system consists of a control cabinet and a soundproof case, which houses a 250 Vac 3-phase 500-watt (alternator) rotary inverter turning 3600 rpm driven from a 30-volt battery supply. The control cabinet supplies a means for manually or automatically controlling the frequency of the supply voltage. The control circuit consists of an elaborate network of transistors and diodes, which maintain the dc voltage constant at the motor terminals driving the alternator. In addition, the alternator is fitted with a centrifugal mechanical governor, which holds the dc voltage approximately constant to the driving motor, when the field current is adjusted for a given value. With the two methods of control, it is possible to hold the frequency to within less than ±½-cycle in 60 Hz. (See Question 3.74.) Provision is also made for manual control of the camera speed for special effects shots, where the camera must run faster or slower than 24 frames per second. Three meters are installed on the control unit for monitoring the dc driving voltage, generated 3-phase voltage, and frequency read in frames per second. (See Question 3.72.) The 1.5-Vac sync pulse for the tape recorder is supplied through a step-down transformer from one leg of the 230 Vac. Individual controls are provided for adjusting the field current for manual and automatic control modes.

When such a system is operated in the manual mode, the frequency must be continuously monitored to assure a constant frequency of 60 Hz. In the automatic mode as soon as the start switch is thrown (the generator settles down, in 1 to 2 seconds), the frequency is held constant at 60 Hz and requires no further adjustment, provided the field current has been adjusted for a particular camera load in advance of the shooting, since cameras vary in their load requirements. Operating in the automatic mode saves considerable camera film, and is an important factor, particularly if the picture is being shot in color. To prevent damage to the transistors, a fan in the control unit automatically starts if the internal temperature of the control cabinet rises above 100°F. The fan is automatically turned off during a take.

When more than one camera is used, it is generally necessary to add capacitance across each leg of the 3-phase alternator to provide power-factor correction. For convenience, external capacitor banks consisting of two banks of 10 and 20 uF each, are plugged into the power supply for switching across each leg of the alternator. The same value of capacitance is always used across each phase. The capacitors should be oil-filled. *Under no circumstances are conventional electrolytic capacitors permissible*, because of their internal leakage and the possibility of exploding. Nonpolarized electrolytic capacitors can be used; however, the paper or oil-filled type is preferred. In most portable alternator sets capacitors permanently connected across the three phases serve two purposes. First, they act as noise suppressors for the slip rings; second, they provide power-factor correction for a single camera.

When two or more cameras are to be driven, the cameras are run in advance and the capacitance adjusted for proper operation. In Fig. 3-88B, is shown a typical capacitor bank, consisting of six 10-uF banks, that may be connected for either 10- or 20-uF operation.

3.89 ***Describe a printed-circuit motor.***—The printed-circuit motor is a direct-current electrical machine, in which the conventional wire windings of a cylindrical armature have been replaced by printed circuits. Printed-circuit motors may be operated from one revolution per day up to several thousand revolutions per minute, and may be designed to deliver torques from a few ounce-inches to hundreds of pound-feet. Since the motor is basically of pancake form, it is well adapted to applications where the motor would be integral with the driven device. Because of its design, there is no magnetic attraction between the rotating armature and the stationary portion of the machine.

An exploded view of a motor, manufactured by Printed Motors, Inc., is shown in Fig. 3-89A. At the left is a flux plate, a portion of the housing supporting the brushes which make contact with the 132 segments shown in the center. A similar winding is printed on the reverse side, and connected to the front side by plated through holes. At the right is the opposite portion of the housing, with a permanent magnet stator.

An important aspect of the printed motor is the fact that the printed armature is a single-wave winding. As a result, the operating voltage is low in comparison to conventional motors. For an armature diameter of 3.5 to 8.5 inches, the terminal voltage will range from 6 to 150 volts dc. Consequently they are suited to control circuits employing semiconductor devices.

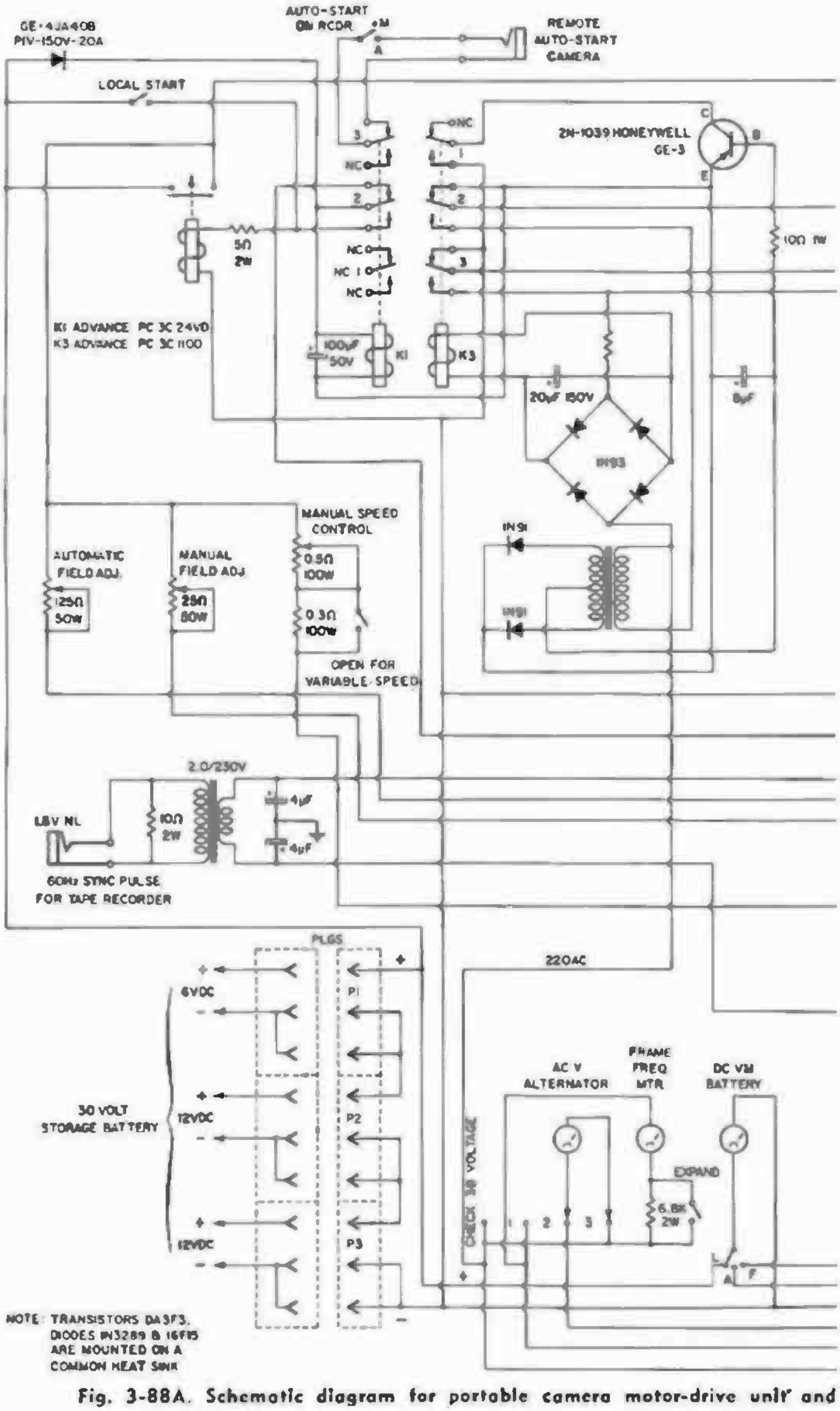
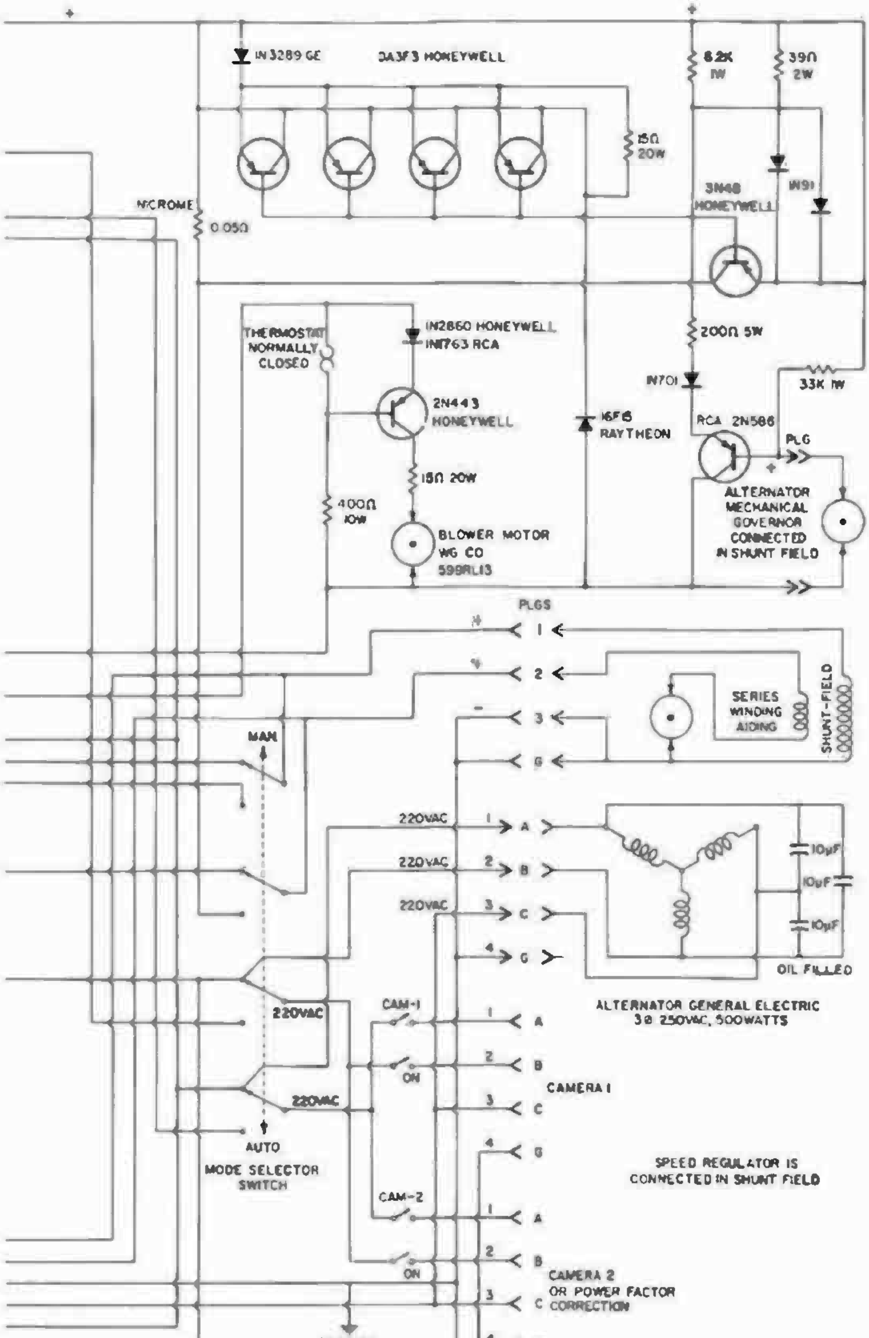


Fig. 3-88A. Schematic diagram for portable camera motor-drive unit and

The armature consists of a disc printed and electroplated with the pattern of conductors rotating in a planar air gap permanent-magnet structure. Commutation is provided by the use of brushes bearing directly on the flat surfaces of the armature segments. Since the effective number of commutator bars is large and the rotating armature contains no iron, the motor provides a smooth modulation-free output torque in response to a constant input current.



250-volt, 3-phase 500-watt alternator. Manufactured by Magnetic Sales Corp.

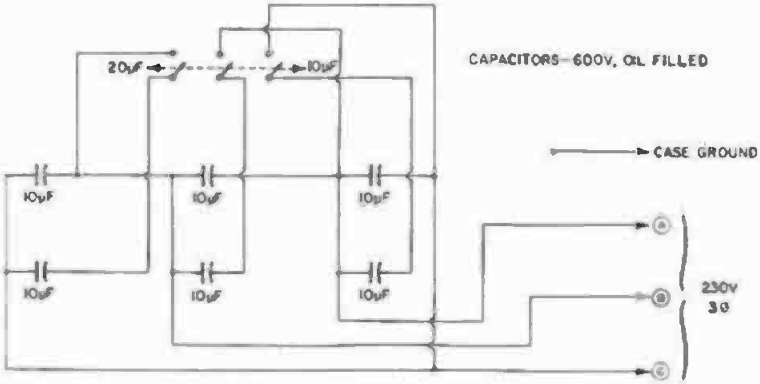


Fig. 3-88B. Camera motor power-supply capacitor bank for correcting power factor. Switch open, two cameras. Switch closed, three cameras. In some instances all the capacitance may be required for two cameras.

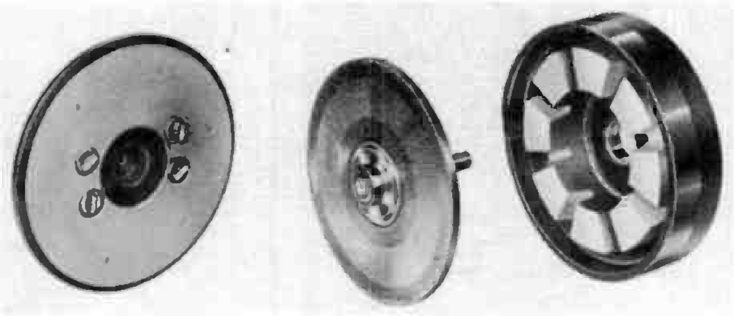


Fig. 3-89A. Exploded view of a printed-circuit motor. At the left is the housing cover with the brushes, armature, and permanent-magnet stator. (Courtesy, Printed Motors, Inc.)

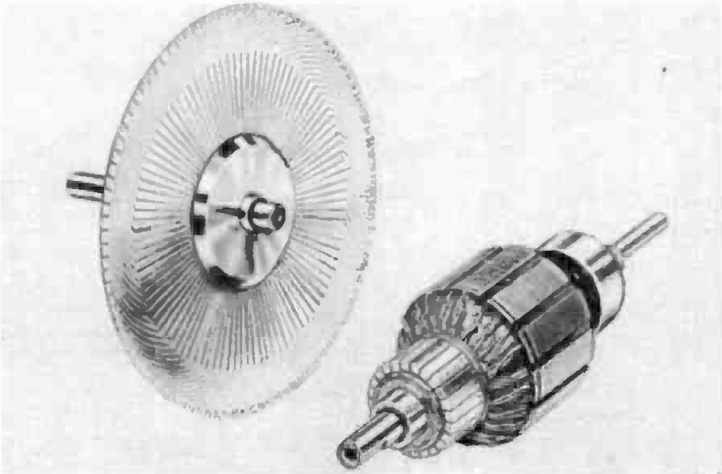
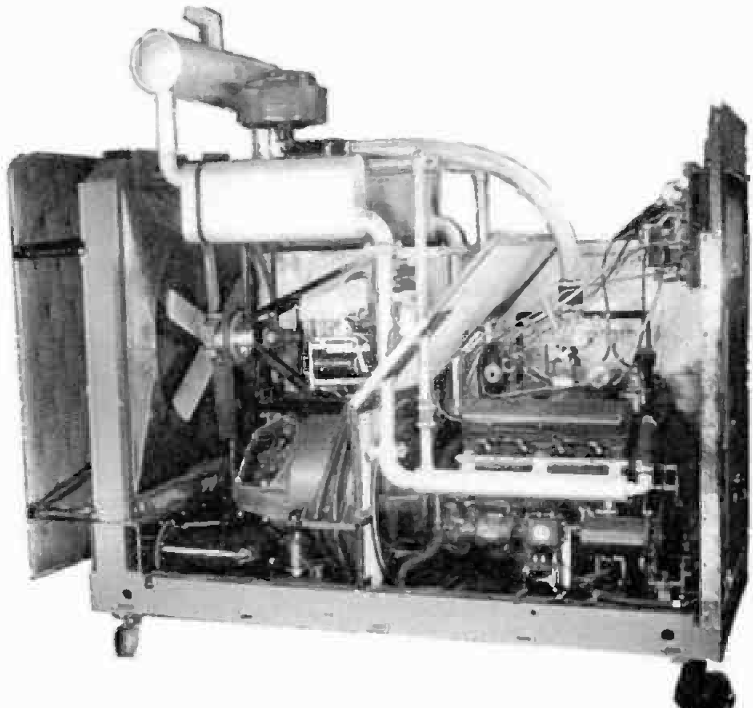


Fig. 3-89B. Printed-circuit motor armature at left and the conventional armature it replaces. (Courtesy, Printed Motors, Inc.)

A printed-circuit armature, compared to a conventional armature it replaces is shown in Fig. 3-89B. Since the magnetic field in a printed motor is supplied by a structure which does not rotate with the shaft, the rotation of the armature does not tend to modulate or vary the permeability of the magnetic path. Due to the design, cogging has been eliminated and there is the-

(a) *View of exterior.*



(b) *View of interior.*

Fig. 3-90. Model-700 portable dc generator for motion picture set lighting. Manufactured by Mole-Richardson. Generator is capable of developing 650 amperes at 120 Vdc.

oretically no limit on the angular accuracy.

The lack of armature inductance displayed by a printed motor is significant in applications where fast response is desired. If the mechanical time-constant is defined as the time for the machine to complete 63 percent of the velocity transition resulting from a step change in terminal voltage, printed motors lie between 0.010 and 0.050 second. Printed motors are best suited to applications involving direct drive of loads at speeds in the range from zero to several thousand rpm, where a premium is placed on the smoothness of generated torque, freedom from preferred armature position, fast response, and high accuracy.

3.90 Describe the construction of a portable dc generator, for motion picture set lighting.—Although portable lighting plants are under the jurisdiction of the electrical department in a motion-picture studio, in small organizations, very often the sound engineer is responsible for them; therefore, a basic knowledge of their design is useful.

Generating plants designed for motion picture use are generally portable either as a unit that may be loaded on a flat-bed truck, or built on a truck as an integral unit. Generally the plant consists of a gasoline motor driving a dc generator with automatic voltage control. A portable plant, manufactured by Mole-Richardson, is shown in Fig. 3-90. This plant is capable of generating 650 amperes at 120 Vdc. The generator is driven by a V-8 Cadillac gasoline engine. Depending on the electrical load demands, the output voltage may be controlled either manually by means of a rheostat, or automatically by a voltage regulator which controls the speed of the engine. The ripple voltage from the commutator is less than 0.5 percent; thus, the noise generated by arc lamps is at a minimum. Such power plants may be used for lighting either incandescent or arc lamps.

The silencing of a generator for motion picture use is a compromise between portability and the degree of noise reduction required. The housing wall of this particular plant consists of 20-gauge sheet metal outer skin, with 3M Co. undercoating applied to the inner surfaces. A fibrous asbestos is sprayed over the undercoating to form an additional sound absorbing layer about ¾-inch thick. This is protected by two coats of casein base paint and metal hardware cloth. The bottom of the plant is closed with covers consisting of ½-inch *Celotex* set between 18-gauge steel sheets. An acoustical partition within the housing, made of ½-inch *Celotex* faced on both sides by ⅛-inch *Transite*, prevents the engine noise from escaping through the radiator. All access doors are gasketed. A blanketed baffle spaced a short distance in front of the radiator reduces air and fan noises. The engine exhaust is muffled by a series-parallel system of silencers. One 3-pass muffler is connected to the exhaust of each 4-cylinder bank of engine, with the outputs of the mufflers joined at the input to a third muffler.

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Section 4

Microphones

Since the invention of the microphone by Dr. Alexander Graham Bell, many types and designs of microphones have been developed over the years, each having served well until its replacement.

Microphones may be classified according to their physical design, such as carbon, capacitor, ribbon-velocity, moving-coil, semiconductor, crystal, and ceramic. They also may be classified according to their polar patterns as omnidirectional, bidirectional, directional, polydirectional, superdirectional, and cardioid. Nomenclature given to microphones designed for special use, such as wireless, dual-stereophonic, in-line, and high-intensity, might be considered yet another category. All of these microphones have been covered in this section, and also their electronics, phasing, and such ancillary equipment as cables, wind and rain screens, and concentrators are discussed in this section. In some instances, the description of a particular microphone is only basic since the design will vary with different manufacturers.

4.1 What is a microphone?—A transducing device which converts acoustical energy into electrical energy. Also called an electroacoustic transducer.

4.2 What are the basic principles of microphone operation?—Microphones are divided into two categories of operation, velocity and pressure. Pressure-operated microphones employ a diaphragm with only one surface exposed to the sound source. The displacement of the diaphragm is proportional to the instantaneous pressure of the sound wave. At lower frequencies such microphones are practically nondirectional.

A velocity microphone is one in which the electrical output substantially corresponds to the instantaneous particle velocity in the impressed sound wave. A velocity microphone is also referred to as a gradient microphone. A gradient microphone is a microphone in which the output corresponds to the gradient of the sound pressure. The quality of a microphone can be judged by the frequency response, sensitivity, distortion, internal noise, and field pattern.

4.3 What are the terms and equations associated with microphones?—Average threshold of the human ear $= 0$-dB SPL (sound pressure level) $= 0.0002$ dynes/cm^2 $= 0.0002$ microbar $= 10^{-16}$ watt/cm^2 (free field in air) $= 10^{-13}$ watts/ft^2.

$$1 \text{ microbar} = 1 \text{ dyne/cm}^2 = 10^{-6} \text{ bar} = 74\text{-dB SPL}$$

$$1 \text{ atmosphere} = 1 \text{ bar} = 14.5 \text{ lb/in}^2 = 10^6 \text{ dynes/cm}^2 = 194\text{-dB SPL.}$$

Microphone sensitivity is defined as output voltage in dB re: 1 volt for a sound pressure of 1 dyne/cm^2, or 74-dB SPL. As an example: the sensitivity rating for a given microphone of minus 85 dB re: 1 volt/dyne/cm^2 states; the output voltage is 85 dB below 1 volt for 74-dB SPL; for 174-dB SPL, the output voltage would be $174 - 74 + (-85) = +15$ dB re: 1 volt. An output voltage 15 dB above 1 volt equals 5.6 volts.

4.4 What types of microphones are classed as pressure operated?—The carbon, crystal, dynamic, pressure, capacitor, and frequency-modulated.

4.5 What type of microphone is classed as velocity operated?—The ribbon. The electrical response corresponds to the particle velocity.

4.6 What is a pressure gradient microphone?—A ribbon velocity microphone such as described in Question 4.50.

4.7 What is a polar field pattern?—It is a plot employing polar coordinates, to show the magnitude of a quality in some or all directions from a given point for three hundred sixty degrees. Polar plots are used to present the directional pattern of microphones, loudspeakers, and other devices having directional characteristics. The term field pattern is also used to denote directional qualities; however, either term may be used interchangeably. In some instances the term directional pattern or characteristic is used. The term free-field denotes the device under measurement is in space, in an area free from obstructions and reflections, such as would be found in open air or in an anechoic chamber. Polar plots are also used in the measurement of magnetic fields and many electrical devices not having acoustical properties. Typical polar plots or field patterns are shown in Fig. 4-57.

At (a) is an omnidirectional or circular or nondirectional field pattern for the crystal, dynamic (moving coil), capacitor (condenser), carbon, electronic, frequency modulated, and inductor type microphones.

At (b), a semidirectional pattern obtained with an adjustable field pattern microphone is illustrated. The microphone is directional at the higher frequencies but nondirectional at the low frequencies.

The bidirectional pattern obtained with a ribbon microphone is shown at (c). The microphone is essentially dead to pickup at the sides. This pattern is generally referred to as a figure-8 field pattern.

The pattern for a unidirectional microphone, called a cardioid pattern, is shown at (d). Microphones are available that will permit the field pattern to be varied to fit almost any situation and include all of the foregoing patterns in some form or other.

4.8 How is a carbon microphone constructed and what is its principle of operation?—Several hundred small carbon granules are held in close contact in a brass cup called a button, which is attached to the center of a metallic diaphragm. Sound waves striking the surface of the diaphragm disturb the carbon granules, thus changing the contact resistance between their surfaces. The change in contact causes a current from a battery connected in series with the carbon button and the primary of a transformer to vary in amplitude, resulting in a current waveform similar to the acoustic waveform striking the diaphragm. After leaving the secondary of the transformer, the minute changes of current through the transformer primary are amplified and reproduced in the conventional manner. The circuit diagram and construction of a carbon microphone are shown in Fig. 4-8. The output voltage from a carbon or pressure microphone is proportional to the displacement of the diaphragm. The field pattern is omnidirectional. This is shown in (a) of Fig. 4-7.

One of the principal disadvantages of the carbon microphone is that it has continuous high-frequency hiss caused

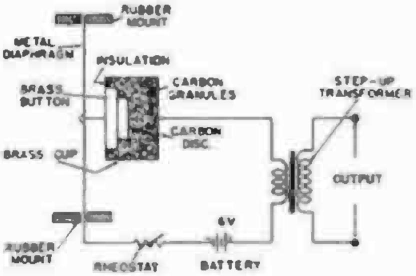
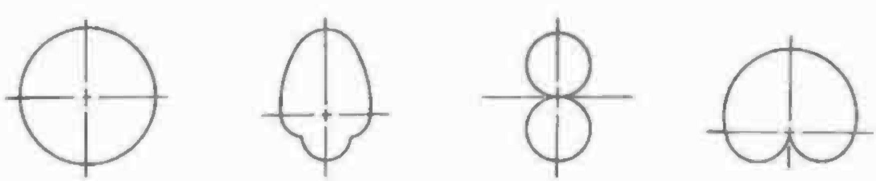


Fig. 4-8. Connection and construction of a single-button carbon microphone.



(a) Omnidirectional. *(b) Semidirectional.* *(c) Bidirectional.* *(d) Cardioid.*

Fig. 4-7. Basic microphone field patterns.

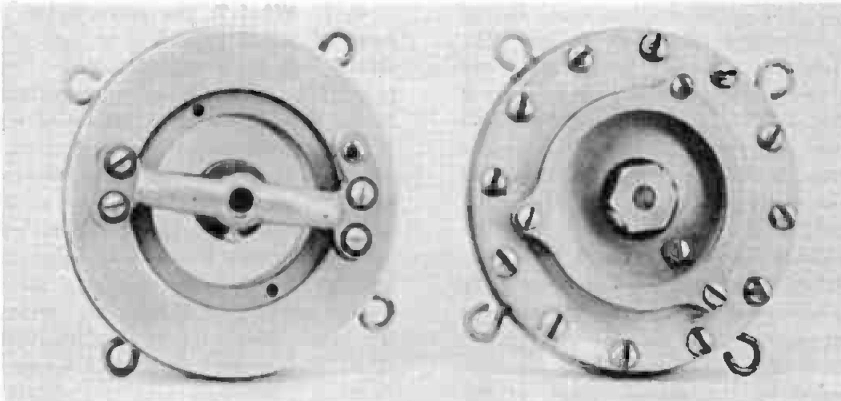


Fig. 4-9B. Early model Western Electric type 600A, double-button, stretched-diaphragm, carbon microphone.

by the changing contact resistance between the carbon granules. In addition, the frequency response is limited and the distortion is rather high.

4.9 ***What is a double-button carbon microphone and how does it function?***— The double-button microphone employs two carbon buttons similar to those used in the single-button type. One button is mounted on each side of the diaphragm. Pressure waves striking the surface of the stretched diaphragm cause it to move and disturb the contact resistance of the carbon granules in the buttons in a manner similar to the single-button microphone described in Question 4.8. As the diaphragm moves, the contact resistance of the granules in the button mounted on the pressure-wave side is reduced, while the resistance of the button on the opposite side is increased. When the pressure wave reverses itself, the reverse action takes place in the carbon buttons. Thus, the current through the buttons corresponds to each half of the pressure wave at the diaphragm. This action is somewhat similar to that of a push-pull amplifier stage. The circuit connections and construction are shown in Fig. 4-9A. The exterior appearance of an early model double-button carbon microphone, manufactured by Western Electric Co., is shown in Fig. 4-9B.

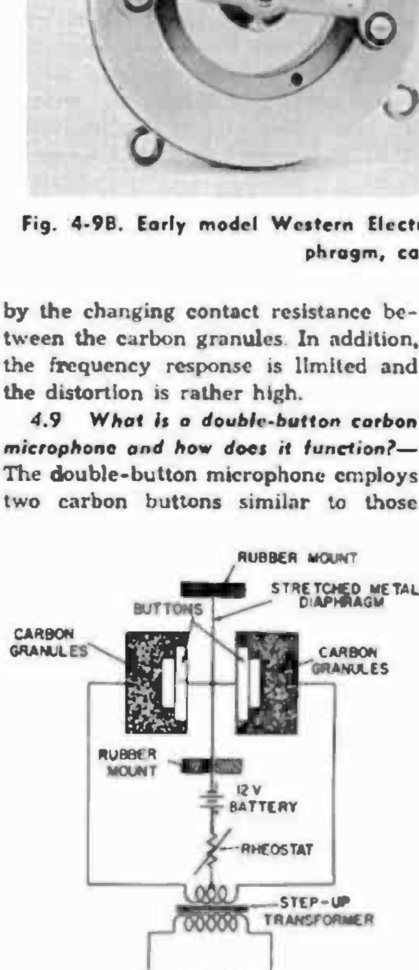


Fig. 4-9A. Connections and construction of a double-button carbon microphone.

The disadvantages of the double-button microphone are about the same

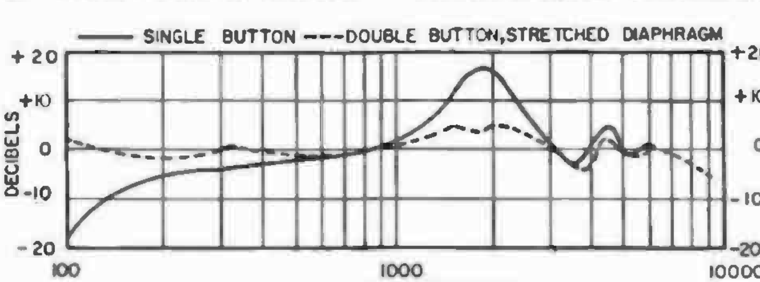


Fig. 4-9C. Frequency response of early model single- and double-button carbon microphones.

as for the single-button type, except the waveform distortion is less. The frequency responses for single- and double-button microphones are given in Fig. 4-9C. The solid line is the single button and the dotted line the double button. It will be noted the resonant peaks of the diaphragm have been reduced in the double-button type by stretching the diaphragm.

4.10 What precautions should be observed when using carbon button-type microphones?—The current through the buttons should not exceed that recommended by the manufacturer, or the carbon granules may be fused. If the microphone is of the double-button type, the currents through each button must be the same when the diaphragm is at rest. Carbon microphones should not be subjected to heavy jars when the current is flowing, unless they are designed for such service.

4.11 What causes packing of the carbon granules in a carbon microphone?—Excessive current through the buttons and moving the microphone when the current is flowing. However, carbon microphones used in communications work are designed to be moved while there is current through them.

4.12 Describe a crystal microphone.—A microphone employing one or more Rochelle salt crystals placed in such a manner that when their surfaces are struck by a pressure wave they are bent or twisted out of shape. This action results in the generation of an electrical current because of the piezoelectric effect of such crystals. A typical crystal microphone is shown in Fig. 4-12.

Fig. 4-12. A typical crystal microphone.

4.13 What is the piezoelectric effect?—When a crystal is subjected to strain, electrical polarization takes place. The polarization is proportional to the mechanical strain and sine with it. The inverse effect is produced when electrical current is applied to the crystal. The mechanical movement in this case is proportional to the applied current. Advantage is taken of this characteristic in the design of crystal microphones, pickups, speakers, and recording heads. The subject of piezoelectric effect is discussed further in Question 25.191.

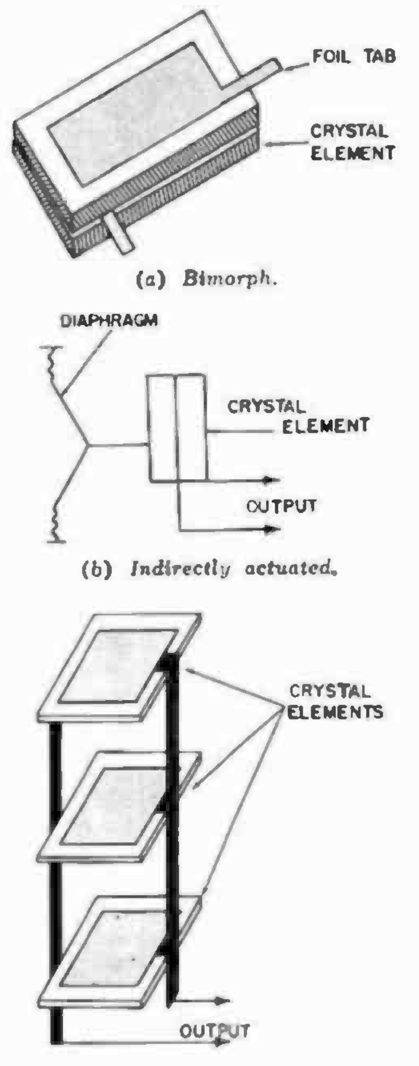
4.14 What type crystals are used in crystal microphones?—Rochelle salt crystals. They are grown from a supersaturated solution of sodium potassium tartrate tetrahydrate by cooling at a temperature of 40°C. Such crystals should not be operated or stored in temperatures exceeding 140° F. Crystals of this nature should not be confused with the quartz crystals used in radio transmitters. (See Questions 25.102 and 25.103.)

4.15 What type crystal microphones are in general use?—The direct-actuated, Fig. 4-15, part (a) and the indirectly actuated, Fig. 4-15, part (b). In the former type, the sound waves strike the surfaces of the crystals creating mechanical strain. In the latter, the sound waves impinge on a diaphragm attached mechanically to the crystal elements.

4.16 What is a sound-cell crystal microphone?—One in which a number of crystal elements are stacked in a pile, as shown in Fig. 4-15, part (c).

4.17 What is a Bimorph crystal?—A type of construction used in crystal microphones to increase their sensitivity. The crystal consists of two slabs cut on axes which determine whether they are to be benders or twisters. The two slabs are separated by a thin piece of foil which connects to one side of the external circuit. The outer surfaces of the crystal slabs are covered with foil and connected to the other side of the external circuit as shown in Fig. 4-15, part (a). The name *Bimorph* is a trade name of the Clevite Corp., Piezoelectric Div.

4.18 What is a bender element?—A crystal element cut on an axis which results in a piezoelectric effect only as a result of bending.



(a) *Bimorph.*

(b) *Indirectly actuated.*

(c) *Sound cell.*

Fig. 4-15. Direct and indirect actuated crystal elements used in microphone construction.

4.19 What is a twister element?—A crystal element cut on an axis which results in a piezoelectric effect only as a result of twisting.

4.20 What is the field pattern for a crystal microphone?—Circular, as for all pressure microphones. See Fig. 4-7, part (a).

4.21 What is the output impedance of a crystal microphone?—Generally on the order of 60,000 to 100,000 ohms.

4.22 Which type crystal microphone has the greatest output, the diaphragm or the sound cell?—The diaphragm type has the greatest output voltage, but the frequency response is limited because of the construction. The sound cell, Fig. 4-15, part (c) has the best frequency response but the lowest output level.

4.23 How is the greatest output voltage obtained from a crystal microphone with the lowest output impedance?—By connecting the crystal elements in series-parallel.

4.24 What is the maximum length of cable that may be used with a crystal microphone?—Approximately 50 feet. Because of the high impedance, a crystal microphone must not be separated too far from the input of the amplifier. The capacity of the cable acts as a capacitor in parallel with the crystal elements. A long cable attenuates the over-all level but has little effect on the frequency response.

4.24 What is the internal capacitance of a crystal microphone?—Generally on the order of 0.03 uF, for the diaphragm-actuated type. For the sound-cell type, 0.0005 to 0.015 uF.

4.26 What is the frequency response of a crystal microphone?—About 80 to 6500 Hz for the diaphragm type, with a peak occurring around 3500 Hz. High-quality crystal microphones will respond up to 16,000 Hz; however, they show a rise in the higher frequencies necessitating equalization to obtain a uniform frequency response. The frequency response of a typical high-quality crystal microphone of the sound-cell type is shown in Fig. 4-26.

4.27 What type amplifier input circuit is recommended for crystal microphones?—Resistance-coupled similar to that shown in Fig. 4-27. Because a crystal microphone appears as a capacitance, it may be considered to be in series with the grid resistance of the input stage. The grid resistance (R_g) should be from 3 to 5 megohms for a microphone of the sound-cell type and from 1 to 5 megohms for the diaphragm type. Lowering the resistance in the grid circuit attenuates the low frequencies.

4.28 What types of microphones are considered to be high impedance?—Crystal, ceramic and the head of a capacitor microphone (condenser) are included in this category. The ceramic microphone may be considered from an impedance standpoint to be similar to the crystal microphone. The head of a capacitor microphone is of extremely

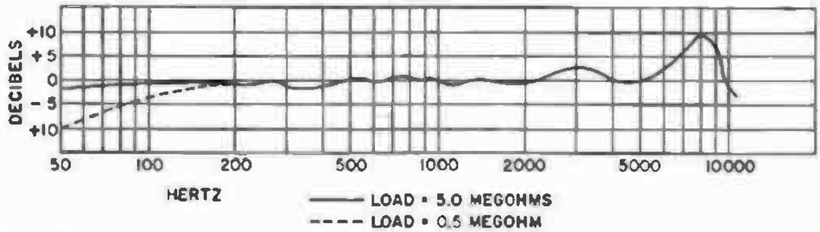


Fig. 4-26. Frequency response of a high-quality sound-cell type crystal microphone.

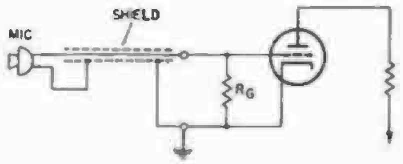


Fig. 4-27. Method of coupling a crystal microphone to the input of an amplier.

high impedance, running into several megohms; therefore, it must be operated adjacent to a preamplifier. An output transformer in the amplifier supplies a low output impedance, ranging from 50 to 250 ohms. Dynamic microphones may be obtained, with impedance ranging from 50 to 40,000-ohms.

4.29 Describe a ceramic microphone?—A microphone similar in characteristics to the crystal, except that it employs a barium-titanate slab in the form of a ceramic. It has piezoelectric characteristics and may be operated in higher temperatures and humidity. Generally speaking a ceramic microphone may be used wherever a crystal microphone is employed; however, the equalization may have to be revised slightly. Ceramic microphones, like the crystal microphone, must be operated into an input impedance from 1 to 5 megohms. Such microphones are ideal for walkie-talkies, hearing aids, dictating machines, public-address systems, and many other services. Typical output levels are minus 59 to 50 dB, where zero dB equals 1 volt/microbar, with an output impedance of 100,000 ohms.

4.30 Describe the construction and characteristics of a dynamic microphone.—A dynamic microphone employs a small diaphragm and a voice coil, similar to a dynamic loudspeaker, moving in an intense permanent magnetic field. Sound waves striking the surface of the diaphragm cause the coil to be moved in the magnetic field, thus generating a voltage proportional to the sound pressure at the surface of the diaphragm. This microphone is also referred to as a pressure or moving-coil microphone (nicknamed "eight ball" because of its appearance). Typical examples of this design are the Western Electric 630B, shown in Fig. 4-30A, and the Altec-Lansing Model 633a/c (nicknamed "salt shaker"), shown in Fig. 4-30B. Due to their similarities, the following explanation will suffice for both types, although the case and frequency response of the Atlec-Lansing 633a/c is somewhat different than the Western Electric 630B.

Shown in Fig. 4-30C is a cross-sectional view of the 630B. The diaphragm (A) is of Dural, approximately 0.5-mil in thickness, and it weighs 25 milligrams. Cemented to the rear of the diaphragm is a voice coil (B) constructed of edge-wound Dural ribbon 8 mils thick × 8 mils wide. The body of the microphone (C) consists of a molded spherical housing, containing a permanent magnet (D) with a center pole piece over which the voice coil is centered. The edge of the diaphragm is hinged and supported at the edges by the housing. The outer surface of the diaphragm is protected from mechanical injury by a perforated grid (E). A two-layer 16-mesh circular screen baffle (F) with layers of silk between the screens is placed in front of the diaphragm. The perforated grid and screen act as acoustical equalizers to improve the omnidirectional characteristics. At the lower left is a small metal tube (G) termed an acoustical equalizer. Its function is to release air pressure behind the diaphragm to prevent distortion of the diaphragm during its inward travel. Working in conjunction with this tube are two air-release vents (H) under the voice coil, to provide acoustical resistance. External pins (I) provide connections to the voice-coil leads (J).

Fig. 4-30A. Western Electric 630B dynamic (moving-coil) microphone.

Microphones of this type do not employ an output transformer; the output voltage is taken directly from the voice-coil winding. The frequency response of this microphone is little affected by the angle of incidence up to 120 degrees as may be seen in Fig. 4-30D. The output impedance is 20 ohms, but is operated into a 30- to 50-ohm preamplifier. The polar pattern is omnidirectional, with a frequency response as shown in Fig. 4-30D. The output level is minus 55 db/10 dynes/cm².

Fig. 4-30B. Altec-Lansing Model 633a/c dynamic (moving-coil) microphone. The 8b baffle is used to assist in achieving a directional characteristic.

4.31 What is a pressure microphone?—It is a dynamic microphone

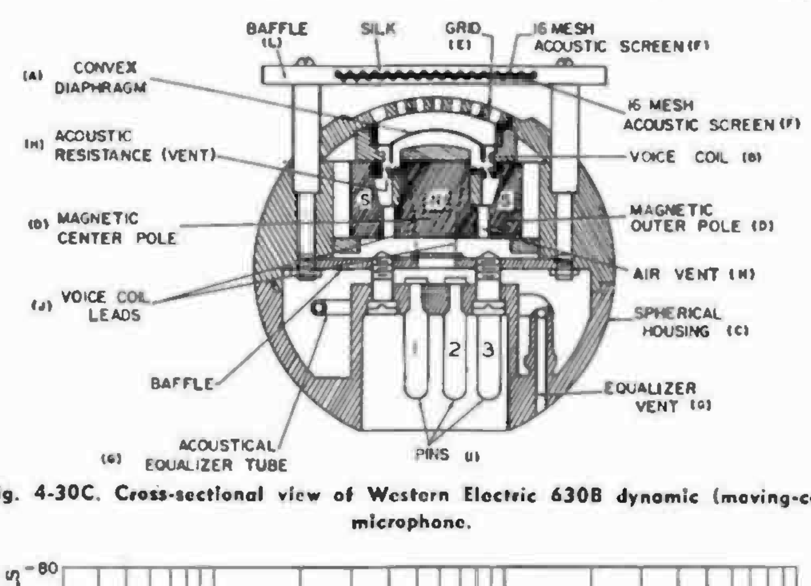


Fig. 4-30C. Cross-sectional view of Western Electric 630B dynamic (moving-coil) microphone.

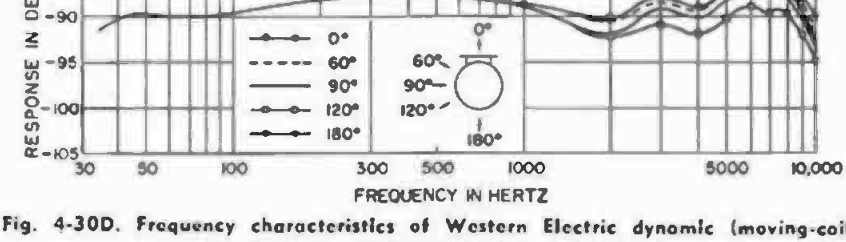


Fig. 4-30D. Frequency characteristics of Western Electric dynamic (moving-coil) microphone for different angles of incidence.

Fig. 4-31. American Microphone Co., type D-22 pressure (moving-coil) microphone with windscreen and boom hanger.

similar to that described in Question 4.30. A typical pressure microphone manufactured by American Microphone is pictured in Fig. 4-31, with a wind screen and hanger for fish-pole operation. The construction of the wind screen is discussed in Question 4.90.

4.32 What is a moving-coil microphone?—A dynamic or pressure microphone such as the ones described in Questions 4.30 and 4.31.

4.33 What is an acoustical equalizer?—A small metal tube located at the rear of a dynamic microphone to release the pressure behind the diaphragm and thus prevent distortion of the diaphragm. The acoustic equalizer (see Fig. 4-30) also tends to smooth out the frequency characteristic of a dynamic microphone.

4.34 What is the field pattern of a dynamic microphone?—Omnidirectional. Some directional effect may be obtained by hanging the microphone overhead with the diaphragm downward; however, this will attenuate the high frequencies to some extent.

4.35 Show an exploded view of a dynamic microphone. — An exploded view of an Electro-Voice Model 635A dynamic microphone is shown in Fig. 4-35. Starting at the left, is a protective screen (A) which combined with elements (B) and (C) compromises a four-stage pop-filter for close talking. Element (B) consists of a special *Acoustifoam* material developed by the manufacturer. Element (D) is a fine screen of magnetic material for attracting metal dust particles to prevent them from falling on the diaphragm and its associated magnetic assembly (F) and (G). The plate (E), with the diaphragm assembly, forms a Helmholtz resonator in front of the diaphragm (F). The inertance of the holes in the plate resonate with the compliance of the air between the retainer and the diaphragm. The moving elements consist of an *Acousticalloy* diaphragm (a special material developed by Electro-Voice) which is quite stable to pressure and temperature changes. A voice coil mounted at the rear of the diaphragm is centered in a strong permanent magnetic field (G).

As this microphone is designed to be hand-held if desired, mechanical shock must be reduced to an absolute minimum by isolating the voltage-generating elements from the case. This is accomplished by the use of a low-durometer plastisol sleeve (H). If the isolation between the case and the moving

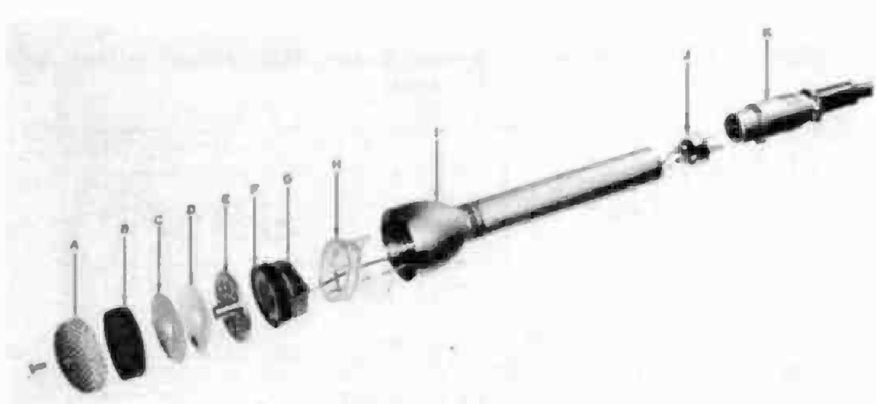


Fig. 4-35. Exploded view of Electro-Voice Model 635A dynamic microphone.

elements is not sufficient, vibrations will be transmitted to the moving coil and diaphragm assembly, resulting in unwanted noise.

The diaphragm is ⅞ inch in diameter and 0.0015 inch in thickness. No output transformer is used, as the moving coil is wound using a number 48 copper wire for an impedance of 150 ohms. Since there is only one opening to the diaphragm, the polar pattern is omnidirectional. This system of controlled leakage allows for pressure equalization on both sides of the diaphragm. The high-frequency response is held up by the use of the Helmholtz resonator (E).

The last three elements are the case (I), the three-pin male connector (J), and the female cable plug (K). The frequency response is 60 to 15,000 Hz, with an output level of minus 55 dB re: 1 mw/10 dynes/cm^2. Microphones of this type are used for professional and public-address work, and in television and broadcast studios.

Since the advent of rock and roll recording where the artist sings or shouts directly into the microphone one or two inches from the mouth, special blast-type microphones have been developed. This has been necessary because of the extremely high sound-pressure levels (SPL) distorting the diaphragm. Special precautions must also be taken to prevent the overloading of the microphone preamplifier input circuit. (See Question 9.51.)

4.36 What precautions must be taken when using pressure-type microphones with an omnidirectional polar pattern?—A characteristic of pressure microphones is their lack of discrimination to low-frequency sounds arriving from random directions. Sounds which originate at the rear of the microphone will be bent around the microphone housing and actuate the diaphragm as if they had arrived from the front. This is especially true of sounds whose wavelengths approach the dimensions of the microphone housing. At the higher frequencies, for sounds originating at its rear, the frequency response will drop off due to diffraction. When recording dialogue with this type of microphone, it may be desirable to increase the directional qualities by the use of a small baffle on the front of the microphone in place of the small screen baffle normally employed. The center hole of the substitute baffle is covered with a single layer of thin silk. The use of a baffle with this type of microphone not only increases its directional qualities, but also increases the response at the lower frequencies. The increase in low-frequency response may be desirable for certain types of pickups but not for dialogue. Therefore, the low-frequency attenuation normally used for dialogue recording may have to be increased.

4.37 How should a pressure microphone be used suspended from a boom?—The diaphragm is tilted downward with the diaphragm almost parallel to the floor. One reason for this suspension is that pressure-type microphones are highly efficient at the higher frequencies and any small loss of high frequency response is not detrimental to the pickup. A second reason is that for large angles of incidence, the directional response changes less rapidly than for small angles of incidence. Thus, a more uniform quality of pickup may be maintained.

4.38 What size baffle is recommended for use with the microphone discussed in question 4.30?—The baffle should be 3⅛ inches in diameter with a 1-inch hole in the center. The material may be ⅛-inch *Bakelite* or Dural.

4.39 Describe a dynamic microphone with cardioid characteristics and a variable-frequency response. — Many times it is necessary to intermix microphones, which generally poses a problem because of the difference in frequency response. The ideal microphone would be one that would permit its frequency response to be varied so as to match other microphones, while retaining its directional characteristics. The

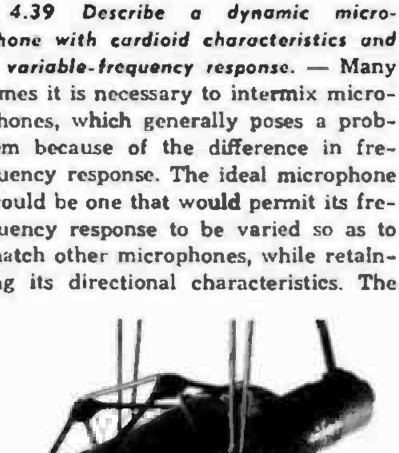


Fig. 4-39A. Electro-Voice Model 668 dynamic cardioid microphone mounted in its wind screen and boom hanger.

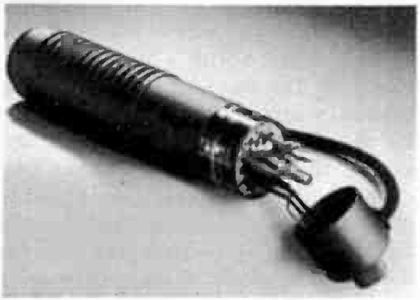


Fig. 4-39B. Electro-Voice dynamic cardioid microphone Model 668. The basic design is of the in-line type. Thirty-six different frequency-response curves are possible by the connections in the base.

microphone to be described is of the in-line family, and has been developed especially for this purpose by Electro-Voice. Complete with wind screen and boom hanger (Fig. 4-39A), it is light enough to be "fish-poled." The microphone proper is shown in Fig. 4-39B, with a cross-sectional view of the wind-screen and hanger construction given in Fig. 4-39C.

Basically the microphone is an in-line type, with two slotted tubes coupled to the back of the diaphragm. The acoustic length of the tube varies inversely with the frequency of the sound source, permitting the phasing-out of unwanted sound from all portions of the spectrum for a maximum front to back ratio. The transducer consists of a single moving-coil element. Contained in the base is a passive equalizer network, which provides a choice of three variations in the high-frequency response (A) (B) and (C) in Fig. 4-39D, and a choice of three variations in low-frequency response identified as (1) (2) and (3). A filter network for reduction of the response above 8000 Hz and below 80 Hz, having a 50 dB per octave cutoff rate, is also provided. The output level is minus 51 dB re: 1 mw/10

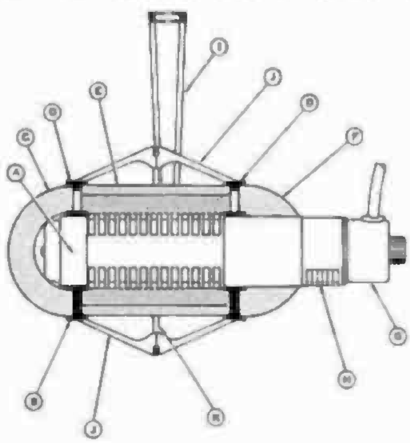


Fig. 4-39C. Cross-sectional view of Electro-Voice Model 668 dynamic cardioid microphone, with its wind screen.

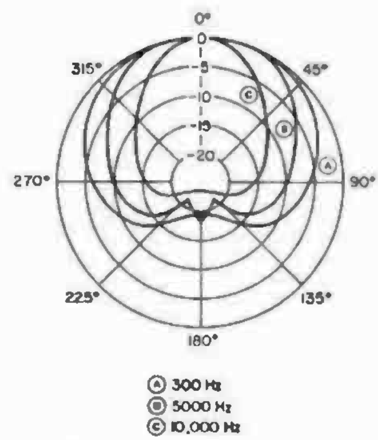


Fig. 4-39E. Polar pattern for Electro-Voice Model 668 dynamic cardioid microphone.

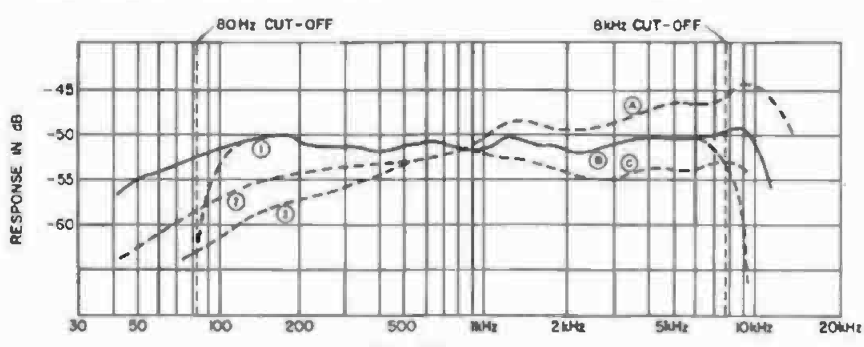


Fig. 4-39D. Frequency response for Electro-Voice Model 668 dynamic cardioid microphone.

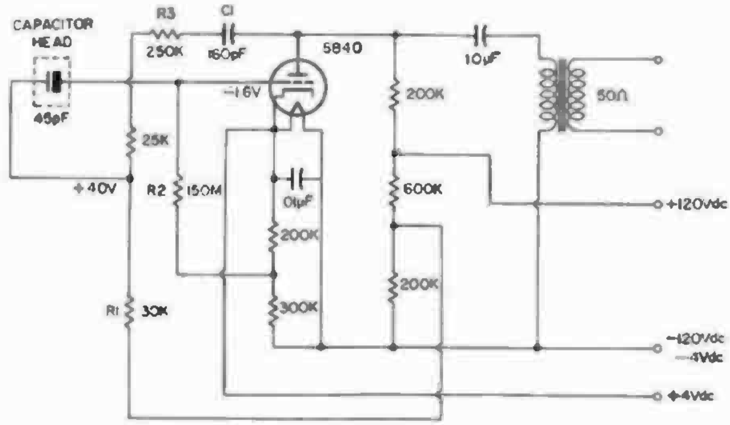


Fig. 4-40A. Typical preamplifier circuit for a capacitor microphone using a single vacuum tube.

dynes/cm². The polar pattern, Fig. 4-39E, is unaffected by the frequency response changes.

Referring back to Fig. 4-39C, at (A) is the microphone. The in-line port may be seen at (B), and the three sections of the wind screen at (C) (E) and (F), held in place by shock-rings (D). A cover (G) encloses a group of terminals for setting the desired frequency response and output impedance. Plate (H) shows graphically the frequency response possible by the plugging-in of the phone tips seen in Fig. 4-39B. The whole assembly is supported by rubber cords (J), held in place by steel rings (K). The hanger is suspended from a boom turret-head by a thumbscrew.

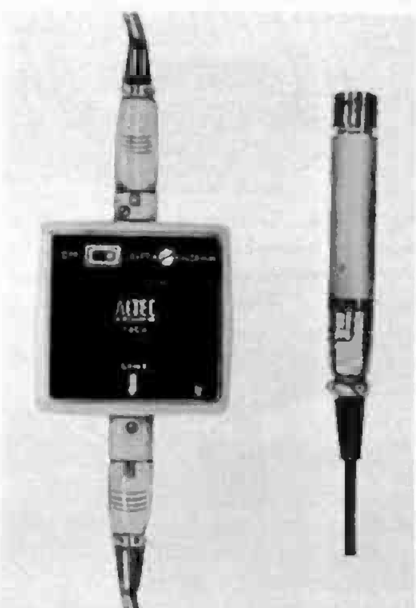


Fig. 4-40B. Altec Model 29B solid-state microphone system.

4.40 Describe the basic principles of a capacitor (condenser) microphone.—The head of a capacitor microphone consists of a small two-plate capacitor, about 40 to 50 pF in capacitance. One of the two plates is a stretched diaphragm; the other is a heavy back plate or center terminal (Fig. 4-41A). The back plate is insulated from the diaphragm and spaced approximately 0.001 inch from, and parallel to, the rear surface of the diaphragm. Mathematically, the output from the head may be calculated:

$$\frac{E_{oc}}{P} = \frac{E_o a^2}{8\,dt}$$

where,

P is the pressure in dynes per square centimeter,
t is the diaphragm tension in dynes per centimeter,
a is the radius of active area of the diaphragm in centimeters,
d is the spacing between the back plate and diaphragm in centimeters,
E_o is the dc polarizing voltage in volts.

If referred to a reference level of 1 volt:

$$dB = 20\text{Log}_{10}\frac{P}{E_{o}C} = 20\text{Log}_{10}\frac{8\,dt}{E_o a^2}$$

Fig. 4-40A is a schematic diagram of a typical preamplifier used with a capacitor microphone. Here, the back plate is connected to the control grid of the tube, and the diaphragm polarized at 40 Vdc, through a 30k resistor, R1. Thus a fixed charge accumulates on the diaphragm. As sound waves enter the head cavity through the side openings, the pressure causes minute changes in the spacing of the two head members, thereby varying the internal capacitance. The resulting signal voltage, which is proportional to the pressure wave, is applied to the control grid of the tube, amplified, and passed on through the output transformer.

It will be observed, the lower end of the 150-megohm resistor R2 is returned to a position of plus 2.4Vdc above ground. Also, the cathode is connected to the plus 4 Vdc of the heater circuit. This makes the control-grid minus 1.6 Vdc with respect to the cathode. Negative-feedback elements R3 and C1 reduce the distortion, and aid in the control of the frequency response. In some capacitor-microphone amplifiers, the grid resistor R2 is omitted, since the microphone head is a capacitor, and to obtain a good low-frequency response the shunt resistance must be as high as possible. The source impedance, as seen by the control grid of the tube, consists of capacitance and resistance in parallel. The thermal noise is least when the resistance is the highest. The effective grid resistance is established by the tube itself.

A microphone system, manufactured by Altec-Lansing and illustrated in Fig. 4-40B, consists of a single diaphragm; the directional (cardioid) head is of small size and low mass, for good high frequency and transient response. The cardioid characteristic is obtained by controlling the relative phase of the sound pressure reaching the back side of the diaphragm. This method provides the greatest front-to-back discrimination over a wide frequency range.

The microphone head is mounted on a base containing a field-effect transistor (FET) and the necessary components. The FET converts the extremely high impedance of the microphone capacitor head to a low impedance suitable for connection to a standard two-conductor cable. One conductor of the cable carries both signal current and the direct current for the FET, while the second provides a microphone polarizing potential of 64 volts. The output from the microphone amplifier is unbalanced until it joins the power supply unit which also contains an output transformer.

The power-supply unit is provided with both male and female cable plugs, to permit it to be inserted at any point in the microphone line when long cable runs are being used, thus increasing the signal-to-noise ratio of the signal in the cable. Power is supplied by two 4.2-volt and three 21-volt mercury batteries, with an expected life of about 2500 hours. A meter is provided on the rear of the power supply case for monitoring the batteries. The frequency range of this microphone is 20 to 20,000 Hz, with output level of minus 53 dB re: 1 mw/10 dynes/cm².

4.41 Describe the construction of a capacitor microphone head. — In earlier types of this microphone (Fig. 4-40A), the internal capacity of the head varied from 200 to 400 picofarads, for a diaphragm diameter of 1¾ inches. Present day capacitor microphone heads have an internal capacity of 40 to 50 picofarads with a diaphragm diameter of 0.5 inch. The impedance of such a head is approximately 30 megohms; therefore, it must be operated within a few inches of the amplifier input. Generally, the head and amplifier are assembled as a unit. A cross-sectional view of a head assembly manufactured by Altec-Lansing, is shown in Fig. 4-41A. Capacitor microphones may be designed to have omni, bi, or unidirectional (cardioid) polar characteristics. A capacitor

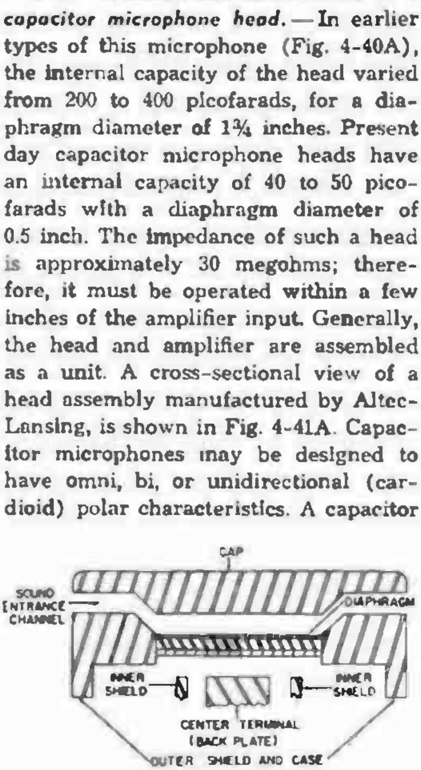


Fig. 4-41A. Cross-sectional view of a capacitor microphone head used with the Altec capacitor microphone pictured in Fig. 4-40B.

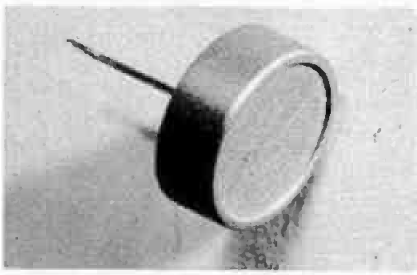


Fig. 4-41B. Capacitor head for Neumann Model U-64 capacitor microphone. (Courtesy, Gotham Audio Corp.)

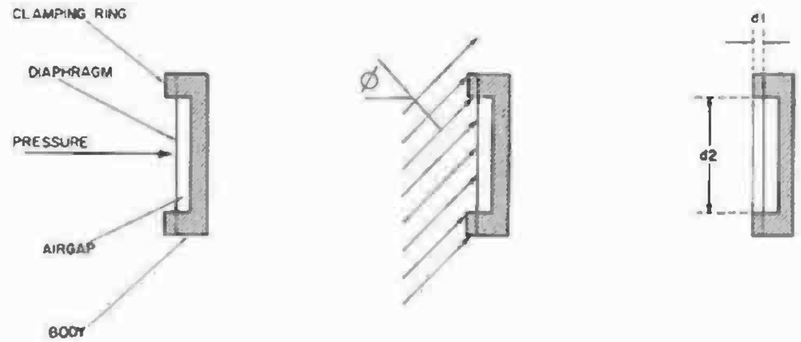
head for the Neumann Model U-64 capacitor microphone is shown in Fig. 4-41B. The diaphragm consists of an extremely thin mylar base, gold plated, less than ¾ inch in diameter. Although the head shown has only a single diaphragm and back plate, capacitor microphone heads designed for directional characteristics employ two diaphragms, with a common polarized back plate centered between the two. (See Question 4.115.)

4.42 ***What is the output level of a capacitor microphone head?***—Approximately minus 90 dBm.

4.43 ***What is pressure doubling in a capacitor microphone?***—It is the result of pressure waves, which are small in comparison to the diaphragm dimensions, coming to a complete stop at the face of the diaphragm. Pressure doubling in earlier types of microphones often caused a peak of 4 to 6 dB in the midrange frequencies. This peak caused extreme sibilance and high-frequency distortion. This defect has been eliminated in present day designs by reducing the physical size of the capacitor head. (See Question 4.41.)

4.44 ***Describe the effects of cavity resonance in a capacitor microphone.***—In the older model capacitor microphones, cavity resonance was quite common because of the large diaphragm (about 2 inches), the design of the case, and the method of stretching the diaphragm. Most of these effects have been eliminated in modern microphones by the reduction in size of the diaphragm to diameters of ¾ inch or less, and by improved methods of stretching the diaphragm. Three common causes of cavity resonance are given in Fig. 4-44. In part (a) are shown the effects of diffraction, which result when the reflected component of the sound-wave incidence upon the surface of the diaphragm causes a standing wave sufficient enough to produce pressure doubling and tripling. Part (b) shows the phase-difference effect, which is caused when the wavelength of the sound wave is comparable to the diameter of the diaphragm and occurs when the incident wave front strikes the diaphragm at an angle theta (θ) to cause pressure variations across the surface of the diaphragm. Part (c) shows cavity resonance caused when the wavelength of the sound wave is comparable to the inside dimensions of the diaphragm assembly. This causes an increase in the sound pressure in proportion to the ratio of the diaphragm d1 to the dimension of the cavity depth d2 (d1/d2).

4.45 ***What are the general characteristics of a capacitor microphone?***—Capacitor microphones have wide frequency response, low distortion, and little internal noise. Although the earlier models were omnidirectional, they may now be obtained with omni, bi, and unidirectional characteristics. A typical frequency response for a capacitor microphone, using an omnidirectional polar pattern is shown in Fig. 4-45. Capacitor microphones require a preamplifier as an integral part of the housing and a source of polarizing voltage for the head, plus a source of power



(a) Diffraction effect causing frequency doubling or tripling.

(b) Phase-differences caused by pressure differences across the face of the diaphragm.

(c) Pressure increases because of the resonance of the cavity area at certain wavelengths.

Fig. 4-44. Cavity effects in front of a capacitor microphone.

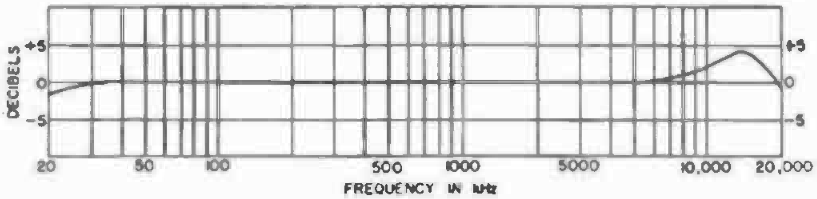


Fig. 4-45. Typical frequency response for a capacitor microphone using a transistor amplifier. Omidirectional pattern.

for the amplifier section, which may be either transistor or vacuum tube. Such microphones, if operated in a high humidity, may develop noise due to moisture getting into the head, and cause arcing between the diaphragm and back plate; however, in present day microphones, this has been almost completely eliminated. If the device is being operated in high humidity, the head should be stored in a desiccator jar when not in use.

4.46 Describe a completely self-contained capacitor microphone.—The microphone shown in Fig. 4-46A, Model S-10, manufactured by Syncron Corp., is completely self-contained and employs a field-effect transistor (FET), operated from a 8.4-volt mercury battery. Fig. 4-46B shows a cross-sectional view of the internal construction. The capacitor head is at (A) with a series of portholes (B) to permit the sound to enter the rear of the capacitor head, thus producing a cardioid pattern. The solid-state amplifier (C) consists of an FET, capacitors, resistors, and an output transformer (D). A solid-electrolyte polarizer (E) supplies a permanent polarizing potential of 62 volts to the capacitor head. This polarizing system has a capacity of 15,000 microampere hours, and because of careful control of leakage, the expected life is around 20 years. At (F) is a Type TS-126 8.4-volt mercury battery with an expected life of 1000 hours. The output transformer supplies a balanced output, with a nominal impedance of 200 ohms.

One of the interesting features of this microphone is its transient re-

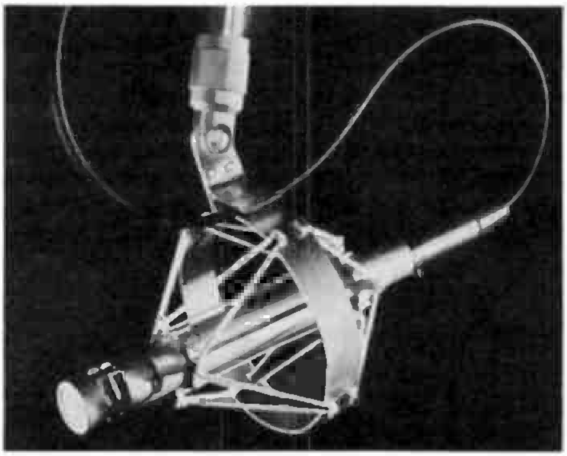


Fig. 4-46A. Synchron Corp. Model S-10 capacitor microphone.

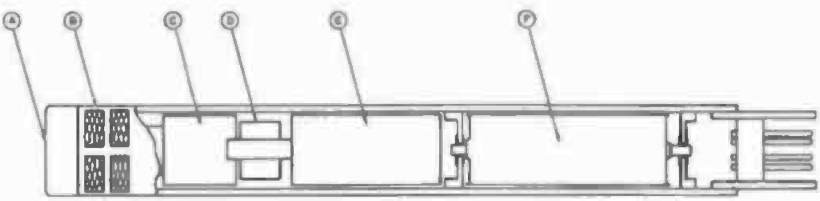


Fig. 4-46B. Cross-sectional view of Syncron Corp. Model S-10 capacitor microphone.

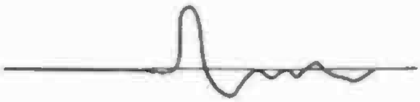


Fig. 4-46C. Rise time oscillograph for Syncron S-10 capacitor microphone.

sponse, as seen in Fig. 4-46C. The oscilloscope display shown was made with a shock-wave having a rise time of 1 microsecond, followed by a small amount of ringing and overshoot. The frequency response for three angles of incidence is shown in Fig. 4-46D, and the polar pattern appears in Fig. 4-46E.

The total harmonic distortion is less than 0.5 percent for an SPL up to 124 dB. The outpnt level is minus 53 dB re: 1 mv/10 dynes/cm². The nominal outpnt impedance is 200 ohms; however, it is unaffected by loading from 30 ohms to infinity.

4.47 What is a unidirectional microphone?—One having a greater sensitivity to sound pickup in one direction than another. The average unidirectional microphone has a back-to-front picknp of 20 to 26 dB; that is, it has 26 dB greater sensitivity to sound waves approaching from the front than from the rear. Directional characteristics may be obtained by using capacitor, dynamic, or ribbon-velocity design. Unidirectional microphones are used extensively for motion picture and television work.

4.48 Describe a dual-type microphone for stereophonic recording.—Microphones of the dual type are especially designed for the recording of stereophonic sound and may be of the capacitor or ribbon-velocity type. The first type to be discussed is the Neumann Model SM-2, manufactured in West Germany by George Neumann, previously known under the name, Telefnnken (Fig. 4-84A). An interior view of its construction is shown in

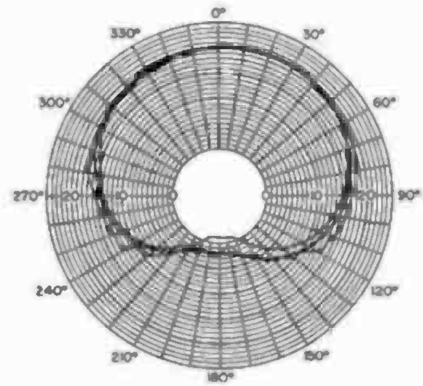


Fig. 4-46E. Polar pattern for microphone shown in Fig. 4-46A.

Fig. 4-48B. The microphone consists of two completely independent but similar capacitor microphones in a single case, one above the other (A) and (B). The upper unit may be rotated to achieve the MS (midscale) method of stereophonic recording technique, discovered by Lauridson, and sometimes termed the intensity system.

Each microphone system is of the pressure-gradient type head, employing two diaphragms (A and AA) and (B and BB), monnted on either side of two fixed-polarized electrodes. Each half of the system, (A) and (B), can be combined in a different manner, in order to obtain three different polar patterns: omnidirectional, cardioid, and bidirectional (figure-8). Below the capacitor head assembly are two preamplifiers, each employing a single AC-701K vacuum tube (C and CC). The second amplifier, not shown, is opposite (C). The polar pattern of each microphone can be independently and remotely adjusted for the three polar patterns and six intermediate patterns by changing the value of the polarizing voltage. This permits great flexibility and the variation of the polar patterns for optimum results.

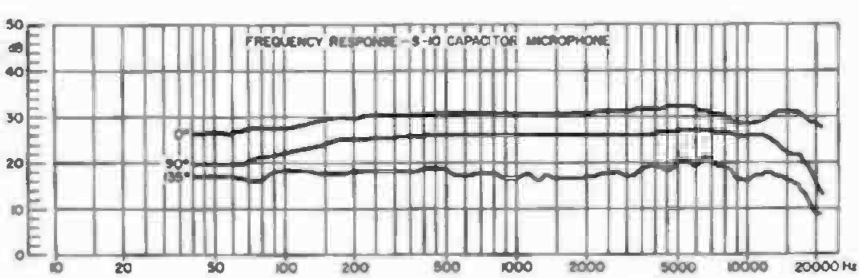


Fig. 4-46D. Frequency response for three angles of incidence.

Fig. 4-48A. Neuman Model SM-2 stereophonic microphone in its case. (Courtesy, Gotham Audio Corp.)

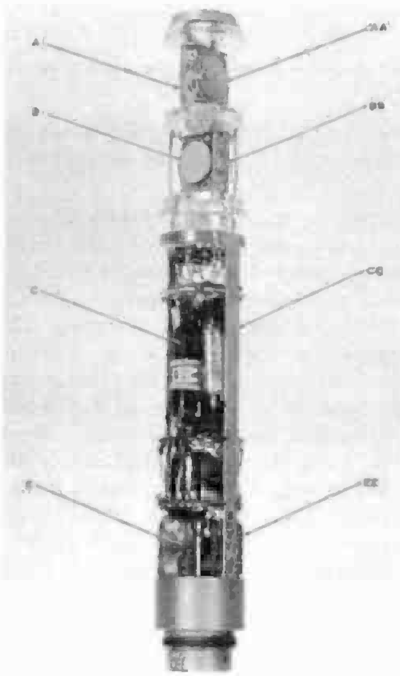


Fig. 4-48B. Interior view of Neumann Model SM-2 stereophonic microphone. The two capacitor heads may be seen at the top. (Courtesy, Gotham Audio Corp.)

Because of the small diameters of the microphone capsule (less than ¾ inch) the directional properties are almost independent of frequency; thus, the rear rejection ratio does not increase at the higher frequencies, relative to the middle or lower frequencies. This permits the microphone to be used close up in a more reverberant sound field. The high frequency rise which is generally required to compensate for the narrower pickup angle is unnecessary. The frequency response as measured in a linear sound field shows only a small rise at the high frequencies.

The frequency response for the three polar patterns is shown in Fig. 4-48C. Harmonic distortion is less than 0.4 percent over the entire frequency range, up to an SPL of 110 dB. The output impedance may be set for either 50 or 200 ohms, with an effective output level of minus 43 dBm. An external power supply model NSM supplies the operating voltages for the preamplifiers, and polarizing voltage for the capacitor heads. The output transformer is wound hum-bucking, to eliminate magnetic fields. The schematic diagram for a single side is shown in Fig. 4-48D.

A second dual-type microphone to be discussed is Model-200 stereophonic

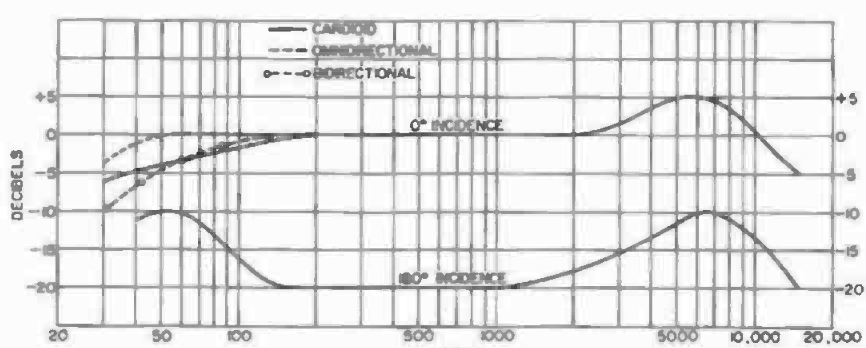


Fig. 4-48C. Frequency response for Neumann Model SM-2 stereophonic capacitor microphone, for three polar patterns.

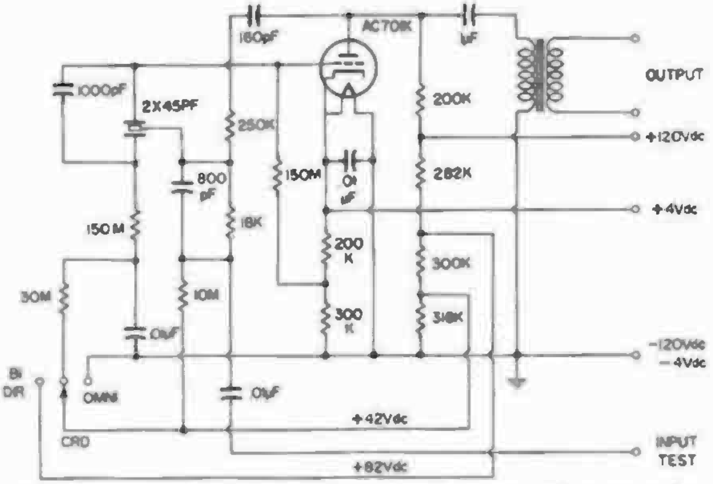


Fig. 4-48D. Schematic circuit for one side of Neuman SM-2 stereophonic capacitor microphone.

microphone, manufactured by Bang and Olufsen of Denmark (Fig. 4-48E). The device consists of two individual monophonic pressure-gradient bidirectional ribbon-velocity microphones, arranged one above the other, with provisions for rotating the upper unit 100 degrees.

The lower section is converted to stereophonic by plugging a second monophonic microphone in the top of the lower element. Such a design reduces phasing problems, since for practical purposes, both microphones occupy nearly the same space with respect to the source of sound pickup. The lower section includes a three-position switch (T) for talking, in which the low frequencies are rolled off at the rate of 3 dB per octave, whereby 125 Hz is minus 9 dB with respect to 1000 Hz. This is accomplished by muting the upper section and connecting a small inductance in parallel with a portion of the output transformer, making the microphone suitable for dialogue recording at distances of 8 to 20 inches. The polar pattern appears in Fig. 4-48F, and as can be observed, is quite constant from 200 to 10,000 Hz. The frequency response in music position is shown in Fig. 4-48G. The output level is minus 60 dB re: 1 volt/microbar. The output impedance is 150 and 250 ohms.

The moving element of each individual element consists of a dural ribbon, 0.0001 inch in thickness and weighing 1.3 milligrams. The ribbon, because of its weight, eliminates resonance in the normal operating frequency range. The magnetic structure consists of an anisotropic permanent Ticonal-E magnet in a magnetic circuit having negligible leakage. To obtain a smooth frequency response, the microphone must be hung on a soft suspension rather than on a stand; otherwise, a rise in frequency response below 50 Hz may be noted.

In stereophonic recording it is of extreme importance that the electrical

Fig. 4-48E. Bang and Olufsen, Struer Denmark Model 200 stereophonic ribbon microphone. Left, single monophonic unit; center, stereophonic; right, stereophonic with upper and lower units separated. (Courtesy, Dynaco Inc.)

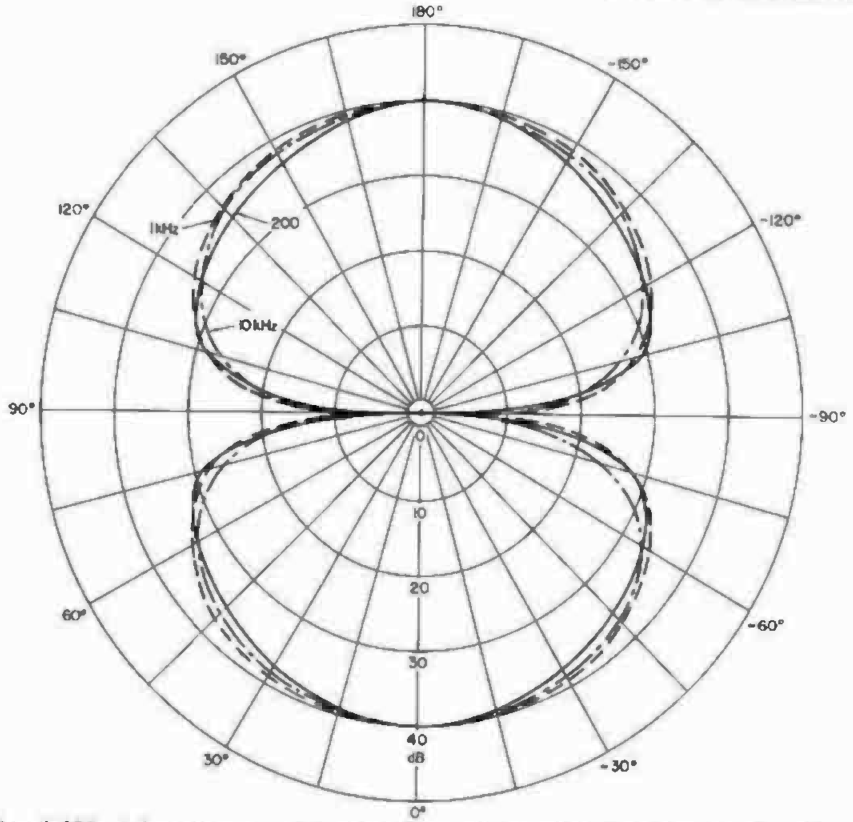


Fig. 4-48F. Polar pattern for Bang and Olufsen Model 200 stereophonic microphone. (Courtesy, Dynaco Inc.)

outputs from both microphone systems be in phase.

4.49 What is a bidirectional microphone?—A microphone which picks up at the front and back equally well, with little or no pickup at the sides. The field pattern is a figure-8. (See Fig. 4-7, part c.) Ribbon-velocity microphones have this characteristic.

4.50 What is a ribbon-velocity microphone?—A microphone in which a very light metallic ribbon is suspended in a strong magnetic field. Pressure waves cause the ribbon to vibrate in the magnetic field generating a voltage corresponding to the particle velocity of the pressure wave. Velocity microphones may be designed to have a wide frequency range, good sensitivity, low distortion, and low internal noise. Typical examples are the RCA 77DX and the MI-10001-C ribbon-velocity micro-

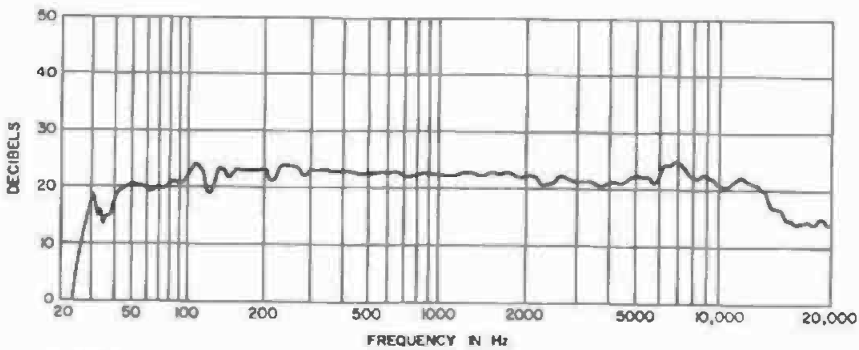


Fig. 4-48G. Frequency response for Bang and Olufsen Model 200 stereophonic microphone. (Courtesy, Dynaco Inc.)

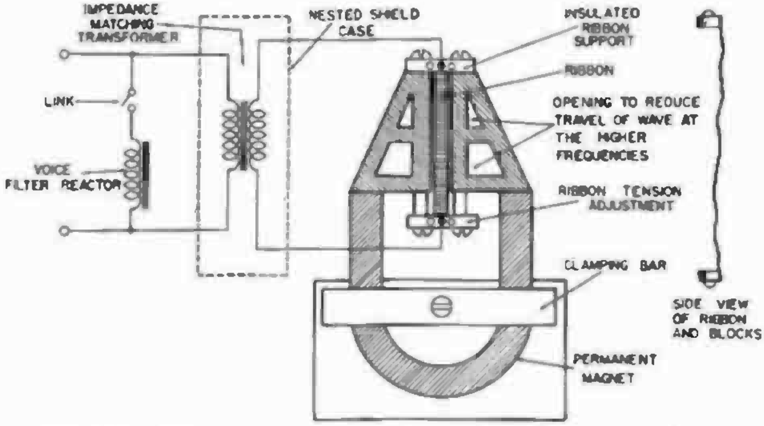


Fig. 4-51. Internal construction of a ribbon-velocity microphone.

phones discussed in Questions 4.59 and 4.77 respectively.

4.51 Describe the construction of a ribbon-velocity microphone.—Referring to Fig. 4-51, the magnetic structure consists of a horseshoe magnet with extended tapered pole pieces. A ribbon made of aluminum foil approximately 0.001 inch in thickness and corrugated throughout its full length for greater flexibility is suspended between the pole pieces on insulated supports.

4.52 Why are the pole pieces of a ribbon microphone tapered?—Because of the mechanical design, the magnetic structure acts similar to a baffle at the sides, obstructing the sound waves in their travel from the front to the rear surfaces of the ribbon. When a pressure wave passes, the ribbon is actuated by the variations in pressure and a voltage is generated by the movement of the ribbon in the magnetic field. This voltage is applied to the primary of an impedance-matching transformer and then to the input of a preamplifier. The ribbon in a velocity microphone may be looked upon as an inductive reactance. The higher the frequency, the higher the reactance of the ribbon. This causes the ribbon velocity to vary with frequency; however, the pressure on the ribbon surfaces will also increase with frequency and will continue to do so up to a point where the baffle dimensions equal one quarter the wavelength of the pressure wave frequency.

The greater the baffle dimensions (distance from front to rear surfaces of the ribbon) the greater will be the pressure; hence, a greater output results at the lower frequencies. At the higher frequencies the baffle affects the frequency response by reducing the output as the wavelengths of the pressure wave approach the baffle dimensions. When this situation occurs, the pressure is at a minimum for both surfaces of the ribbon. Holes are provided in the upper section of the magnetic pole pieces to reduce the length of travel for the higher frequencies. As a result, a greater output is obtained at the higher frequencies with a smoothing out of the frequency response.

When sounds originate at the side of the microphone in the same plane with the ribbon, a cancellation effect will take place, since the pressure on both the front and rear surfaces of the ribbon is equal. This cancellation action causes the field pattern to take the form of a figure-8, as shown in Fig. 4-7, part (c).

4.53 What is the impedance of the ribbon in a ribbon-velocity microphone?—Approximately 0.10 ohm, a value almost eqnivalent to its dc resistance. An impedance-matching transformer brings this low impedance up to the line impedance.

4.54 What is the resonant frequency of the ribbon in a ribbon-velocity microphone?—The ribbon is generally tuned so that the resonant frequency falls between 30 and 40 Hz. For microphones with a higher cutoff frequency, the resonant frequency is generally made higher.

4.55 What is a voice filter used with a ribbon-velocity microphone?—A small reactor connected in parallel with the output of the microphone to reduce the low-frequency response for speech.

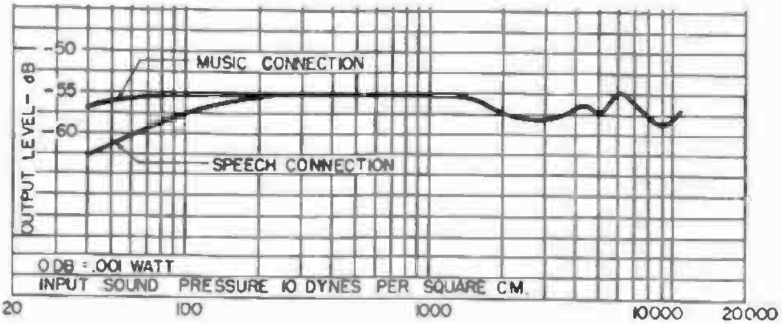


Fig. 4-56A. A typical frequency response for a bidirectional ribbon-velocity microphone.

At frequencies above 200 Hz, the reactor has little or no effect. Below this frequency, the response is slowly tapered off to the point where 40 Hz is attenuated 7.5 dB with respect to 1000 Hz. This is shown graphically in Fig. 4-56A. When recording music, the reactor is not nsed. Connections for the voice filter are shown in Fig. 4-51.

4.56 Show the frequency response and polar pattern for a bidirectional, ribbon-velocity microphone.—The frequency response for a typical bidirectional ribbon-velocity microphone is shown in Fig. 4-56A, with and without a voice filter. The polar pattern is shown in Fig. 4-56B. It will be noted that the response at 90 degrees is zero. This is caused by the pole-piece construction as described in Question 4.52.

4.57 Show the relationship of frequency to distance of a sound source for a bidirectional ribbon-velocity microphone.—As the distance between the sound source and the microphone is de-

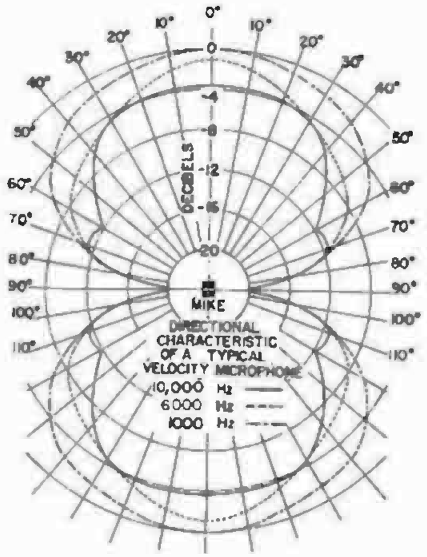


Fig. 4-56B. Polar pattern showing the bi-directional characteristics, of a typical ribbon-velocity microphone.

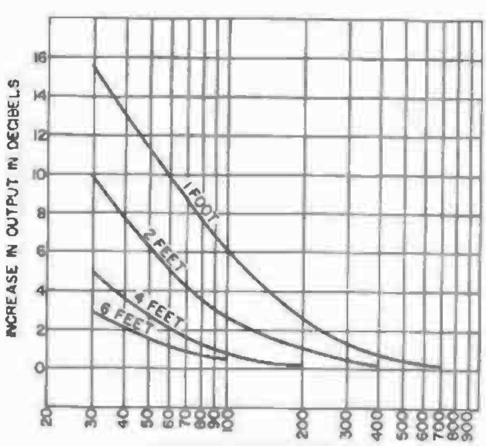


Fig. 4-57. Typical relationship of microphone distance to frequency response for ribbon-velocity bidirectional microphone.

creased, the low-frequency response will rise, as shown in Fig. 4-57. If the microphone is being used for dialogue pickup, the low-frequency response below 200 Hz must be rolled off, otherwise the reproduction at the low end will be excessive and tubby. The rolloff may be accomplished by the use of a voice filter, generally installed as an integral part of the microphone, or by the use of one of several dialogue equalizers discussed in Section 6.

Ribbon-velocity microphones should not be used closer than 4 feet from the sound source without a low-frequency rolloff. With the rolloff they may be used to within a few inches of the sound source. The increase in low-frequency response is characteristic of this type microphone. Ribbon-velocity microphones employing a directional characteristic give considerably less increase at the low frequencies for a given distance from the sound source, as illustrated in Fig. 4-57. This effect is referred to as the proximity effect.

4.58 What type microphones are generally used for motion-picture and television recording?—Capacitor or dynamic or ribbon-velocity microphones may be used if they provide a directional pattern and are suitable for boom operation. Typical types are the Altec-Lansing M-30 system and the 689BX, Electro-Voice 642 and 6681, and RCA MI-11010-A. Choice is not limited to the above microphones, but they do represent typical microphones developed for these express purposes.

4.59 Describe the construction of a unidirectional ribbon-velocity microphone.—A typical example of a unidirectional cardioid microphone designed for motion picture and television production recording, and broadcasting, is the RCA MI-10001C, shown in Fig. 4-59A, with its boom-suspension mechanism. The internal magnetic structure is similar to that of the bidirectional ribbon-velocity type, discussed in Question 4.51, with the exception of the added mechanism for obtaining the directional characteristics (Fig. 4-59B). The moving element is a single extremely light corrugated dural ribbon, suspended in an air gap, between the poles of a very highly charged permanent magnet (A). Thus, the ribbon can vibrate freely with the motion of the air particles of the sound wave. The voltage generated due to the movement of the ribbon cutting the magnetic field of force is the electrical equivalent of the velocity of the air particles. An acoustical labyriuth in the cylindrical section below the magnetic structure is filled with hair-

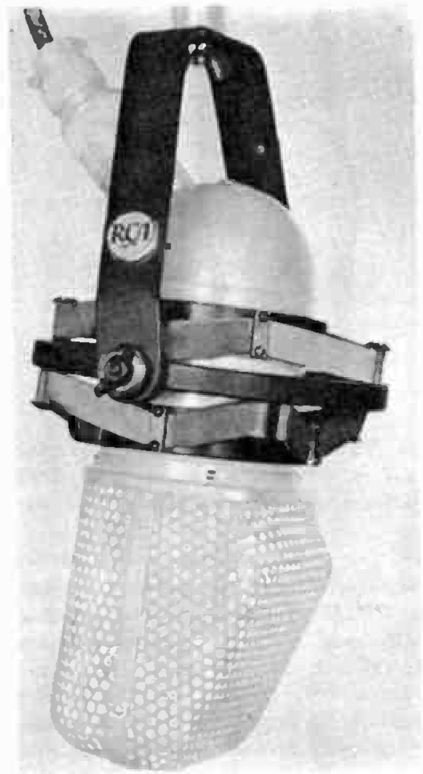


Fig. 4-59A. RCA MI-10001C unidirectional ribbon-velocity microphone for motion picture and television use.

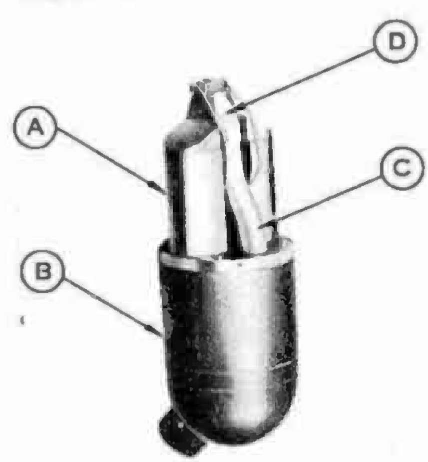


Fig. 4-59B. Interior view of RCA MI-10001C microphone.

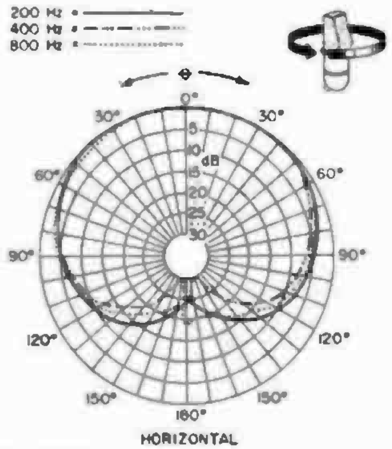


Fig. 4-59C. Horizontal polar pattern for RCA MI-10001C ribbon-velocity unidirectional microphone—200 to 800 Hz.

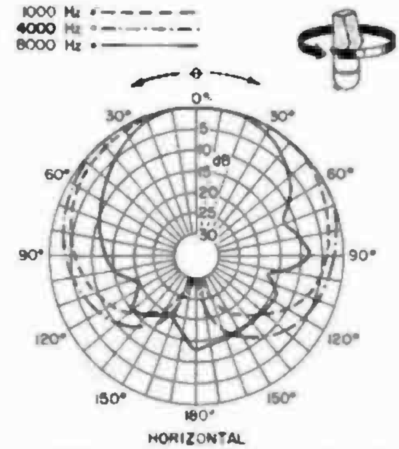


Fig. 4-59D. Horizontal polar pattern for RCA MI-10001C ribbon-velocity unidirectional microphone—1000 to 8000 Hz.

felt and terminates at the rear of the magnetic air gap by a tubular connector (C). The tubular connector is sealed to the rear of the air-gap, and contains a small silk covered opening, facing toward the ribbon (see Fig. 4-65). The area of the opening is of suitable dimensions to provide a cardioid or unidirectional polar pattern.

The ribbon and magnetic assembly are enclosed in a circular silk-lined grill, to provide protection to the mechanism and prevent metallic dust particles from entering the air gap. The hemispherical shell (B) also contains an impedance-matching transformer, with taps at 30, 150, and 250 ohms. The output signal is taken through a standard 3-pin male connector. Such microphones have a quite uniform frequency response for sound incidence at the zero axis (front of microphone), with output decreasing as the angle of the sound source with zero axis increases. The pickup angle is approximately plus or minus 50 degrees from the zero axis, with less than 1-dB difference in the output level, and approximately plus or minus 90 degrees before the output level decreases 6 dB over the frequency range normally employed. The broad angle of relativity equalizes the response, thereby reducing and simplifying "panning" the microphone, where two or more persons are concerned. The polar patterns in Fig. 4-59C and Fig. 4-59D are based on a plane-wave sound incidence. Referring to the plot in Fig.

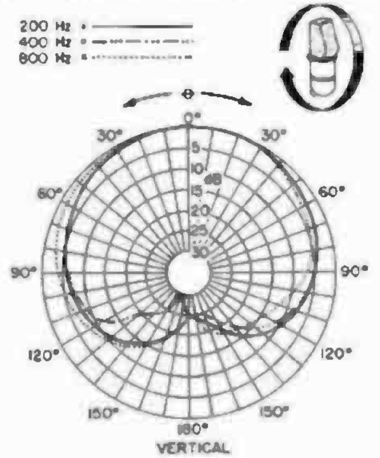


Fig. 4-59E. Vertical polar pattern for RCA MI-10001C ribbon-velocity unidirectional microphone—200 to 800 Hz.

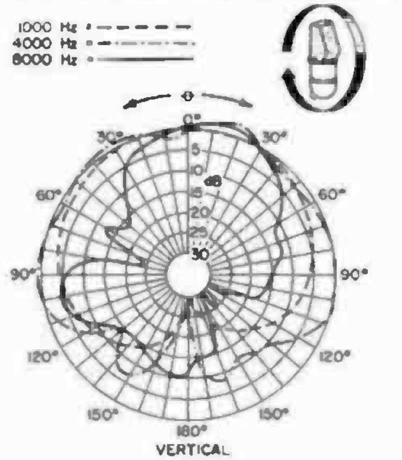


Fig. 4-59F. Vertical polar pattern for RCA MI-10001C ribbon-velocity unidirectional microphone—1000 to 8000 Hz.

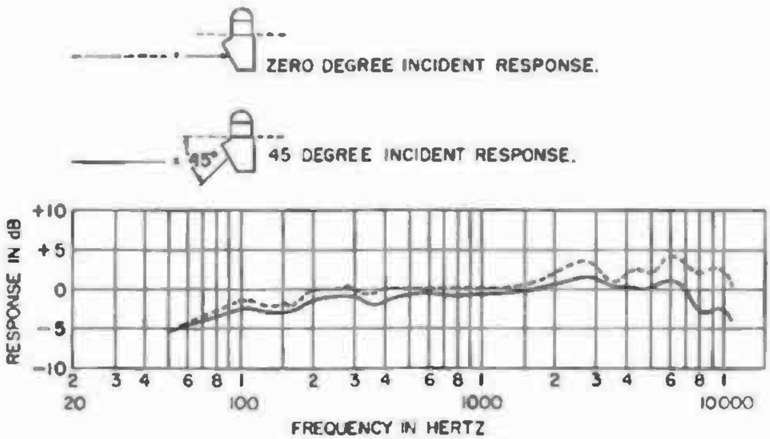


Fig. 4-59G. Frequency response of the RCA MI-10001C unidirectional microphone.

4-59C, cancellation for a plane wave at the rear is such that the cancellation is 20 dB or more at 180 degrees incident angle over a broad band of frequencies. As the distance between the microphone and sound source is decreased and the wavefront becomes increasingly curved, cancellation at the lower frequencies is reduced.

A loss of directional characteristic at the lower frequencies and short distances is such that the output level at a distance of 1 foot from the sound source is very nearly the same at the back and the front. As an example, the 80-Hz output is approximately plus 1.5 dB at 180-degree incident angle, re to zero dB at 1000-Hz for a zero-degree incident wavefront at one-foot distance. The 80-Hz zero axis output is accentuated approximately plus 3 dB, due to the low frequency tip-up characteristic of a velocity microphone.

It can be observed from Fig. 4-59G that the frequency response is relatively uniform over a range of 50 to 10,000 Hz. However, when the microphone is operated close to the sound source, the response will exhibit to a certain extent the low-frequency accentuation of a velocity microphone, although to a lesser degree than the conventional bidirectional ribbon microphone discussed in Questions 4.51 and 4.56. Under normal conditions the microphone under discussion should not be worked closer than 3 feet from the sound source, and preferably 4 feet, as shown in the curves of Fig. 4-59H.

As a rule, the RCA MI-10001C microphone is suspended at an angle of 45 degrees to the floor, with the microphone just outside the camera angle. The zero-degree axis (perpendicular to the front of the ribbon) should be directed toward the sound source, with the back facing unwanted sound sources, such as noises from camera, lights, traffic, and reflections from hard walls of a set. Referring to Fig. 4-59G, again the response changes as the angle of incident sound varies from zero to 45 degrees, and as the distance between the sound source and microphone is decreased (approaching the minimum of 3 feet), the lower frequencies are accentuated. Vertical polar patterns are shown in Fig. 4-59E and Fig. 4-59F for frequencies ranging between 200 and 800 Hz. Sensitivity to extraneous magnetic fields is quite low. For an exciting field of 0.001

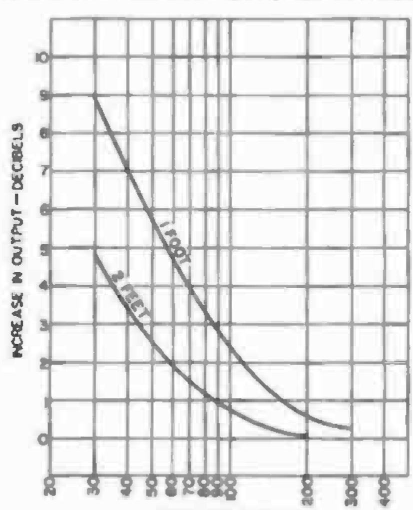


Fig. 4-59H. Relationship of microphone distance to frequency response for ribbon-velocity, unidirectional microphone.

gauss, the output level is minus 128 dBm. The signal output at 1000 Hz for a pressure of 10 dynes/cm² at the ribbon is minus 49 dB, using the 30-ohm output impedance.

Since random energy response of a unidirectional or cardioid microphone is one-third that of a nondirectional microphone, for the same permissible reverberation it may be used 1.7 times the distance of a nondirectional microphone, for a given set of conditions. The characteristics discussed were obtained operating the microphone into a preamplifier, using an unterminated input transformer. (See Questions 4.76 and 12.170.)

The hanger mechanism is designed for boom operation. The housing is clamped by two rings, supported by heavy rubber bands from a larger ring supported by the hanger arms, and hung from a boom turret-head by a single wing nut. Although the grill with the silk covering acts somewhat as a wind screen, it is not satisfactory for outside work if the wind is blowing more than 7 miles per hour. A large ball type wind screen that slips over the outside is generally employed.

As the distortion is quite low for this type microphone, it may be used for either dialogue or music recording. For dialogue, the low end is rolled off starting at 800 Hz to where it is down 8 to 12 dB at 100 Hz. This is the standard dialogue characteristic used by most motion picture studios. (See Questions 2.109, 6.80, and 18.81.)

4.60 Describe an early model unidirectional microphone.—One of the original unidirectional microphones developed for motion picture sound recording was the Western Electric RA-1142A. The magnetic structure of this microphone was quite similar to the 639A microphone used for broadcasting work. However, certain modifications were necessary to meet the requirements for sound recording and boom operation. The perforated metal case was added to improve the high frequency response and a boom hanger added with a shock mounting to protect the magnetic structure. Although this microphone has been obsolete for many years, its design is still of interest. The frequency response is given in Fig. 4-60B. (See Question 4.66.)

4.61 Describe the construction of a unidirectional microphone using two dynamic units.—The construction of the model D-202es unidirectional microphone, manufactured by AKG G.M.B.H. of Vienna, Austria, is shown in Fig. 4-61A. It employs two independent dynamic-unit (moving-coil) microphones, placed one above the other in a single

Fig. 4-60A. Interior view of an early (1935) Western Electric RA-1142A unidirectional microphone for motion-picture sound recording. Now obsolete, but it is still of historical interest.

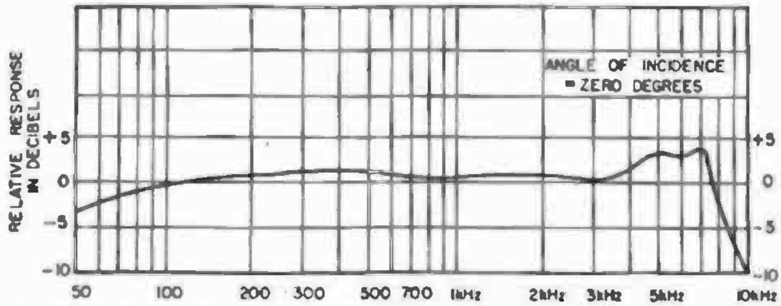


Fig. 4-60B. Frequency response of Western Electric RA-1142A unidirectional microphone.

housing, and connected electrically by a phase-correcting, dividing network. The frequency response of each dynamic unit is adjusted for optimum response at the low and high frequencies, crossing over at 500 Hz.

Referring to the cross-sectional view of the interior in Fig. 4-61B, at (A) is shown the high-frequency unit, employing a domed diaphragm (B) with a compensating coil (C) for eliminating the effect of extraneous magnetic fields. The high-frequency unit is mounted above the low-frequency unit by mounting plate (G). The domed diaphragm and its moving coil may be seen at (D), its magnets at (E), supported on shock mounts (F) and (H). A mass tube (I) projects into the low frequency microphone unit and connects with a smoothing chamber communicating with the outside sound field by means of a series of slotted openings (O), covered with a damping material to the lower end of housing (P). Because of the long sound-bypass distance (approximately 5½ inches) afforded by the mass tube and the subdividing of the frequency response between the high- and low-frequency units, it is possible to achieve a reduction on the distance effect or change in frequency response as a function of distance to the sound source.

The high-frequency diaphragm has a diameter of approximately ¾ inch and a sound bypass of ½ inch. A phase-correcting network, consisting of an RC network combined with an LC network, corrects the phase and frequency response at the higher frequencies. At (J) is a central screw for holding the assembly together, at (K) the crossover network, (L) an off-on switch, and at (M) a low-frequency attenuation switch and control that permits the low-frequency response at 50 Hz to be rolled off continuously to minus 20 dB, with reference to 1000 Hz. Item (N) is a standard 3-pin male connector. The upper end of the microphone housing (P) is covered with a sintered bronze cap (R) attached by a mounting ring (Q).

The frequency response is 30 to 15,000 Hz plus or minus 2 dB, with an output level of minus 53 dB re 0.2mv/microbar. The output impedance of 200 ohms is practically independent of frequency. The polar pattern is cardioid with a front-to-back ratio of 20 dB minimum over the entire frequency range. The off-axis response (90 degrees) is parallel with the on-axis (zero degrees) front curve. The microphone is designed for either boom or stand operation.

4.62 Describe the basic methods used for recording stereophonic sound.—The so-called *classical* method of recording stereophonic sound (sometimes called the AB or XY method) involves the use of two bidirectional microphones with identical characteristics.

Fig. 4-61A. Unidirectional microphone Model D-200es manufactured by AKG G.M.B.H, Vienna, Austria. (Courtesy, Sonocraft Corp.)

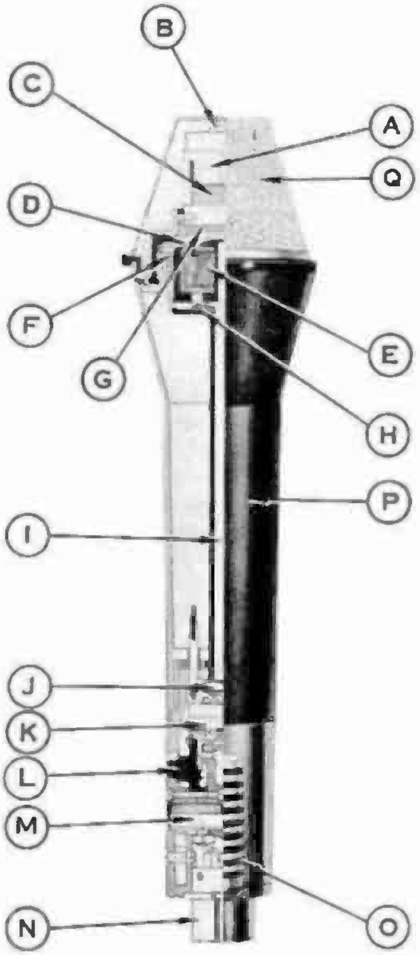


Fig. 4-61B. Cut-away view of unidirectional microphone Model D202-es manufactured by AKG, G.M.B.H. Vienna, Austria. (Courtesy, Sonocraft Corp.)

The microphones are placed side by side, spaced 12 to 48 inches apart, the exact distance depending on the desired working angle, relative to the source of sound. The microphones are connected as given in Fig. 4-62A. If additional vocal or solo microphones are required, they should consist of pairs placed fairly close together. An additional control (P5) is connected across the output from the vocal or solo microphones to provide a leakage path for reducing the effects of movement by the vocalist and the possibility of apparent movement when reproduced from one loudspeaker to the other.

For the listener to locate the sound source from the two loudspeakers, the speakers must be in the same position as were the microphones, to avoid differences in the distance from the sound source to the microphones. This may be accomplished in two different manners. The first method is by placing two microphones of similar electrical characteristics on a support as shown in Fig. 4-26B, suggested by Madsen, 12 to 20 inches apart, separated by a small baffle or sphere, thus creating a dummy head to enhance the left-to-right impressions. The use of the dummy head or baffle causes laterally displaced sound sources to be shadowed, resulting in a more uniform sound reproduction. The baffle has the effect of reducing the crowding of the sound source. Diffraction of the sound around the dummy head or baffle causes attenuation of the high frequencies on the far side. The resulting field pattern is shown in Fig. 4-62C.

The second method requires the mounting of the two microphones by a support, as shown in Fig. 4-62D. Here two capacitor-type microphones are placed one above the other as suggested by Lauridsen. The microphones must have similar electrical characteristics and their physical size must be small enough that they do not distort the sound field when they are placed in close proximity. They are rotated to a position where each microphone picks up the sound from half the studio.

The German (MS) method, also developed by Lauridsen, eliminates many of the drawbacks to the intensity system developed above. In the MS method (midside stereophony), one cardioid microphone supplies a complete pickup, similar to a single microphone pickup for monophonic recording. The second microphone having a bidirectional polar pattern is placed either above or below the cardioid and rotated to where its null point (x) meets the axis in the polar pattern of the cardioid microphone (Fig. 4-62E). Referring to Fig. 4-62F, if the output from the two microphones (A) and (B) in Fig. 4-62E are interconnected by means of differential transformers (S) and (M) to form an (A + B) and an (A — B) signal, two channels will result. In each channel one half the pickup area is preferentially received, relying on the fact the two principal axes of the pressure gradient microphone correspond to voltages of opposite polarity. Again

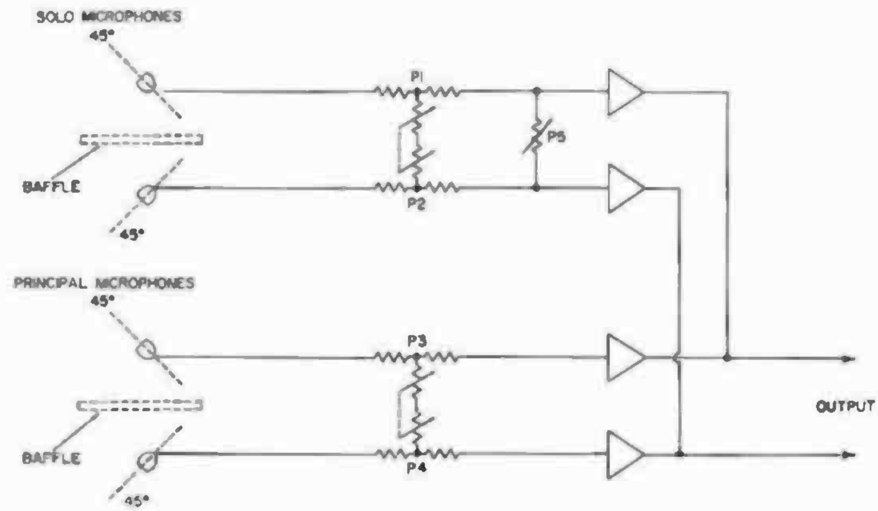


Fig. 4-62A. Classical two-channel method of recording stereophonic sound.

referring to Fig. 4-62E, and assuming the instantaneous value of the sound from the left microphone produces an instantaneous positive voltage (B) in the bidirectional microphone, the sound source on the central axis (M) will rise to the voltage (A) in only the cardioid

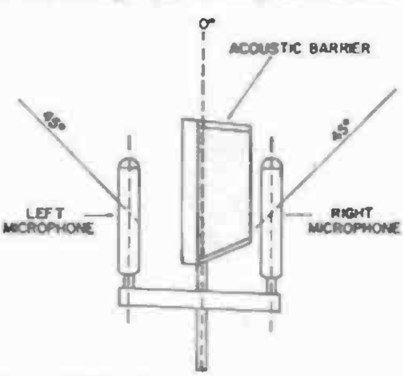


Fig. 4-62B. Stereophonic microphone placement using two ribbon-velocity bidirectional microphones set at an angle of 45-degrees each side of zero. (After Madsen.)

microphone, thus creating a central impression. Sound sources making angle (A1) with the central axis, give rise to a voltage (A + B) at the left loudspeaker, and (A − B) at the right loudspeaker. In a condition where (A = B) only the left loudspeaker is operative, and to the listener, the source is from the left. For sound sources at angle (A2) the sound appears to be coming from the right loudspeaker. For smaller angles in the recording studio, they correspond to the apparent direction. The size of the angle (A1 + A2) may be varied by changing the relative gain of the microphone channels. Experience indicates that sound sources lying outside angle (A1 + A2) will be more centralized, because the output from the bidirectional microphone predominates, causing the loudspeakers to be driven in opposite phase, resulting in a loss of direction to persons listening in a central position. This indicates that an individual microphone is necessary for a vocalist.

The advantages gained by the use of the MS method lie in the fact that one channel, namely the midchannel, produces a satisfactory single channel transmission; thus, the recording may be reproduced stereophonically or monophonically. To satisfactorily record stereophonically using the MS method requires dual microphones (see Question 4.48) or two individual microphones placed as near as possible to the same position (Fig. 4-62D). Dual-type microphones, when added for vocal or solo position, may be connected as shown in Fig. 4-62H.

The British E.M.I. system employs two bidirectional ribbon microphones (figure-8 pattern) placed at a 45-degree angle to the source of sound. The field patterns of the two bidirectional microphones create a pattern similar to a four-leaf clover (Fig. 4-62G). Since both microphones have equal sound pickup in the designated areas, A1 and A2, they may be connected in or out of

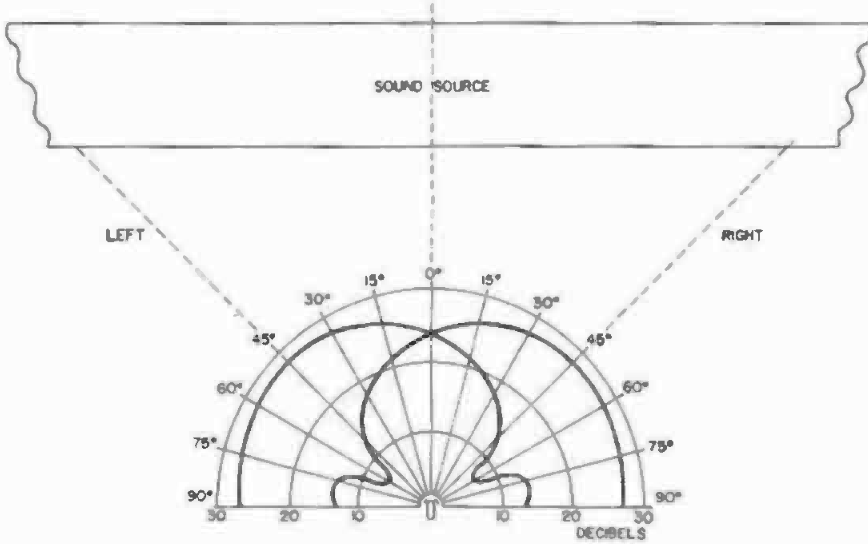


Fig. 4-62C. Polar pattern for two bidirectional ribbon-velocity microphones using an acoustic barrier between the microphones. Each microphone is turned 45 degrees from the zero point. (After Madsen.)

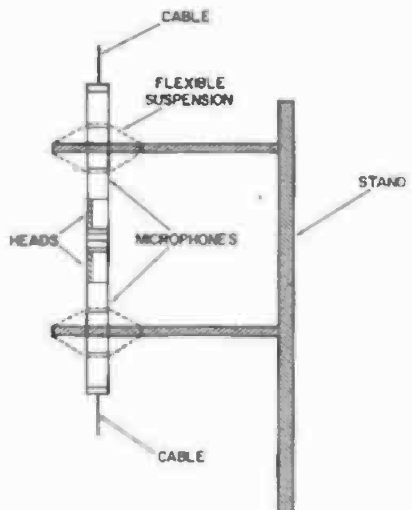


Fig. 4-62D. Two cardioid capacitor microphones mounted vertically head-to-head for stereophonic recording.

phase to produce a given signal at the cutting-head coils.

Connecting the output of the microphones in phase produces a lateral signal component equivalent to the (M) component in the German MS system (Fig. 4-62E). If the outputs are connected out of phase, the resulting field pattern is that of the (S) or vertical component of the German system. For

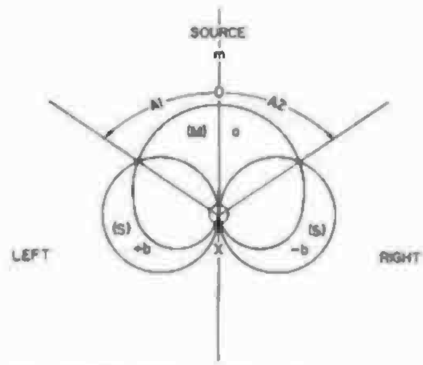


Fig. 4-62E. Polar pattern for a microphone combination using cardioid and bidirectional microphones.

signals of random differences, the resulting signal at the cutting head is a complex one, a combination of both lateral and vertical motion.

In the United States and Europe, many variations of the methods of sound pickup described above as used in combination with the 45/45 degree system of disc recording developed by Davis and Frayne of the Westrex Corp., (USA). Reproducing characteristic in the United States is that standardized by the Record Industry Association of America (RIAA), Fig. 13-95. As a rule, each recording activity develops a system of microphone placement and method of operation which is peculiar to itself.

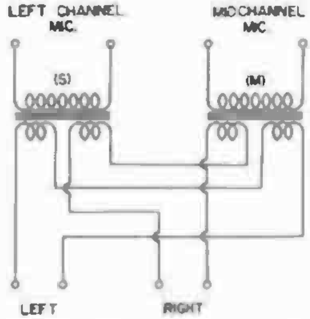


Fig. 4-62F. Differential transformers connected at the output of two microphones to obtain sum and difference voltages for MS stereophonic recording.

4.63 How does the transient response of a dynamic microphone compare to a capacitor microphone?—The transient responses of a dynamic and capacitor microphone are compared graphically in Fig. 4-63. It can be observed that the capacitor microphone has a much faster rise time than the dynamic microphone. The dynamic microphone has a longer rise time because of the diaphragm and the inductance of the moving coil. Although the capacitor microphone *appears* to have better transient response, in both instances *the rise time amounts to microseconds;* the capacitor rising from 10 percent of its rise time to 90 percent in approximately 15 microseconds, while the rise time for the dynamic microphone is on the order of 40 microseconds. The measurement of rise time is discussed in Question 22.74.

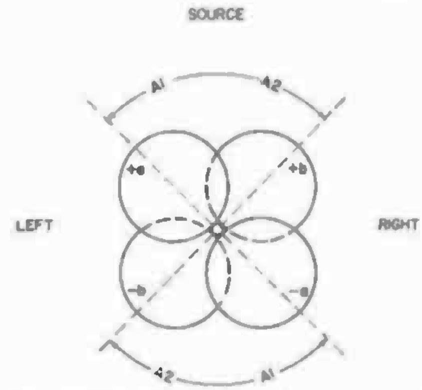


Fig. 4-62G. Microphone placement and field pattern for British E.M.I. method of recording stereophonic sound.

4.64 Show the methods used for grounding microphones and microphone cables.—The grounding of microphones and their interconnecting cables is of extreme importance, since any hum frequencies or noise picked up by the cables will be amplified along with the audio signal. Professional systems gen-

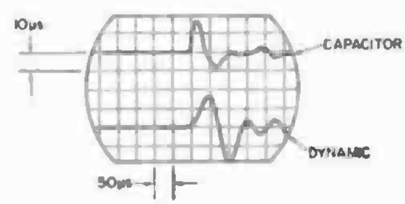


Fig. 4-63. Transient time of capacitor microphone compared to that of a dynamic microphone.

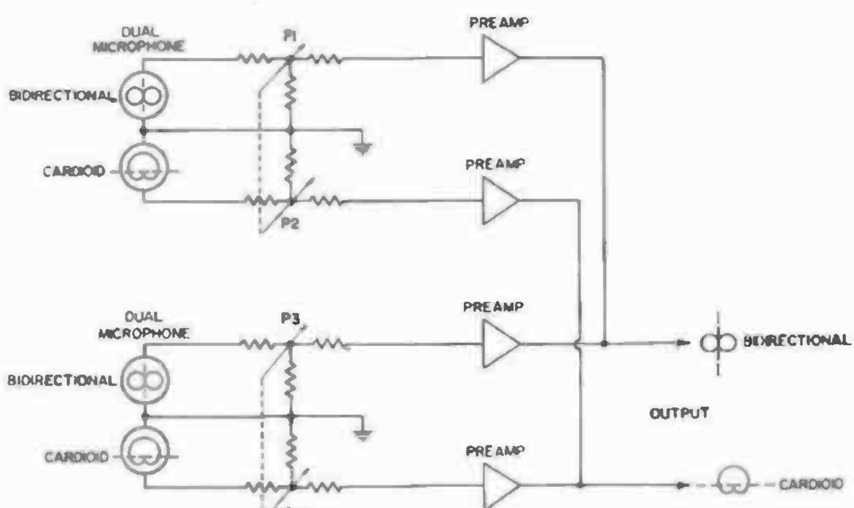


Fig. 4-62H. Mixing circuit for two dual-type microphones when added for vocal or solo positions.

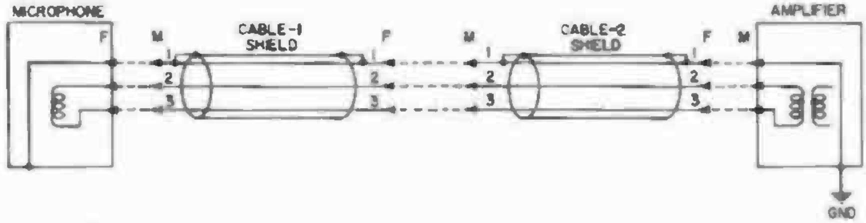


Fig. 4-64A. Grounding method for 3-conductor microphone cables. The physical ground connection is made at the amplifier or mixer console.

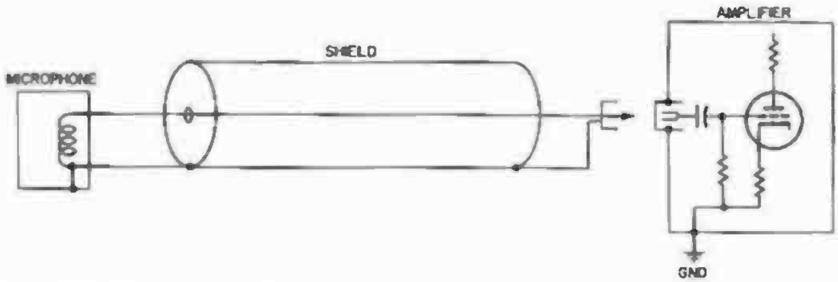


Fig. 4-64B. Grounding method used for 3-connector microphone cables using pin-plugs. The physical ground is connected at the amplifier end only.

erally use the method shown in Fig. 4-64A. Here the signal is passed through a three-conductor cable to the input transformer of a preamplifier. It should be observed, that the cable shield is connected to conductor and pin number 1, and the audio signal is carried by conductors and pins 2 and 3. The actual physical ground is connected at the preamplifier chassis, and carried to the microphone case over conductor number 1. In no instance is a second ground ever connected to the far end of the cable, as to do this will cause the flow of ground currents between two points of grounding. The grounding of electronic circuits and their associated equipment is discussed in Section 24.

In the making up of microphone cables, precautions must be taken to establish a color code and follow it through all cables. The pin count shown is that used by RCA. The plus or minus side of the audio signal (hot side) is connected to pin number 2.

In systems designed for semiprofessional and home use, the method in Fig. 4-64B is used. It will be noted that one side of the audio signal is carried over the cable shield to a piu-type connector. The bodies of both the male and female connector are grounded, the female to the amplifier case and the male to the cable shield. The microphone end is connected in a similar manner; here again the physical ground is connected only at the preamplifier chassis. Pin counts are also discussed in Questions 24.42 and 24.100.

4.65 ***Describe how a cardioid polar pattern is obtained with a ribbon-velocity microphone.***—The basic principles of securing a directional patteru with a ribbon-velocity microphone are shown in Fig. 4-65. The dural ribbon is suspended in a strong magnetic field, as described for the ribbon-velocity microphone in Questions 4.51 and 4.52. The ribbon is anchored at its mechanical center and grounded to the case. This effectively divides the ribbon into two separate parts. Directly behind the upper ribbon is a metal tube which connects to an acoustical labyrinth located in the base of the microphone case. The labyrinth is filled with hair felt and dissipates sound waves traveling down the pipe in the form of heat. With this type construction, the upper portion of the microphone becomes a pressure-operated device, while the lower section operates as a normal ribbon-velocity microphone. The overall device now becomes a combination of a pressure and a velocity microphone. The extreme ends of the ribbon are connected to a transformer.

Pressure waves striking the front of the ribbons cause voltages, which are in series and in phase, to be generated

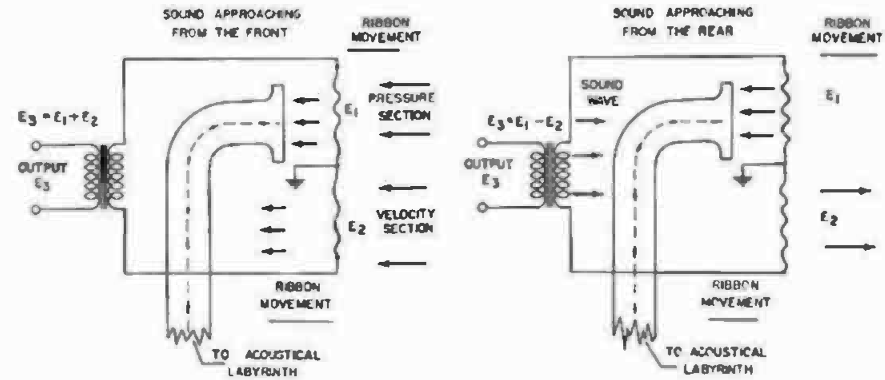


Fig. 4-65. Basic principles of operation of a unidirectional ribbon-velocity microphone.

simultaneously in both the upper and lower ribbon sections. Hence, these voltages are additive. However, pressure waves, either reflected or direct, which strike the ribbons from the rear will have little effect, since they generate voltages which are in series, but out of phase. Furthermore, it is difficult for waves approaching from the rear to strike the upper ribbon, due by the mechanical interference offered by the pipe connecting to the acoustical labyrinth below. Thus, the effect of waves approaching from the rear is minimized, since the waves are able to actuate only the lower half of the ribbon and have little or no effect on the voltages generated by waves striking the front of the ribbous. (See Question 4.59.)

4.66 Describe the construction of a directional microphone employing both a ribbon-velocity and pressure unit.—Microphones of this design (Fig. 4-66A) make use of both a pressure and ribbon-velocity unit to obtain a cardioid omnidirectional or bidirectional polar patterns. The output voltages from the two microphones are used independently, or combined in various proportions to obtain a variety of polar patterns. Six different patterns are available by means of a switch at the rear of the microphone housing. The pressure uuit is a moving-coil dynamic unit microphone, similar to the Altec-Lansing 633a/c, described in Question 4.30, and when used alone, has an omnidirectional polar pattern. The ribbon-velocity section has a bidirectional pattern. Combining the output voltages of these two units results in the phasing of the voltages in such a manner to produce a cardioid polar pattern. The output impedance is approximately 35 ohms and is operated into a 30- to 50-ohm input. The output level is minus 52 dB, re: 10 dynes/cm².

Fig. 4-66A. Altec-Lansing Model 639B microphone. Six different field patterns are made available by means of a switch at the rear of the housing.

4.67 Show the frequency response and angle of incidence for the microphone discussed in Question 4.66.—Both the frequency response and field patterns for this microphone are shown in Fig. 4-67. The single letter at the top of each characteristic may be interpreted as follows: R-ribbon, D-dynamic, C-cardioid. Numbers 1, 2, and 3 are variations of C.

4.68 What is a polydirectional microphone?—A microphone in which the polar pattern may be changed from omnidirectional to a bidirectional, a cardioid, or a combination of the three.

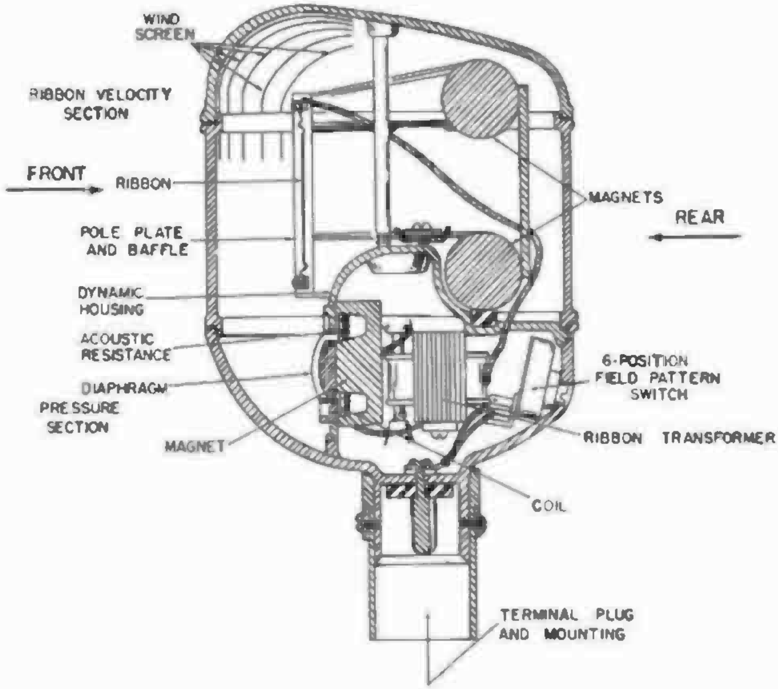


Fig. 4-66B. A cross-sectional view of Altec-Lansing Model 639B microphone.

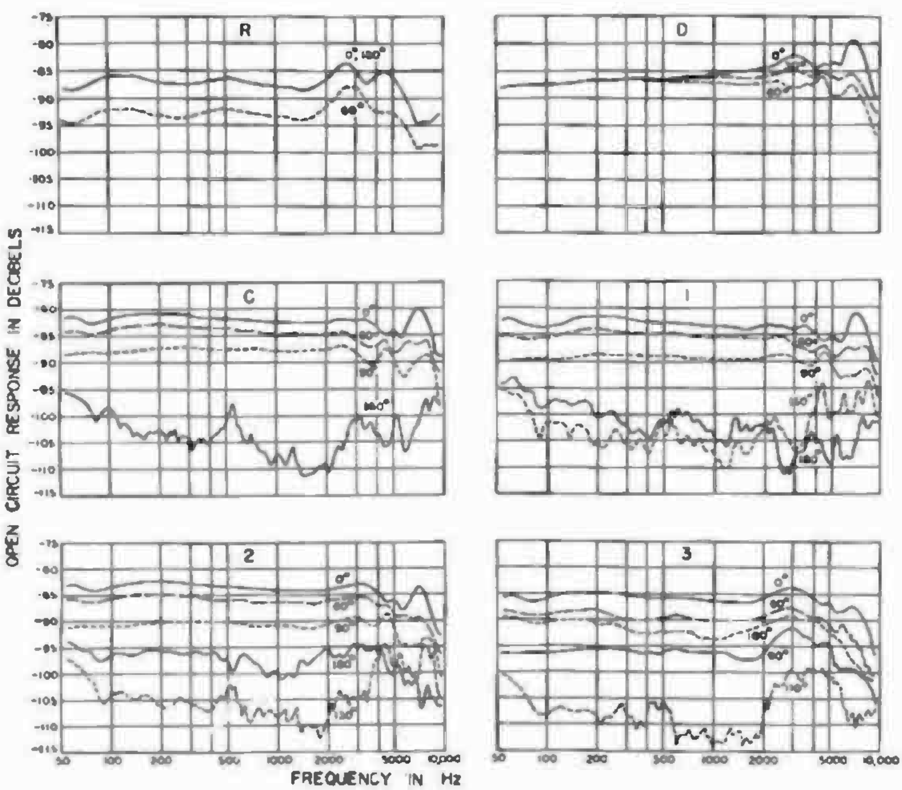


Fig. 4-67. Frequency response for Altec-Lansing 639B microphone, for various angles of Incidence.

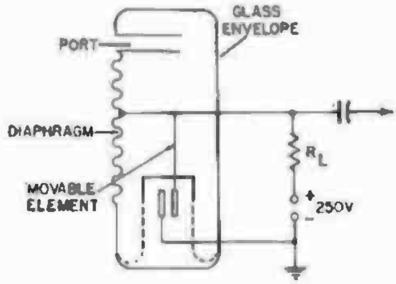


Fig. 4-69. The RCA type 5734 mechano-electronic transducer. This device is a vacuum-tube type microphone.

Such a microphone is discussed in Question 4.77.

4.69 What is an electronic microphone?—A microphone constructed in much the same manner as a vacuum tube. One of the internal elements is connected to an external diaphragm. Pressure waves striking the diaphragm move the internal element causing it to be displaced. This action causes a change in plate current proportional to the diaphragm displacement. The output voltage is amplified in the usual manner. Since the device is pressure operated, its field pattern is circular. A cross-sectional drawing of its construction is shown in Fig. 4-69. It is sometimes referred to as a mechano-electronic microphone. It was developed by RCA but is not used commercially.

4.70 What is a frequency-modulated microphone?—A capacitor microphone which is connected to a radio-frequency oscillator. Pressure waves striking the diaphragm cause variations in the capacity of the microphone head, which frequency-modulates the oscillator. The output of the modulated oscillator is passed to a discriminator aud amplified in the usual manner.

4.71 Describe a capacitor microphone using a radio frequency oscillator.—Capacitor microphones using a radio-frequency oscillator are not entirely new to the recording profession, but since the advent of the transistor considerable improvement has been achieved in design and characteristics. An interesting microphone of this design is the Schoeps Model CMT26U (Fig. 4-71A) manufactured in West Germany by Schall-Technik, and named after Dr. Carl Schoeps, the designer.

The basic circuitry is shown in Fig. 4-71B. By means of a single transistor, two oscillatory circuits are excited and tuned to the exact same frequency of 3.7 MHz. The output voltage from these circuits is rectified by a phase-bridge detector circuit, which operates over a large linear modulation range with very small radio-frequency voltages from the oscillator. The amplitude and polarity of the ontput voltage from the bridge depends on the phase angle between the two high-frequency voltages. The microphone capsule (head) acts as a variable capacitance in one of the oscillator circuits. When a sound wave impinges on the surface of the diaphragm of the microphone head, the vibrations of the diaphragm are detected by the phase curve of the oscillator circuit, and an audio frequency voltage is developed at the output of the bridge circuit. The microphone-head diaphragm is of metal to guarantee a large constant capacitance. An automatic frequency control (afc) with a large range of operation is provided by means of capacitance diodes to preclude any influence caused by aging or temperature changes on the frequency-determining elements, which might throw the circuitry out of balance.

The internal output resistance is 200 ohms, fed directly from the bridge circuit through two capacitors and delivering an output level of minus 51 to 49

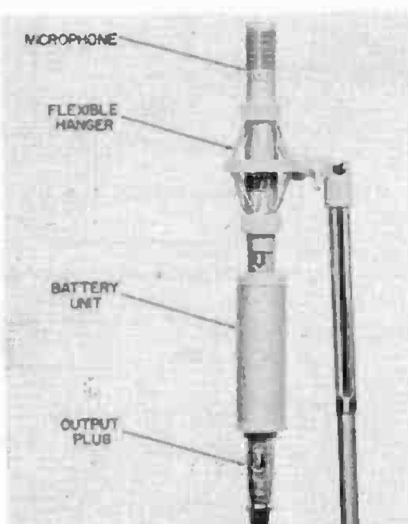


Fig. 4-71A. Schoeps (West Germany) Model CMT26U radio-frequency capacitor microphone, with BZ02 battery supply unit. (Courtesy, International Electroacoustics Inc.)

dB (depending on the polar pattern used), into a 200-ohm load for an SPL of 10 dynes/cm². The signal-to-noise ratio and the distortion are independent of the load because of the bridge circuit; therefore, the microphone may be operated into load impedances ranging from 30 to 200 ohms. The manufacturer suggests that no output transformer be used; however, this will be determined by the mixer input circuits employed. The audio signal and a dc operating potential of 8.5 volts are carried over a two conductor cable; the battery volt-

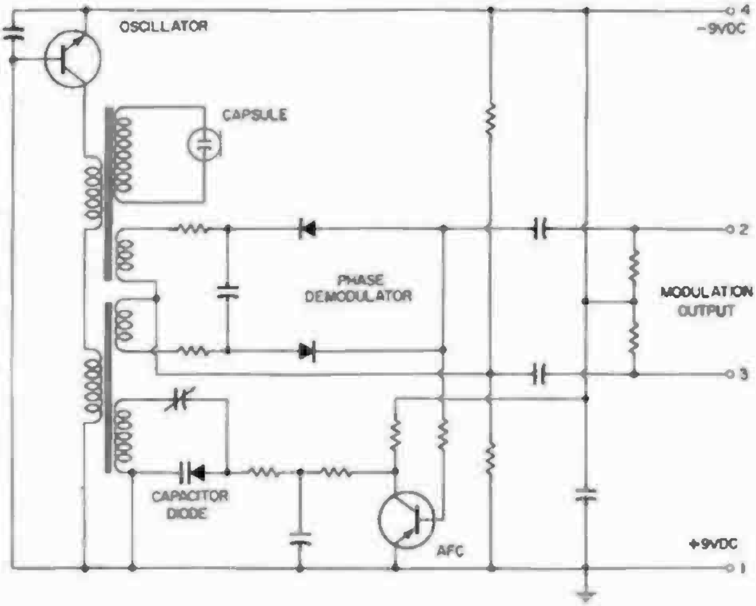


Fig. 4-71B. Basic circuit for the Schoeps radio-frequency capacitor microphones, series CMT.

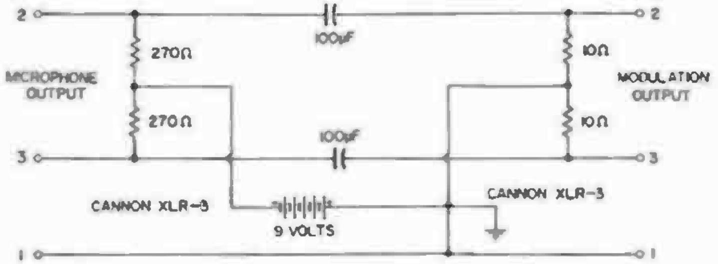


Fig. 4-71C. Schematic circuit for Schoeps battery supply unit BZ02.

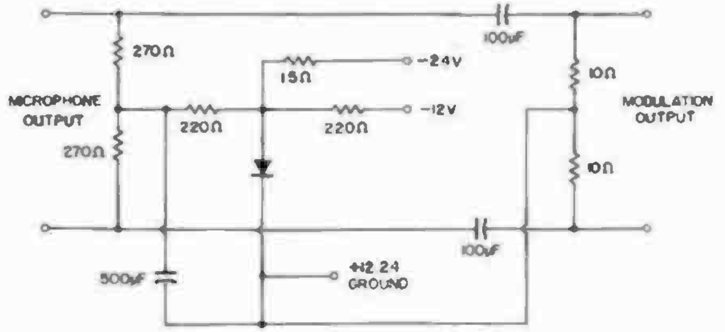


Fig. 4-71D. Network for operating a series CMT microphone from an available power supply, either 12 or 24 volts dc.

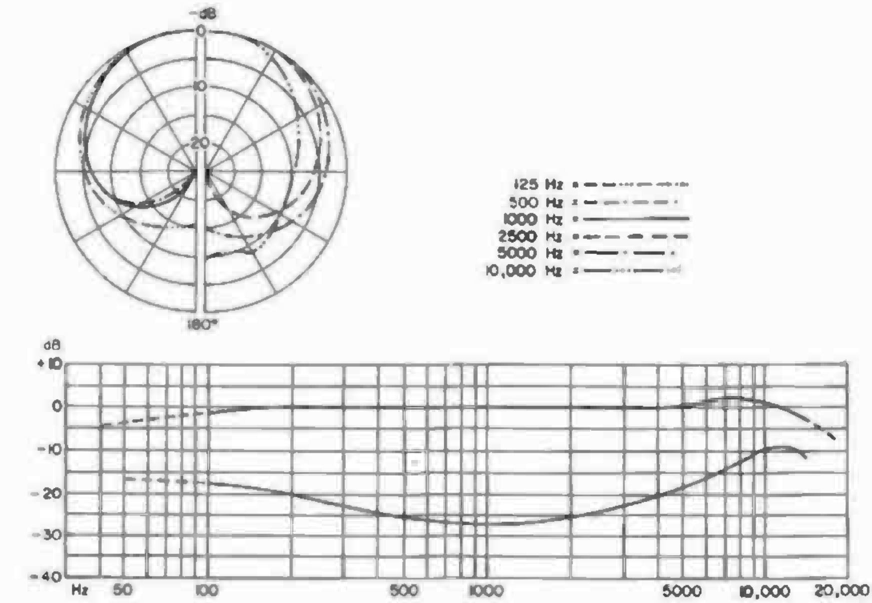


Fig. 4-71E. Cardioid frequency response and polar pattern for Schoeps Model CMT26U capacitor microphone.

age is fed across the center point of the output circuit and the cable shield. The center or neutral point is created by the two resistors across the output of the microphone bridge circuit, and the two resistors on the input side of the battery unit (Fig. 4-71C).

The advantage of supplying the operating voltage in this manner is that since the modulation voltage is not superimposed on the supply voltage, the crosstalk damping is around 100 dB, even when operating several microphones from the same supply source. The internal noise voltage when terminated in 1000 ohms ranges from 0.30 to 0.44 microvolt, depending on the polar pattern used. Hum pickup is negligible

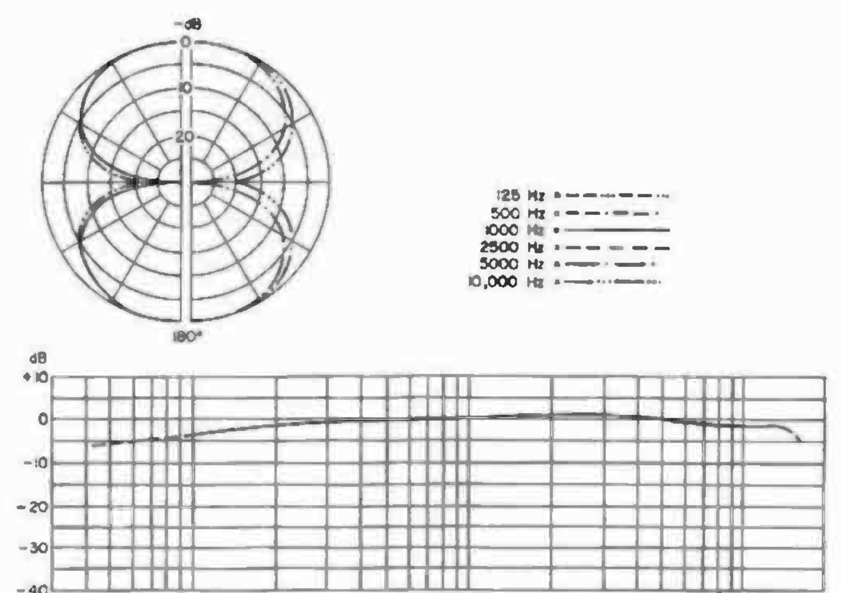


Fig. 4-71F. Bidirectional frequency response and polar pattern for Schoeps Model CMT26U capacitor microphone.

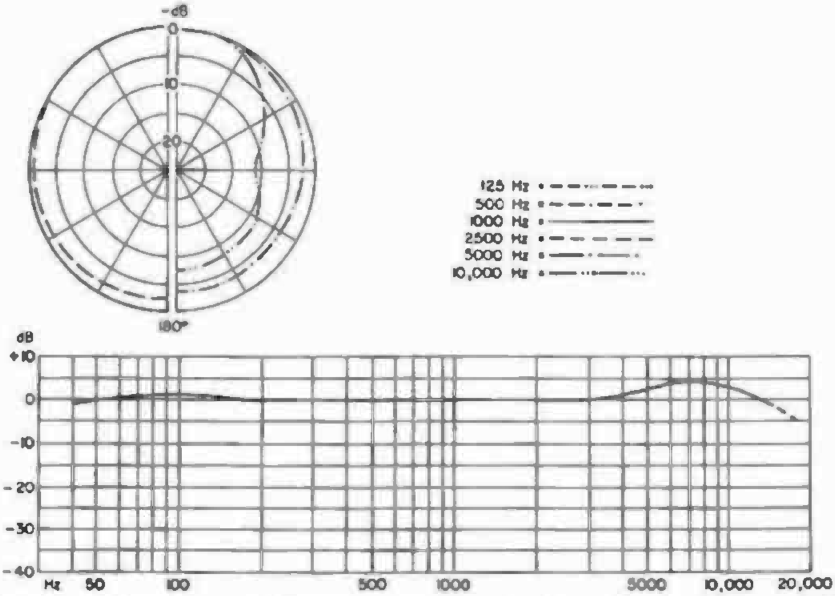


Fig. 4-71G. Omnidirectional frequency response and polar pattern for Schoeps Model CMT26U capacitor microphone.

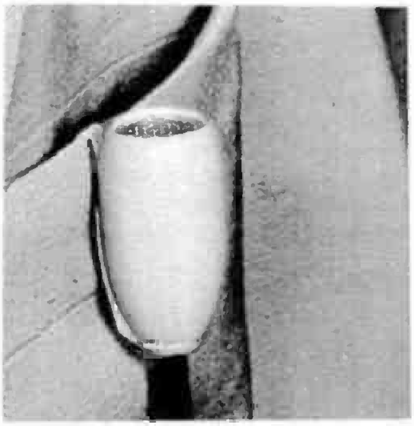


Fig. 4-71H. Microphone head for rf type microphone used as a lapel microphone.

because interference superimposed on the supply voltage does not directly reach the modulation. The phase of the output voltage may be reversed without modification to the power-supply source. A resistive network for supplying the operating voltage from an existing power supply is shown in Fig. 4-71D. Frequency response and polar patterns for three methods of operation are shown in Figs. 4-71E, F, and G.

A second microphone of the capacitor type, developed by Stephens, utilizes the advantages of the capacitor microphone in a circuit that does not require the use of a preamplifier at the capacitor head, as it differs in the manner in which the minute changes of capac-

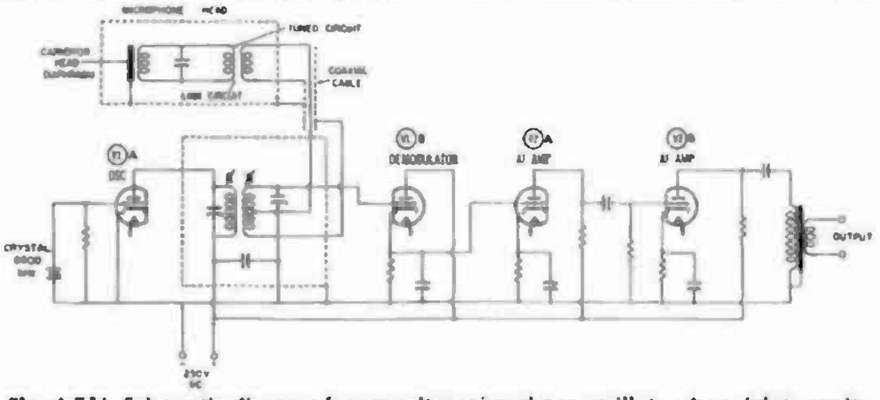


Fig. 4-71I. Schematic diagram for capacitor microphone oscillator-demodulator unit. (After Stephens.)

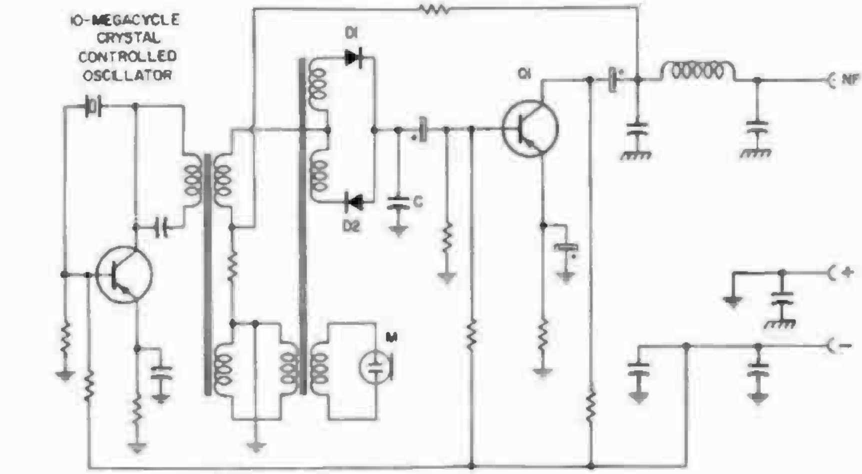


Fig. 4-71K. Basic circuitry for Sennheiser capacitor microphone Model 404 employing an rf oscillator.

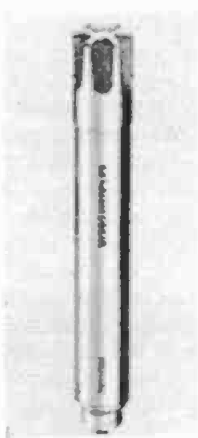


Fig. 4-71J. External appearance of Sennheiser Models MKH404 and MKH405 capacitor microphones having cardioid directional characteristics.

itance are used. No polarizing voltage is required, as in the conventional capacitance microphone. The head assembly, Fig. 4-71H, contains a *resonant* circuit link-coupled by a coaxial cable to a crystal-controlled oscillator. Tuning of the circuit to a frequency that is approximately that of the crystal oscillator is provided by the capacitance of the head (Fig. 4-71I). Sound waves at the diaphragm cause small changes in the head capacitance and shift the frequency of the oscillator either above or below the nominal operating frequency. A demodulator (detector) converts the rf changes into audio frequencies, which are amplified in the usual manner. Typical operating frequencies for the crystal are: 8.6 kHz and 8.725 kHz. The cable length between the capacitor head

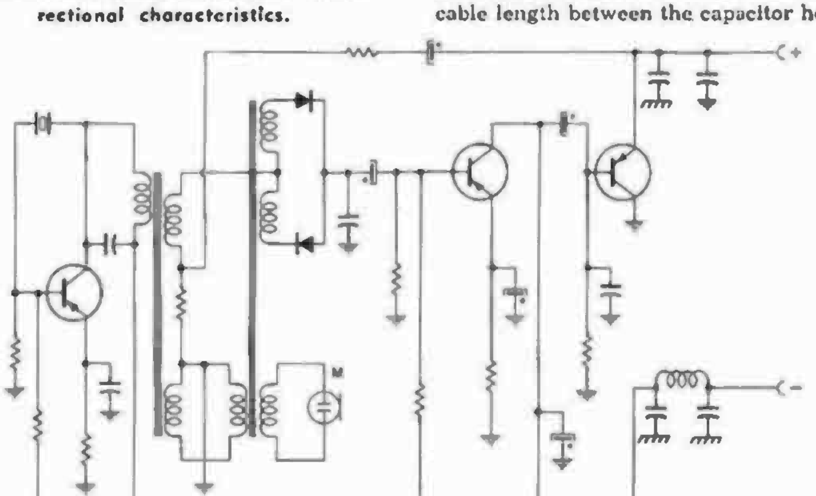


Fig. 4-71L. Basic circuitry for Sennheiser capacitor microphones Model 405 and 805.

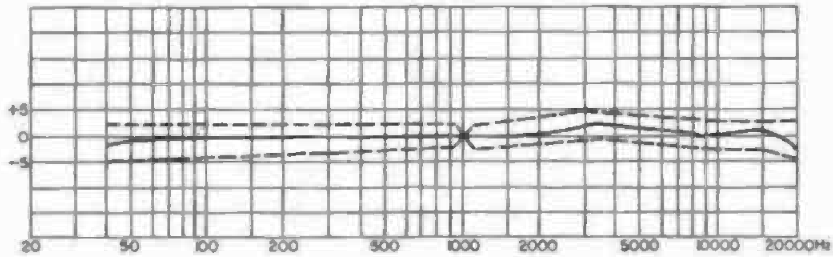


Fig. 4-71M. Frequency response and manufacturing tolerance for Sennheiser Model MKH405 capacitor microphone.

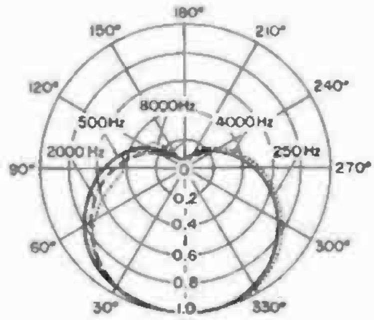


Fig. 4-17N. Field pattern for Sennheiser Model MKH405 capacitor microphone.

and the oscillator is quite critical, and is cut in lengths of 37.5 inches. The oscillator-modulator may be placed up to 400 feet from the capacitor head.

A capacitor microphone of somewhat different design, manufactured by Sennheiser of West Germany and also employing a crystal-controlled oscillator, is shown in Fig. 4-71J. In the conventional capacitor microphone (without oscillator) the input impedance of the preamplifier is on the order of 100 megohms; therefore, it is necessary to place the capacitor head and preamplifier in close proximity. In the Sennheiser microphone, the capacitive element (head) used with the rf circuitry is of lower impedance, since the effect of a small change in capacitance at rf frequencies is considerably greater than at audio frequencies. Instead of the capacitor head being subjected to a high dc polarizing potential, the head in this microphone is subjected to an rf voltage of only a few volts.

Since the preamplifier and crystal-controlled oscillator are of transistor design, they are assembled in an integral unit. An external power supply of 12 Vdc is required. The circuitry for a Model MKH404 microphone is shown in Fig. 4-71K, with the circuitry for a Model MKH405 shown in Fig. 4-71L.

Referring to Fig. 4-71K, the output voltage of the 10-MHz oscillator is periodically switched by diodes D1 and D2 to capacitor C. The switching phase is shifted 90 degrees from that of the oscillator by means of loose coupling, and aligning the resonance of the microphone circuit M under a no-sound condition. As a result, the voltage across capacitor C is zero. When a sound impinges on the diaphragm, the switching phase changes proportional to the sound pressure, and a corresponding audio voltage appears across capacitor C. The output of the switching diodes is di-

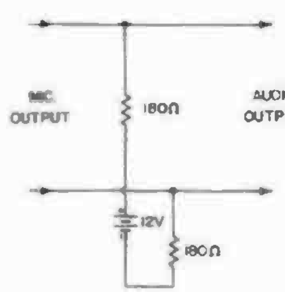


Fig. 4-71O. Duplexing the operating voltage for a microphone over an audio line.

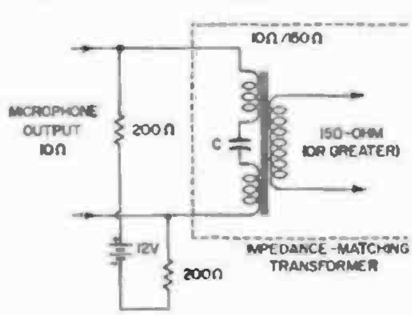


Fig. 4-71P. Impedance-matching transformer connected in the output of a microphone, with the operating voltage duplexed over the audio line.

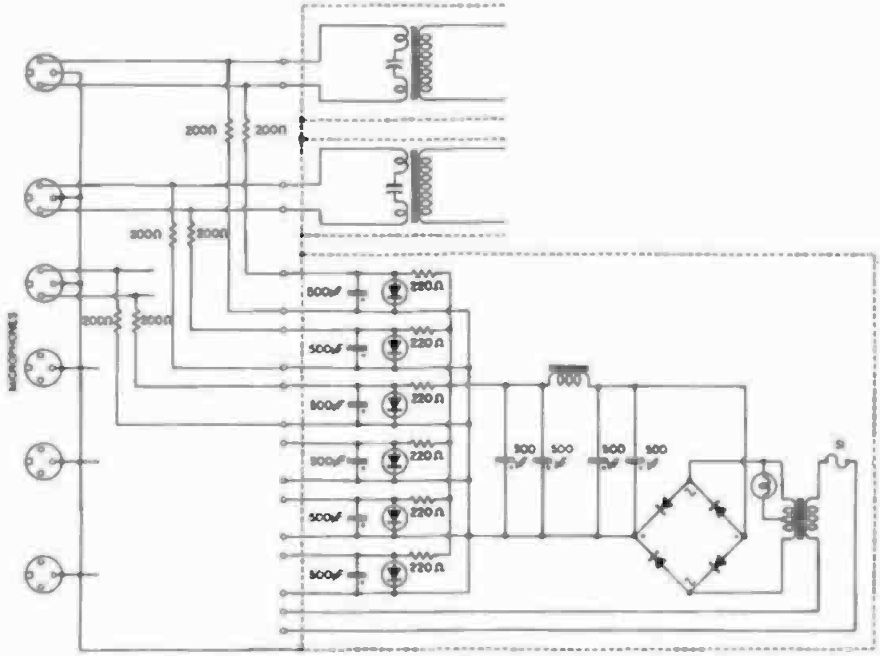


Fig. 4-71Q. Alternating-current power supply for six capacitor microphones. The dc operating voltages are duplexed over the audio lines. The schematic diagram shown is for the Sennheiser Model MZN-6 power supply.

rectly connected to the transistor-amplifier stage, whose gain is limited to 12 dB by the use of negative feedback.

To eliminate the effects of rf oscillator noise, the oscillator circuit is crystal controlled. Noise in an oscillatory circuit is inversely proportional to the "Q" of the circuit, and because of the high "Q" of the crystal and of its stability, compensating circuits are not required, resulting in extremely low internal noise. In Model MKH404 microphone, only one stage of amplification is employed. For model MKH405 and 805 two stages are used as shown in Fig. 4-71L.

The output stage is in reality an impedance-matching transformer. In the Model MKH404, the output is adjusted for 100-ohms, and for the MKH 405 and 805, 10 ohms. With proper precautions, they may be operated into a balanced, symmetrical, or unbalanced input circuit. For an output impedance of 100 ohms, the load impedance must be 2000 ohms or greater; and for the 10-ohm output, 150 ohms or greater. Using an output impedance of 10 ohms permits long cable runs without an appreciable loss of the high frequencies. Radio-frequency chokes are connected in the output circuit to prevent rf interference and also to prevent external rf fields from being induced into the microphone circuitry. Shown in Fig. 4-71M is the frequency response, with manufacturing tolerances shown by the dashed lines; the polar response is shown in Fig. 4-71N. Two types of power supplies are available, mercury battery and ac operated. The battery supply will provide 50 to 60 hours of operation on one set of batteries. The operating voltage (12 Vdc) is duplexed over the audio lines. This is accomplished by connecting the battery across the audio line through two series resistors of 180 ohms each (Fig. 4-71O). With impedance-matching transformers, the circuit shown in Fig. 4-71P is used. Here a capacitor (C) is connected in series with the *split primary* of the input transformer to prevent the flow of current through the windings. If the transformer does not have a split primary, a capacitor is connected in each side of the line.

The schematic diagram for an ac power supply appears in Fig. 4-71Q. This supply is capable of operating up to six microphones simultaneously. The voltages are well stabilized, using zener diodes for each microphone supply circuit. The output level for the Model MKH405 is minus 27.5 dB re: 1 mw/ 10 dynes/cm^2. Frequency response is 40 to 20,000 Hz; THD distortion 0.5 percent; EIA rating 121.5 dB. Mode of operation is pressure gradient, with a cardioid directional characteristic.

4.72 What is a wireless microphone? —It is a miniature microphone attached to a frequency-modulated radio transmitter and is worn on the body of the user. For motion picture and television use, the microphone and transmitter are concealed. Two systems are in use. One system uses a small transmitter worn on a coat lapel or mounted on the housing of a miniature frequency-modulated radio transmitter. The second system employs a hand-held or lavalier microphone housing a transistor radio transmitter (Fig. 4-72A); this is the system to be discussed. The antenna is wrapped around the body of the user and he becomes a part of the antenna system. Or, in the instance of the hand-held microphone, the antenna is attached to the microphone-transmitter or it may be built into a helmet. The transmitted signal is picked up by a remote receiver. The maximum transmitting distance is about one half mile, depending on the frequency used and the antenna system. As a rule, the distance between the transmitter and the receiver is kept to within 200 feet or less. When several microphones are required, such as might be used with several characters in a play, a separate receiver is required for each microphone, each receiver tuned to a different frequency. The transmitter may be tuned from 25 to 45 MHz, with a signal-to-noise ratio of 60 dB, or better, for a plus-minus 20-kHz modulation swing.

The receiver consists of 11 tubes, including radio-frequency amplifiers, limiters, automatic frequency control (afc), intermediate amplifiers, audio stages, and an adjustable squelch control for reducing background noise. The unit also contains a built-in 36-inch collapsible antenna, or it may be used with a 72-ohm remote transmission line and antenna. A tuning eye is provided for setting the receiver to the exact frequency of the transmitter. The sensitivity is 1.5 microvolts for 20-dB quieting. The audio frequency response is 20 to 20,000 Hz. The output has an impedance of 150 or 600 ohms designed to operate directly into a mixer console. A loudspeaker is also included for monitoring purposes. The schematic diagram for the receiver portion is shown in Fig. 4-72B.

The microphone with its transmitter is completely transistorized. Its schematic diagram appears in Fig. 4-72C, and consists of a dynamic microphone element, with a frequency range of 80 to 14,000 Hz. The transmitter has a swing of 20 kHz at normal voice levels, with a power output of 40 milliwatts to the final amplifier, depending on the frequency and the antenna system used. The battery is a 6.5-volt mercury type, with an expected life of 20 hours. The whip antenna, shown on the side of the microphone in Fig. 4-72A, is used for personal interviews or when walk-

Fig. 4-72A. Vega Electronics Corp. wireless microphone system. The microphone and its fm radio transmitter are shown at the right, and the receiver at the left.

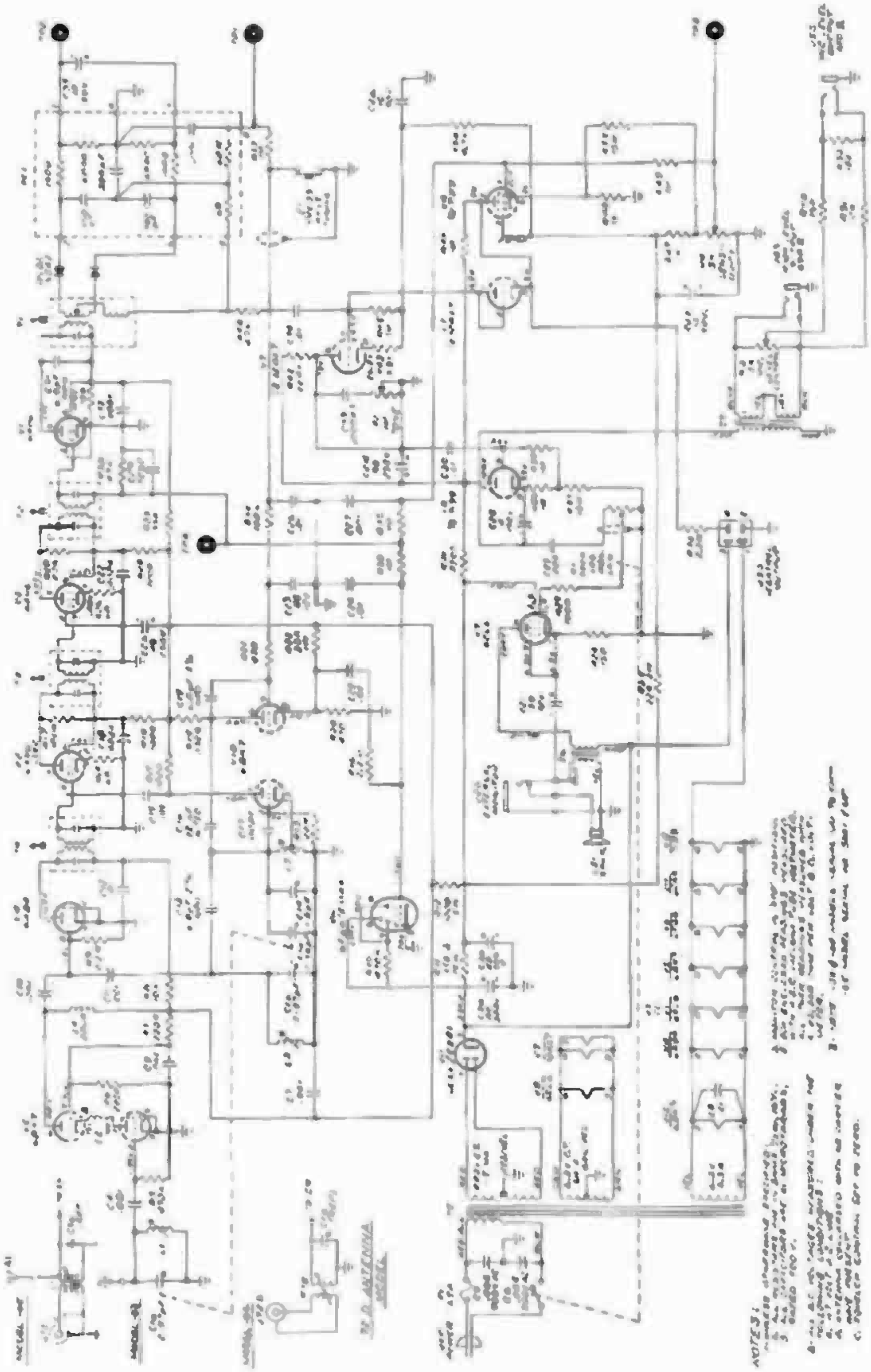


Fig. 4-72B. Schematic diagram for Vega Electronics Corp. wireless microphone receiver.

ing around in a crowd, and may be extended to 23 inches. Frequencies for wireless microphone operation in the United States are under the jurisdiction of the Federal Communications Commission, and such devices are operated in the general business frequencies, 33.14, 35.02 and 42.98 megahertz, or 26.25 for radio and television relay operation. Permission must be obtained from the

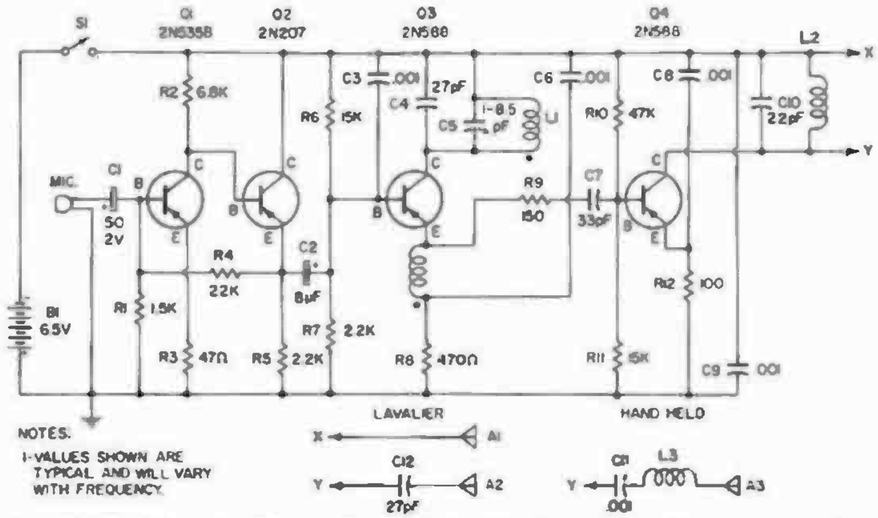


Fig. 4-72C. Schematic diagram for wireless microphone transmitter unit manufactured by Vega Electronics Corp.

FCC in advance, before any type of wireless microphone can be operated.

4.73 How far can a low-impedance microphone cable be run without a preamplifier?—Microphones with an output impedance ranging from 30 to 200 ohms can be run to a distance of 200 feet without seriously affecting the frequency response. It is a fairly safe rule to follow, that the lower the impedance of the line, the greater the distance it can be run. However, unless the cable is well shielded, the signal-to-noise ratio may be seriously affected. Also for such long runs, the cable must be kept clear of all ac lines and equipment. A good practice to follow when long runs are necessary is to insert a preamplifier in the cable run at about 100 feet. This preamplifier may be of transistor design with a gain of 40 to 60 dB adjustable in steps of 20 dB. (See Question 4.76.)

4.74 If an oscilloscope is connected to the output of a microphone, what is the relationship between the pressure wave and the image?—The vertical deflection will be proportional to the amplitude of the pressure wave at the diaphragm of the microphone.

4.75 What does the term open-circuit voltage mean?—It is the voltage measured with a high impedance vacuum-tube voltmeter at the output of an unterminated microphone. As a rule, preamplifiers designed to be used with microphones do not use a termination across the secondary of the input transformer. This method of operation increases the voltage at the control grid of the first tube in the preamplifier 6 dB or a voltage gain of 50 percent, which results in a substantial increase in the signal-to-noise ratio. Open-circuit operation is discussed in detail in Question 12.170.

4.76 What effect does the terminating impedance have on microphone characteristics? — Microphone specifications relative to their frequency-response characteristics and output level are based on an *open-circuit voltage* measurement. Such measurements are made using a high-impedance voltmeter, connected directly across the microphone output terminals (Fig. 4-76). However, in actual practice the microphone is fed into a preamplifier with an input transformer using an *unterminated* secondary. (See Question 12.170.)

If the microphone characteristic is measured using a load impedance approaching the microphone internal output impedance, the performance characteristics will differ from those specified for an unloaded measurement. A typical example of a *loaded* condition is operating the microphone into a low-

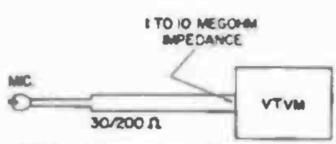


Fig. 4-76. Measuring the open-circuit voltage of a microphone.

level mixing circuit, a transistor preamplifier with a low input impedance, or into an input transformer with a terminated secondary. (See Question 9.2.)

If the microphone is terminated in a *resistive* load equal to its own impedance, only a drop in level will be experienced with no appreciable change in the characteristics. If the load varies with frequency, the overall response will be affected. This can easily occur if the interconnecting cables have an appreciable amount of capacitance such as might be encountered with a high-impedance microphone. The greater the cable capacitance per foot, the more the high frequencies will be attenuated. This is particulary true for microphones having an output impedance of 10,000 to 20,000 ohms. Generally about 18 to 20 feet is the maximum distance that can be run.

Mass-controlled microphones, such as the bidirectional and cardioid types, generally have an impedance curve that decreases sharply as the frequency is increased in the range of 100 to 2000 Hz. Connection of these microphones to a resistive load causes a change in both the output level and the frequency response, particulary at the low frequencies. For professional usage, microphones are fed into an open circuit preamplifier; thus, the problem of loading is eliminated. They are also operated at an impedance of 50 ohms, therefore cable lengths may be up to 600 feet with about a 1-dB loss at 10,000 Hz. (See Question 4.73) Operating a 50-ohm impedance into a 250-ohm input results in a 7-dB loss of output level, but no appreciable change in the frequency response. (See Question 4.98)

4.77 Describe a ribbon-velocity microphone with variable polar patterns.—A variable polar pattern microphone can sometimes be quite useful and convenient as it may be changed from a directional to a bidirectional or omnidirectional polar pattern by a simple adjustment. The microphone to be discussed is the RCA 77-DX MI-4045-F ribbon-velocity microphone (Fig. 4-77A). The general design of this microphone is similar to the unidirectional ribbon-velocity microphone discussed in Questions 4.59 and 4.65. As in the RCA MI-10001-C, the 77DX ribbon is divided into two sections, with the center grounded and the upper portion terminated in an acoustic impedance consisting of a small tube sealed at the back of the ribbon and connected with an acoustic labyrinth situated in the lower portion of the housing. An aperture in the back of the tube near the top is made variable by a rotating plate controlled from the exterior at the rear of the housing. Controlling the size of the aperture determines the types of polar pattern. When the aperture is completely open, the ribbon is not terminated by the acoustic labyrinth; therefore, both sides of the ribbon are exposed to the sound source. Under these conditions, the microphone operates as a ribbon-velocity microphone with a bidirectional polar pattern (Fig. 4-77B).

Closing the aperture to about half size, one half of the ribbon has both surfaces exposed to the sound pickup and therefore operates as a ribbon-velocity microphone; however, the other half of the ribbon is terminated in the acoustic labyrinth and operates as a pressure unit. Operating under these conditions, the voltages generated by the upper and lower sections of the ribbon, tend to reinforce each other for sounds arriving from the front, but tend to cancel each other for sounds arriving from the rear, resulting in a cardioid polar pattern (Fig. 4-77B). By closing the aperture completely, the two sections of the ribbon are terminated in the acoustic labyrinth and the polar

Fig. 4-77A. RCA Model 77DXMI-4045-F polydirectional microphone.

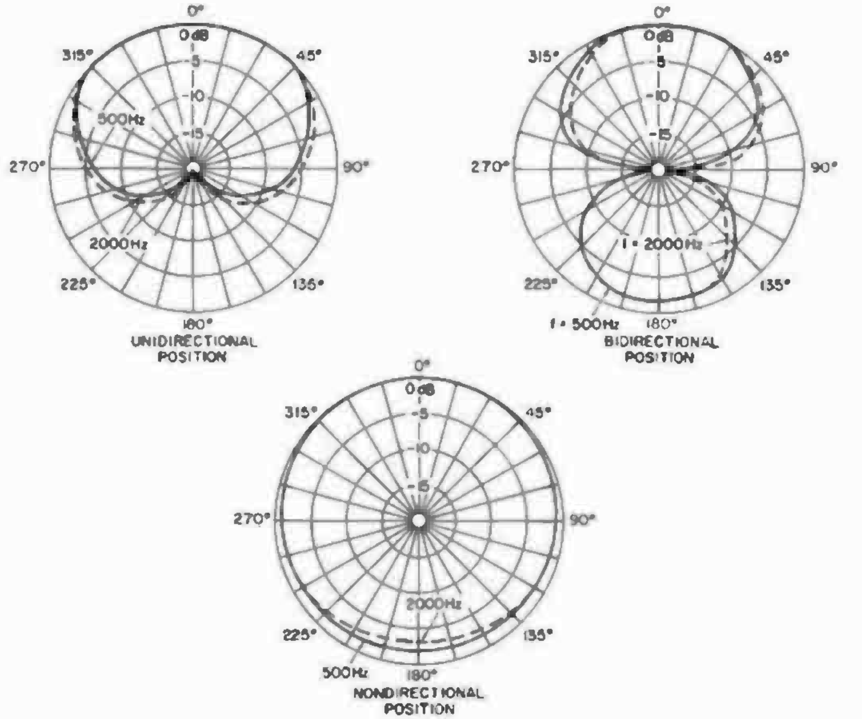


Fig. 4-77B. Polar patterns for RCA 77DXMI-4045-F polydirectional ribbon-velocity microphone.

pattern becomes omnidirectional (Fig. 4-77B). Because the aperture is variable, six polar patterns are possible. In the unidirectional position, the attenuation between front and rear is approximately 20 dB, or a ratio of 10:1.

The bottom of the housing contains an impedance-matching transformer, and a switch for selecting two types of low-frequency rolloff for voice, and a flat position for music. Position V1 is used for sound pickups within 1 foot of the microphone, aud V2 for distances greater than 1 foot (see Fig. 4-59H). Low-frequency attenuation is accomplished by connecting a reactor across the output of the microphone. The smaller the value of the reactor, the greater the attenuation will be.

The frequency response is shown in Fig. 4-77C. In the flat position the response is 30 to 20,000 Hz, within plus-minus 5 dB at the low and high ends, with output impedances of 30, 150 and 250 ohms. The output level ranges from 56 dBm (omnidirectional) to minus 50 dBm for the bidirectional position, for sound pressures of 10 dynes/cm². Hum level is minus 128 dB, referred to a field of 0.001 gauss. Microphones of this type, offering adjustable polar patterns are often termed polydirectional.

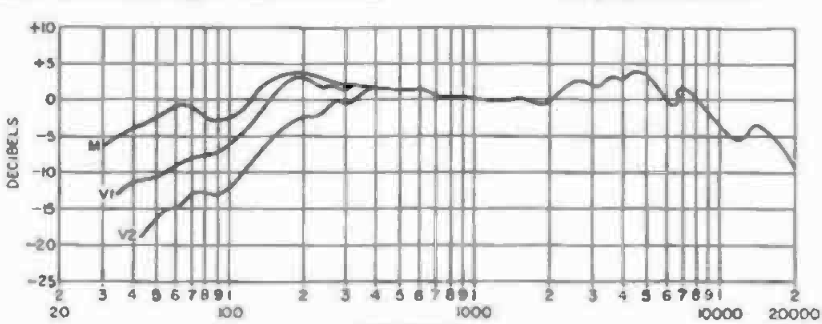


Fig. 4-77C. Frequency response for RCA 77DXMI-4045-F polydirectional microphone.

4.78 What is the impedance of a so-called high-impedance dynamic microphone?—The term high impedance is generally associated with crystal or ceramic microphones and will range from 100,000 ohms to 1 megohm. However, dynamic microphones rated high impedance are usually 40,000 ohms output impedance, although the moving-coil element may be on the order of 1.5 to 50 ohms. An output transformer of 40,000-ohms impedance with taps at 250, 150, and 50 ohms is included in the microphone housing.

4.79 What is a uniaxial microphone?—A unidirectional microphone in which the maximum response is in the principal cylindrical axis.

4.80 What is a cephloid microphone?—A mannikin head with microphones mounted in the head at the normal ear level to create a diffraction pattern and simulate human hearing. Such microphones are used in acoustical laboratories for binaural work.

4.81 What is a lavalier microphone?—A small dynamic microphone worn around the neck as a lavalier. A typical microphone of this type is the RCA BK-6B shown in Fig. 4-81A.

This microphone has been especially designed for correct speech balance when used informally in television broadcasting interviews and public address applications. The frequency response and directional qualities are engineered to complement human speech so that a correct frequency balance is maintained when the speaker is talking off mike.

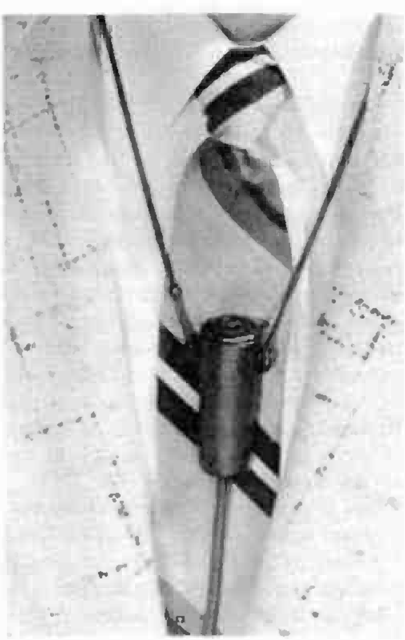


Fig. 4-81A. RCA BK-6B dynamic lavalier microphone, designed to be worn around the neck.

An internal resonator placed in front of the diaphragm reduces the high frequency emphasis while extending the upper frequency limit.

The output impedance may be set for 30, 150, or 250 ohms. The output level is minus 67 dB, for a pressure of 10 dynes/cm². The polar pattern is shown in Fig. 4-81B and the frequency response is shown in Fig. 4-81C.

4.82 What is a throat microphone?—One in which the diaphragm is actuated by being directly in contact with the external portions of the throat. Such microphones are widely used in aircraft for radio and internal intercommunication, or where the ambient noise level is high.

To obtain the maximum intelligibility, the high frequencies are attenuated. The unit may be of the moving coil or button design.

4.83 What is a sound-powered microphone?—A microphone constructed similarly to a dynamic microphone, with the exception that it generates considerably more output power and may be used without amplification over a con-

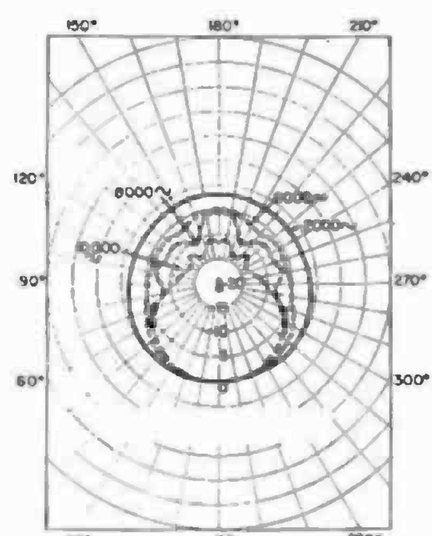


Fig. 4-81B. Polar pattern for RCA BK-6B lavalier microphone.

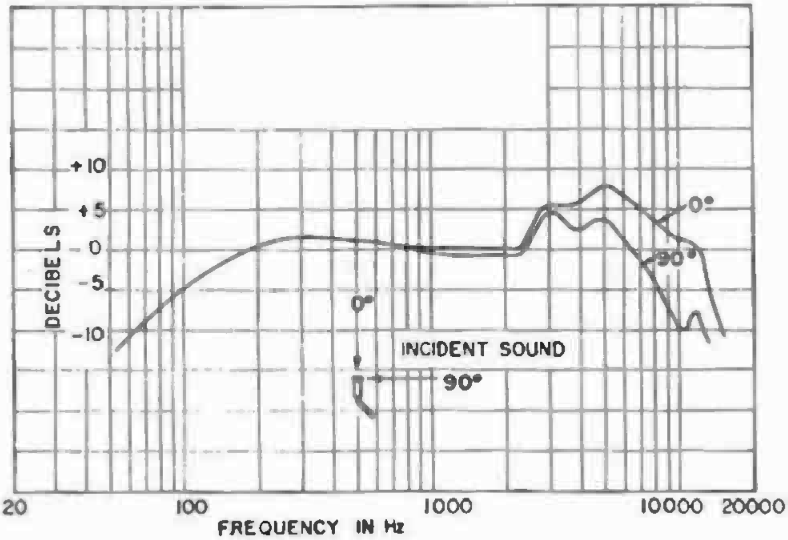


Fig. 4-81C. Frequency response of RCA 8K-68 lavalier microphone.

siderable distance. It is frequently used in intercommunication systems. The frequency response is limited to the voice frequency range.

4.84 How important is the phasing of microphones when making a multiple microphone pickup?—When two or more microphones are being used on the same channel and they are within ten feet of each other, they should be electrically in phase. If they are not, serious distortion may take place due to cancellation.

4.85 How are microphones phased? —The microphones to be phased are placed alongside each other and connected to their respective preamplifiers and mixer inputs. While someone speaks into the microphones, one mixer pot is adjusted for a normal output level as indicated on a VU meter. The setting is noted aud the pot closed. The same adjustment is made for the second microphone and the setting of that pot noted. Now both pots are opened to the above settings and, if the microphones are out of phase, the quality of reproduction will be distorted and there will be a distinct drop in level. Reversing the electrical connections to one microphone will bring them into phase. A second test is then made similar to the first. If the microphones are in phase, the output level will be greater and undistorted.

If the microphone is of the bidirectional type (figure-8) it may be reversed 180 degrees to bring it into phase, and later corrected electrically. If the microphones are of the directional type, only the output or cable connections can be reversed. After phasing a bidirectional microphone, the rear should be marked with a white stripe for future reference.

4.86 Describe the methods used for calibrating microphones.—Generally a microphone is calibrated using one or more of four calibrating instruments—the pistonphone, thermophone, electrostatic actuator, or a reciprocity calibrator.

Pistonphones are used for calibrating below 150 Hz, and consist of a pressure chamber, tightly sealed to the front of the microphone housing to be calibrated (Fig. 4-86A). The voltage developed at the output of the microphone for a given pressure at the diaphragm is noted. The pistonphone is then disconnected, and a voltage of the same frequency is connected in series with the microphone head and adjusted for the same output level as developed by the pistonphone. The response E_o/P is then the ratio of the second voltage to the applied pressure.

Two practical applications of the pistonphone are shown in Figs. 4-86G and H. Referring to Fig. 4-86G, the pistonphone consists of a sealed glass jar, with a mechanically driven piston in the base, which produces known sinusoidal fluctuations in air pressure. The actual pressure is calculated from the volume of the chamber, the dimensions of the

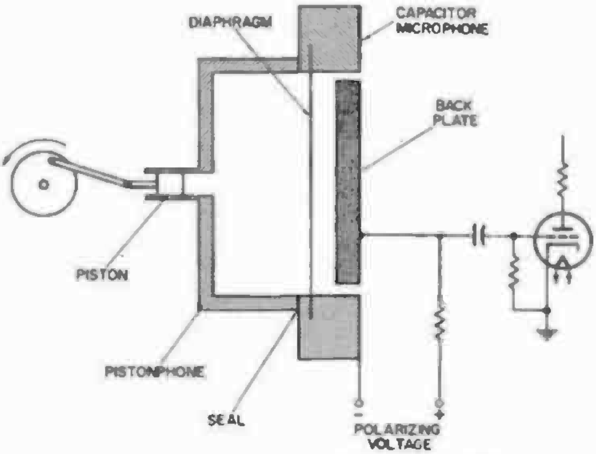


Fig. 4-86A. Pistonphone microphone calibrator.

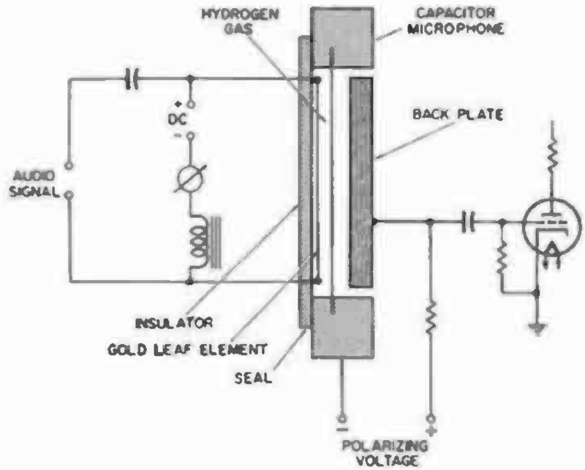


Fig. 4-86B. Thermophone microphone calibrator.

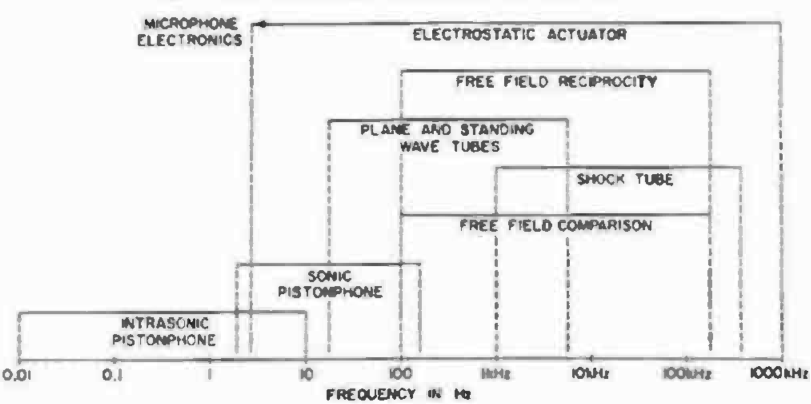


Fig. 4-86C. Frequency range of instruments used in the calibration of microphones.

piston, its stroke, and the constants of the air. With devices of this type, frequencies of a fraction of a cycle are possible, with sound-pressure levels of 94 to 115 dB, re: 0.0002 dyne/cm².

A second design appears in Fig. 4-86H, and is somewhat similar to a dynamic-loudspeaker mechanism. The small piston in the chamber at the left produces the pressure for the measurement. A calibrated eyepiece and a marker on the piston shaft permit the calibration of the piston movement. A capacitor microphone is shown under measurement, with its electronics and case mounted at the right end of the pressure chamber. Noise contributed by the electronics may be measured by replacing the microphone head with a low-leakage capacitor of the same value capacitance as the head, or the microphone may be placed in an acoustic chamber constructed of two lead cases, one inside the other, separated by acoustic absorbent. The noise level is then measured at the output, using a vacuum-tube voltmeter.

The thermophone, shown in Fig. 4-86B, is constructed of one or more very thin gold-leaf strips, supported on insulated blocks in front of the microphone diaphragm. A measured, steady direct current is passed through the gold-foil strips with the audio fre-

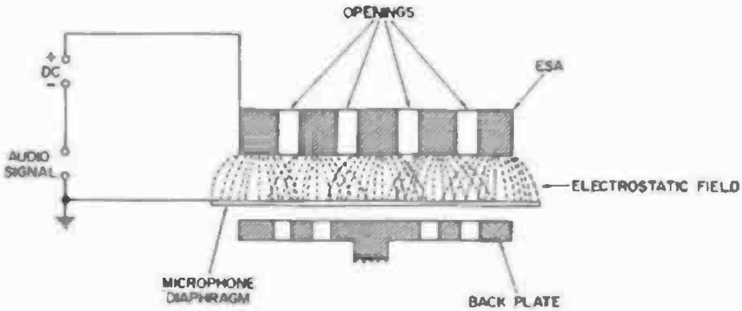


Fig. 4-86D. Cross section of an electrostatic actuator used in the calibration of capacitor-type microphones. (Courtesy, LTV Research Center Western Division)

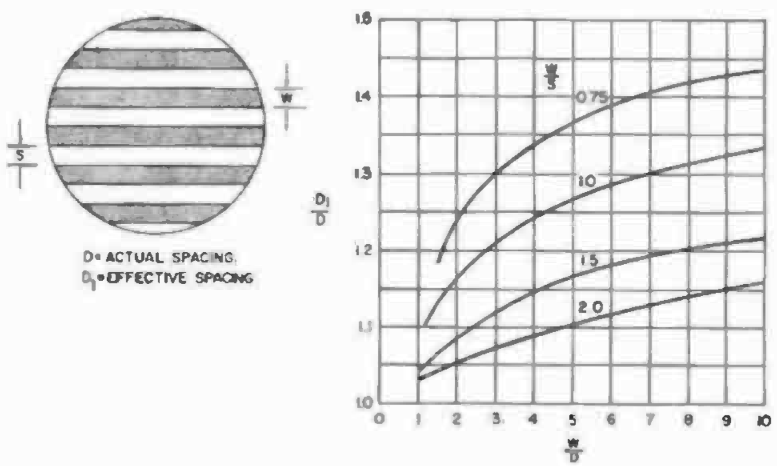


Fig. 4-86E. Effect of actuator-plate geometry on the effective spacing. (Courtesy, LTV Research Center Western Division)

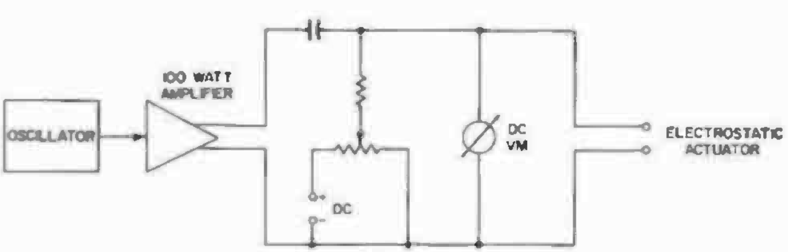


Fig. 4-86F. Driving circuit for electrostatic actuator. The amplifier should be capable of developing 100 watts of power with low distortion. The dc polarizing voltage is approximately 1500 volts. (Courtesy, LTV Research Center Western Division) quency signal superimposed on the dc. The cavity of the thermophone is filled with hydrogen gas to shift standing waves in the cavity beyond the range of measurement. Thus, the variation of pressure in the cavity occurs at the applied frequencies.

Microphones designed for use in space require frequency calibrating ranges from 0.01 to 1 MHz per second, with sound pressures up to 195 dB, re: 0.0002 dyne/cm², also at extremely high and low temperatures, with low static pressures equivalent to those found in high altitudes. The various methods used for microphone measurements and their frequency range is given in the tabulations in Fig. 4-86C. Both the reciprocity and pistonphone techniques provide a high degree of accuracy, the pistonphone being capable of measurements down to 0.01 Hz, with an upper limit of 15,000 Hz, using a closed coupler, and up to 100 kHz in a free field. Calibrations at high pressure levels are performed in standing and progressive wave tubes, the high-intensity pistonphone, and shock tube. However, the above instruments are difficult to apply to calibrations at environmental conditions similar to that experienced in the field. These difficulties have been overcome to a great extent by further development of the electrostatic actuator (Fig. 4-86D) which is used primarily for the calibration of capacitor microphones.

The electrostatic actuator was described by Olson in 1931, by Ballantine in 1932, and further developed by Brown and Dahlke. The actuator consists of a slotted perforated flat metal plate, which is insulated and placed parallel to the diaphragm of the microphone at a known distance. A direct-current voltage applied between the actuator plate and the diaphragm of the microphone to be calibrated generates an electrostatic force, simulating a sound pressure, at the surface of the diaphragm. The sound pressure thus generated is the function of the product of the applied voltage, the square of the effective spacing, the permittivity of the vacuum, and the dielectric constant of the gas between the plates. Since the influence of temperature and static pressure on the dielectric constant is considerably less thau the accuracy in determining the pressure, the simulated

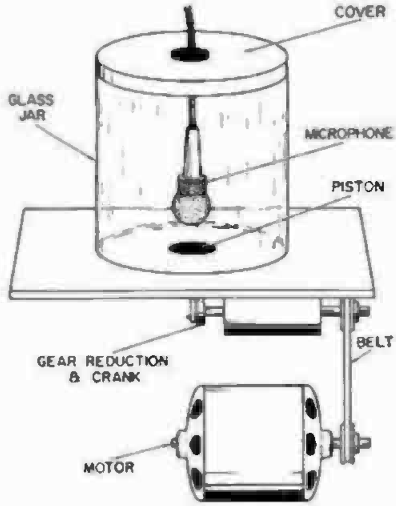


Fig. 4-86G. Pistonphone for calibrating microphones. A piston in the bottom of the glass jar is driven by crank and gear reduction, from the motor at the floor level. The microphone to be calibrated is suspended in the jar over the piston.

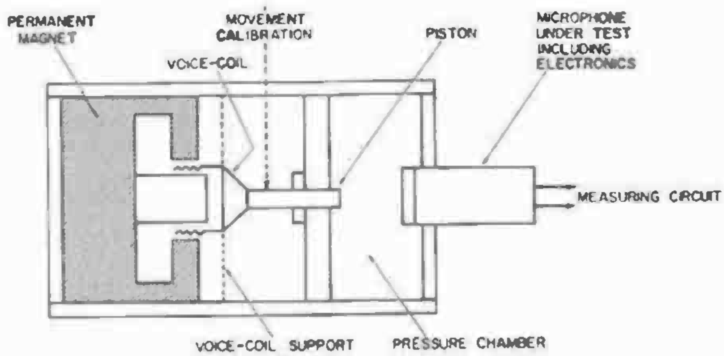


Fig. 4-86H. Pistonphone built along the lines of a dynamic loudspeaker. The piston is actuated by the movement of the voice coil, creating a sinusodial pressure in the chamber containing the head of the microphone (after Glover and Baumzweiger).

sound pressure is not affected by the environmental conditions. In addition, the sound pressure is not a function of the applied frequency.

The effect of the actuator plate and its geometry is shown in Fig. 4-86E. If the effective spacing of the slot D_e, which is not identical to the actual spacing, can be determined, the effective sound pressure can be expressed:

$$P_e = \frac{KE_o}{d^2} E_{oe}$$

where (D) is the effective spacing between plates, E_o permittivity of free-space and equals 8.85×10^{-12} farads/m, K the dielectric constant of the insulator between the plates, and E the dc potential between the plates. For a complete mathematical treatment pertaining to the actuator, the reader is referred to the reference.

Since the simulated sound pressure is produced by an electrostatic force, the signal driving the system will determine the low-frequency response of the electrostatically driven diaphragm. If the driving system cutoff frequency is low enough, the output voltage of the microphone system will be the response of the microphone electronics. To obtain the low-frequency response, the front to back vent of the microphone must be present in the sound field. The electrostatic actuator does not fulfill this requirement.

The schematic diagram for the actuator driving system is shown in Fig. 4-86F. The maximum sound pressure level (SPL) for 5-mil spacing of the actuator and the microphone diaphragm, and a 20-percent safety factor in breakdown voltage is 127 dB. For 2-mil spacing, the SPL rises to 134 dB. However, for this spacing, some difficulty may be experienced in keeping the actuator and diaphragm parallel. To reduce harmonic distortion below 2.5 percent, the ratio of the dc polarizing voltage to the ac signal voltage must be greater than 10:1; as 150 volts of signal is a reasonable value, the dc voltage will be on the order of 1500 volts. Frequency measurements, using this device, are possible to within plus or minus 1 dB. Microphone calibration using the reciprocity method is discussed in Section 23.

Considerable work has also been done in the testing of microphones, termed real voice testing. In this test, a person speaking in a normal tone of voice replaces the usual sound pressure generating device. As the speaker talks, a frequency analysis is made and the average output level established. Thus, the average output for a number of speakers is the real voice calibration. The test equipment consists of the microphone to be tested, preamplifier, and a magnetic tape recorder. Portions of the recording are made into a loop, and the output fed to an octave-band analyzer and graphic level recorder. The result of this test is the average power output from the microphone. It has been found for some types of research, the real voice test is more satisfactory than the usual pressure calibration.

4.87 Describe how the linearity of a capacity microphone is determined at high intensities.—It is tested by measuring its linear response to a step function generated by the use of a shock tube at pressure levels equivalent to 5 pounds per square inch. The linearity of an Altec-Lansing Model 21-BR-200 high-intensity microphone is shown in Fig. 4-87A, for SPL of 140 to 180 dB, re: 0.0002 dyne/cm² at a frequency of 400 Hz. To provide a primary calibration, the sensitivity of a group of eight microphones was measured, using a pistonphone at an SPL of 124 dB at frequencies up to 400 Hz. Sound pressures are then generated in a standing-wave tube to provide nearly undistorted sinusoidal pressures at extremely high levels.

Essentially, a standing-wave tube consists of two cylindrical tubes having lengths which are nonharmonically related. The tube is driven by an electrodynamic driver, which produces a somewhat distorted signal at the amplitudes necessary to generate sound pressures at the desired levels. This distortion is suppressed by the design of the standing wave tube, which acts similar to an acoustic bandpass filter. The standing-wave tube measurements are performed at a single frequency associated with the tube resonance. To verify the microphone linearity at the higher frequencies, shock-tube techniques are employed.

A microphone is placed at the end of the shock tube. A shock is generated by bursting a diaphragm which isolates the drive section of the tube. The mi-

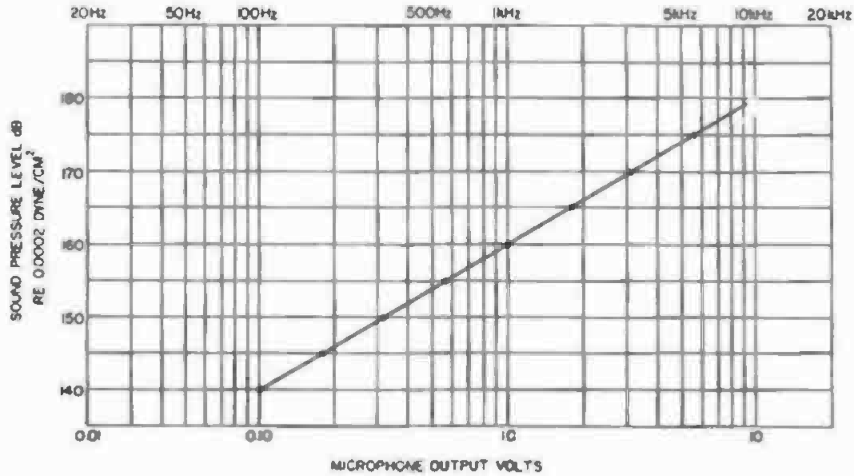


Fig. 4-87A. Standing-wave tube linearity calibration of Altec-Lansing capacitor microphone Model 21-BR-200. Sensitivity −86 dB re: 1 volt/dyne/cm².

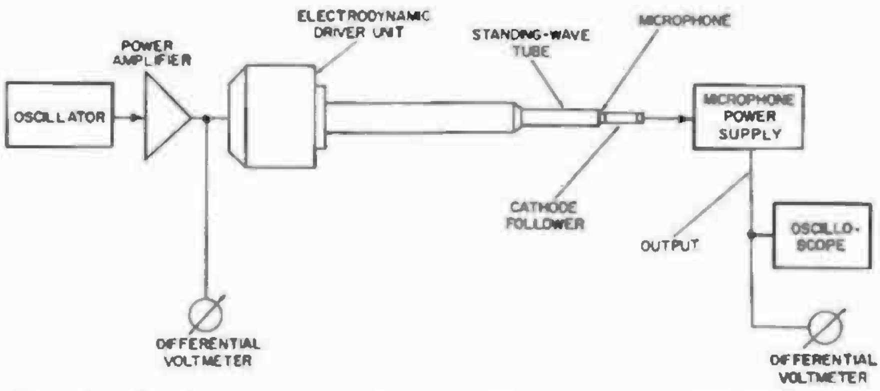


Fig. 4-87B. Standing-wave tube and its associated equipment used in the linearity measurement of microphones. (Courtesy, LTV Research Center Western Division)

crophone experiences a sudden rise in pressure associated with the reflection of the shock wave at the end of the driven section of the tube. The rise-time, overshoot, and decay time of this response exhibit the linear characteristics of a simple system transient response. The system resonance and damping constant are established from this transient response to yield the steady-state response characteristics. The basic test setup for using a standing-wave tube is shown in Fig. 4-87B. A cross-sectional view of a high intensity microphone head is given in Fig. 4-115A. Details for the use of standing-wave and shock-tube devices are given in the reference.

4.88 What does the term "gunning a microphone" mean?—To point the microphone at the source of sound. If the microphone is suspended from a boom on a turret head, it is turned towards the actors as they speak, to secure a uniform sound quality. As the microphone is generally moved quite rapidly, it is referred to as "gunning."

4.89 What happens when a ribbon-velocity microphone is moved rapidly through the air?—A low-frequency flutter is generated due to the passage of air between the ribbon and the pole pieces. It is not practical to use standard ribbon microphones on a boom and attempt to gun them. Ribbon-velocity type microphones used for motion pictures and television are especially designed to be used on a boom and may be gunned without producing noise.

4.90 Describe a microphone wind screen, its construction, and effectiveness.—A wind screen is a device placed over the exterior of a microphone for the purpose of reducing the effects of

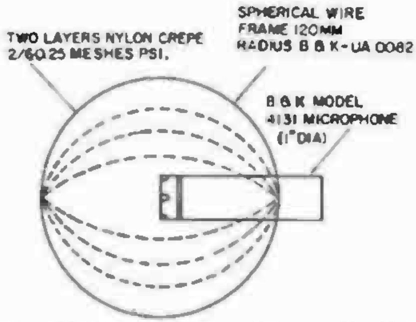


Fig. 4-90A. Typical silk-covered wind screen and microphone. (Courtesy, B and K Technical Review.)

wind noise when recording out of doors, or when panning or gunning a microphone. The screen consists of a wire frame covered with silk or a special type foam-rubber composition. Microphones designed for television and motion picture recording are generally constructed with the wind screen as an integral part of the microphone. (Wind screens are also referred to as wind gags.)

With a properly designed wind screen, a reduction of 20 to 30 dB in wind noise can be expected, depending on the sound pressure level at the time, wind velocity, and the frequency of the sound pickup. Wind screens may be used with any type microphone and they vary in their size and shape. Two types of wind screens employing a special type foam-rubber manufactured by Electro-Voice, and marketed under the trade name *Acoustifoam* are shown in Figs. 4-39A and 4-97E. This material has no effect on the high frequency response of the microphone because of its porous nature. *Standard styrofoam is not satisfactory for wind screen construction because of its homogenous nature.*

A cross-sectional view of a typical wind screen employing a wire frame covered with nylon crepe for mounting on a one inch diameter microphone is shown in Fig. 4-90A. The effectiveness of this screen as measured by Dr. V. Brüel of Brüel and Kjaer is given in Figs 4-90B, C, D, and E. When recording dialogue using a wind screen, its effectiviness is increased if dialogue equalization is used as discussed in Questions 4.114, 6.122, and 18.81.

4.91 What is a rain screen and how is it constructed?—During the filming of rain and snow scenes, a protective covering called a rain screen is placed over the microphone, as shown in Fig. 4-91, to protect the microphone and reduce the sound of the raindrops. The screen consists of an inverted shell approximately 18 inches in diameter. The exterior is covered with hair felt and metal fly screen, to break the raindrops into fine particles. The interior is cov-

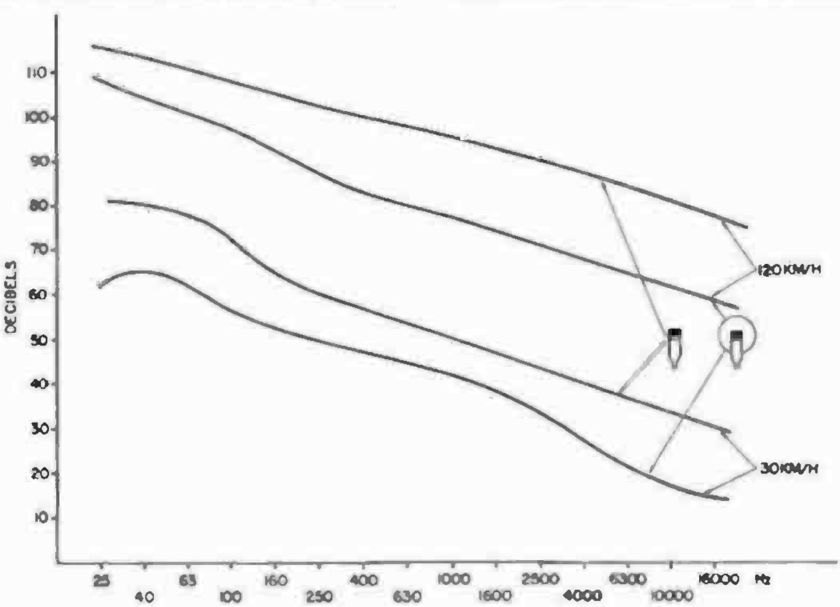


Fig. 4-90B. Wind noise as a function of frequency, measured in ⅓ octaves, but with the wind direction parallel to the membrane. (Courtesy, B and K Technical Review.)

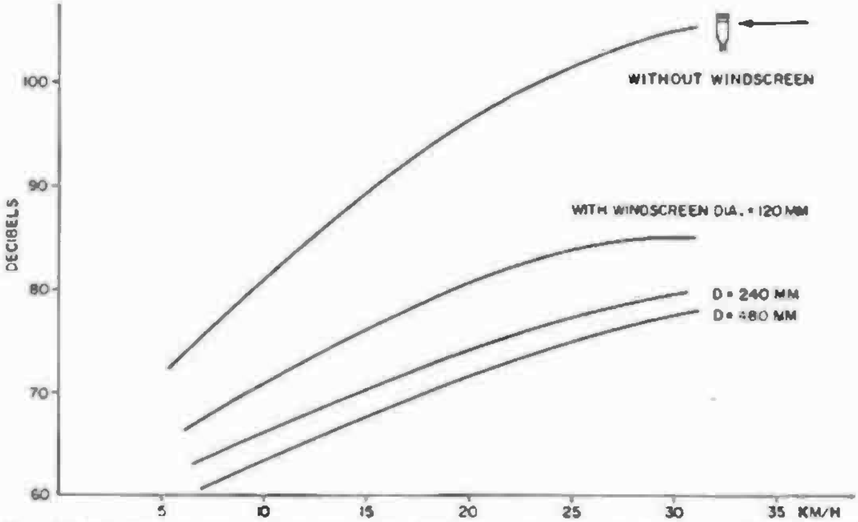


Fig. 4-90C. Wind noise measured with different sizes of windscreens. (Courtesy, B and K Technical Review)

ered with oilcloth and a layer of fly screen. The microphone is suspended inside the screen on a flexible support.

4.92 What type of microphone is recommended for dialogue recording?— It is the practice in the professional field to use either dynamic, capacitor, or ribbon-velocity microphones having *directional characteristics*. This is an absolute necessity to reduce unwanted noise pickup at the sides and rear of the pickup area. Bidirectional microphones should *not* be used, if the microphone is to be suspended from a boom. It is common practice in the broadcast field to mount bidirectional microphones on a floor stand, placing one person on each side of the microphone. Dialogue equalization should always be used for voice pickup, or the microphone should be connected with the internal *voice-filter* in the circuit. Dialogue equalization is discussed in Questions 4.114, 6.122, and 18.81.

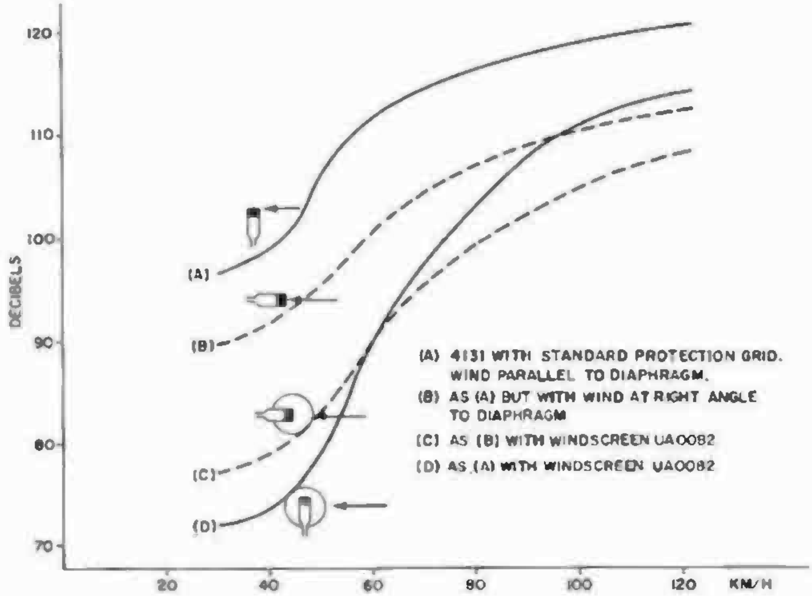


Fig. 4-90D. Wind noise as a function of wind speed in the range 20 Hz to 20kHz. (Courtesy, B and K Technical Review)

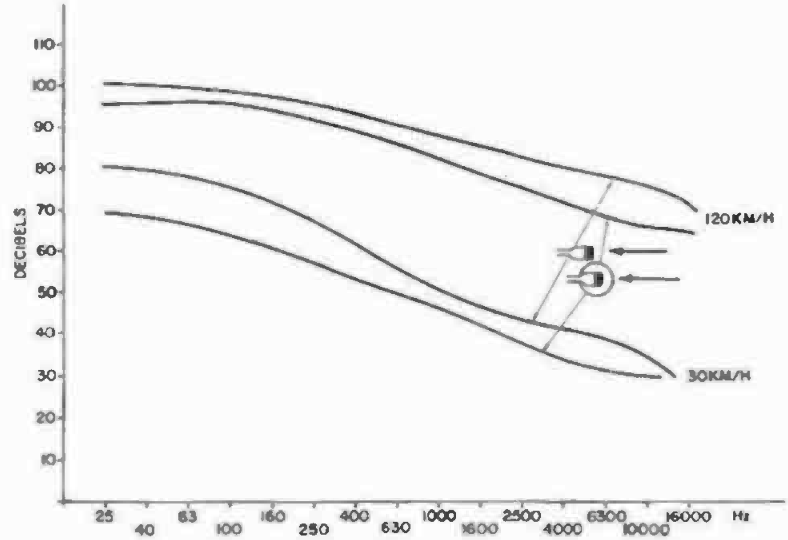


Fig. 4-90E. Wind noise as a function of frequency, measured in ⅓ octaves, with the wind direction at right angles to the membrane. (Courtesy, B and K Technical Review)

4.93 What types of microphones are recommended for music recording?—Capacitor, ribbon-velocity, and dynamic types. Similar types of microphones employing directional characteristics are preferred to aid in the separation of various sections of the orchestra. For stereophonic recording, the dual-type microphone similar to those discussed in Question 4.48 is used.

4.94 How close may two microphones be placed when feeding the same recording channel?—As close as their physical dimensions will permit, provided they are in phase electrically with each other. If they are not in phase, the output will be distorted and show a decided loss in output level. (See Questions 4.84 and 4.85.)

4.95 What is the standard pin count for microphone plugs? — It is standard practice to employ a recessed male plug on the microphone. Microphone cables are then made up using a female plug on one end and a male plng on the other. The pin count may be either of the two that are shown in Fig. 4-95. The pin count should be determined before attempting to put the microphone in service. *Under no circumstances is a volt-ohmmeter or buzzer ever to be applied to the output connections of a microphone.* To do so will severely damage the internal strncture. (See Question 24.100.)

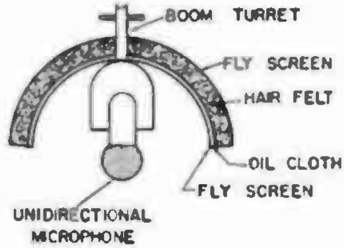


Fig. 4-91. A microphone rain screen.

Fig. 4-95. Pin count used on microphone cables.

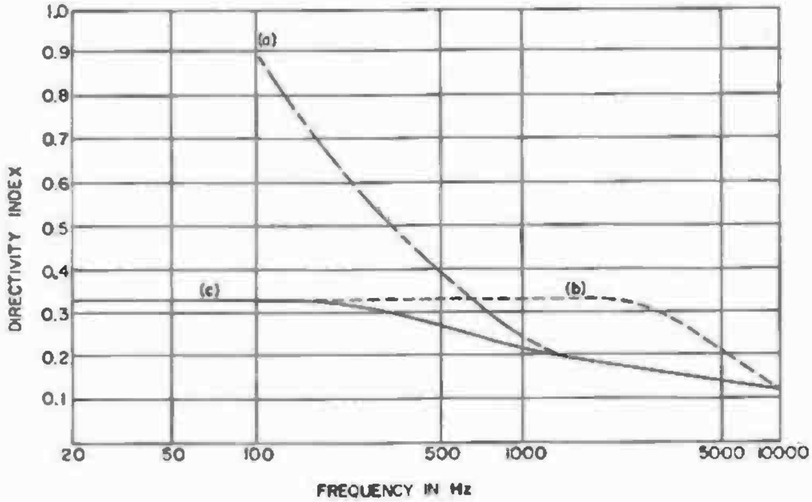


Fig. 4-97A. Directivity index for: (a) 11-inch line microphone, (b) cardioid microphone, (c) Electro-Voice Model 642 Cardiline dynamic microphone.

4.96 How may the signal-to-noise ratio be increased on a long microphone cable run?—See Question 4.73.

4.97 Describe the basic principles of a cardioid in-line directional microphone.—Both television and motion picture production have suffered greatly from the fact that to pick up good sound, the microphone must be reasonably close to the person speaking. On a motion picture set, this causes difficulties for both the camera man and the sound man. Also, getting in close restricts the movement of the microphone boom and is often the cause of poor pickup and excessive camera noise. These problems have been overcome to a great extent by the development and introduction of the cardioid in-line directional microphone, described by Olson in 1938.

Two of the most important characteristics of any microphone are its sensitivity and directional qualities. Assuming a constant sound pressure source, increasing the distance of the microphone from the source requires an increase in the gain of the amplifying system after the microphone. This is accompanied with a decrease in signal-to-noise ratio and an increase in environmental noises, such as reverberation and background noise, to where the indirect sound may equal the direct sound. The pickup then deteriorates to where it is unusable. Distance limitations can be overcome by increasing

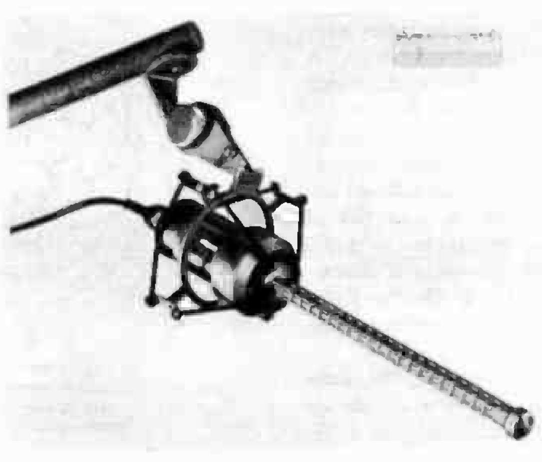


Fig. 4-97B. Electro-Voice Model 642 Cardiline (in-line) directional microphone, with boom hanger.

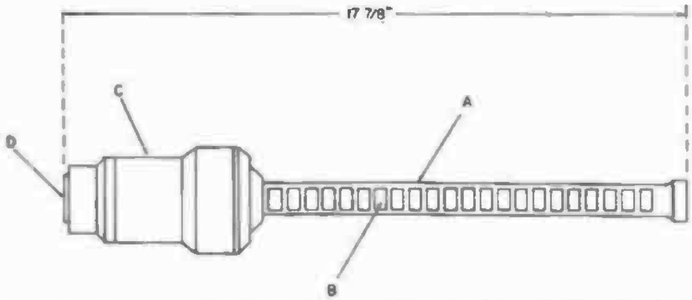


Fig. 4-97C. Basic components of the Electro-Voice Model 642 Cardiline directional microphone.

the sensitivity of the microphone, and the effect of reverberation can be lessened by increasing the directivity of the pattern. The in-line microphone has these two desirable qualities.

Over the years, the most commonly used microphone for boom operation and the recording of dialogue has been a gradient microphone with a cardioid directional polar pattern, as shown in Fig. 4-7, (c) and (d). In addition to the polar plot, the directional characteristics of a microphone may be described by a ratio termed directivity index. Directivity index is a ratio of output voltage from a microphone in a sound field that arrives at the microphone from all directions, to the output voltage of an omnidirectional microphone (with equal axial sensitivity) in the same sound field. The directivity index is a measure of nonaxial response; the lower the directivity index the sharper the polar response.

For microphones of the bidirectional and cardioid type, the index is 3:1; because of baffle effects (see Question 4.52), the polar response becomes narrower at the higher frequencies. A graphical plot of the directivity index versus frequency response for an 11-inch cardioid line microphone, and for a 1½-inch dynamic gradient microphone with a cardioid polar pattern are shown by the curves (a) and (b) of Fig. 4-97A. Curve (c) is for a Model 642 cardioid in-line directional microphone, manufactured by Electro-Voice under the trade name *Cardiline*, pictured in Fig. 4-97B.

The basic components of the construction of this microphone are shown in Fig. 4-97C. Line-tube (A) has a ⅛-inch slot milled its entire length and a group of ports which act as a linear tapering acoustic resistance; they are equally sensitive to equal sound pressures, and will cause equal voltages to be generated at the output of the dynamic transducer unit (C). Since the ports are acoustically connected to the transducer unit by the common tube, acoustic delays are introduced ahead of the transducer element. When placed in a plane-wave sound field, this equally sensitive line with variable delay produces wave interference in the common cavity at the front of the transducer unit. The magnitude of this interference will depend on the angle between the plane wave and the axis of the tube. The directivity of a line microphone is a function of frequency; the lower the frequency, the broader the polar pattern (Fig. 4-97D). Sensitivity is achieved by a rather heavy magnetic structure consisting of an Indox V and Armco magnetic iron, approximately 2½ inches in diameter. Dimensions of this size can be used; because of the in-line tube no baffle effects occur, the directivity being controlled by the

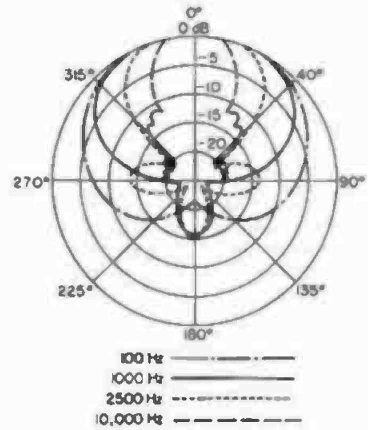


Fig. 4-97D. Polar response pattern for Electro-Voice Model 642 Cardiline directional in-line microphone.

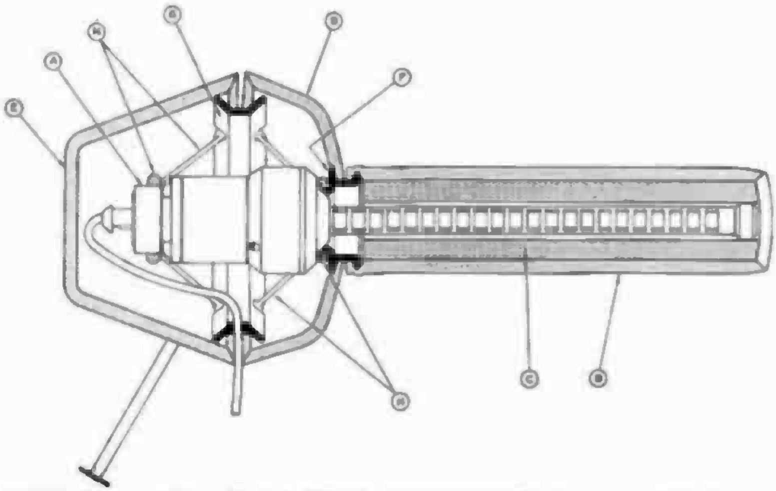


Fig. 4-97E. Electro-Voice Model 642 Cardiline microphone mounted in its support hanger and Acoustifoam wind screen.

openings in the tube structure. The diaphragm of the dynamic unit is 1¼ inches in diameter with a ¾-inch voice coil.

The drawing in Fig. 4-97E shows the microphone with its *Acoustifoam* wind screen attached. The microphone is supported at (A) by two rubber rings and arms (H) in a metal ring (G). A second ring (F) seals the front and rear wind screens (B) and (D). The third wind screen (E) encloses the rear of the microphone assembly. Experience indicates the rear wind screen must be in place to achieve the maximum effect of the wind screen assembly. The *Acoustifoam* wind screen has no effect on either the frequency response or the polar pattern. Since the screen is quite porous, air may be blown through quite easily. Dirty screens may be washed in clear water.

The frequency response is shown in Fig. 4-97F. Three positions are available for adjusting the low-end frequency response; flat, minus 5 dB, and minus 10 dB at 100 Hz, with reference to 1000 Hz. The output level is minus 48 dB re: 1 milliwatt/10 dynes/cm². Hum level is minus 120 dB re: 0.001 gauss. The directivity index is 6:1 with an included angle of pickup of 80 degrees. The output impedance may be adjusted for 50, 150 or 250 ohms. Because of the directional characteristics and sensitivity, this microphone may be operated from 2 to 4 feet farther from the sound source than the conventional microphone.

4.98 What is the effect of a long microphone cable on frequencies above 1000 Hz?—The high frequencies will be attenuated. Assume a microphone with an output impedance of 20,000 ohms is

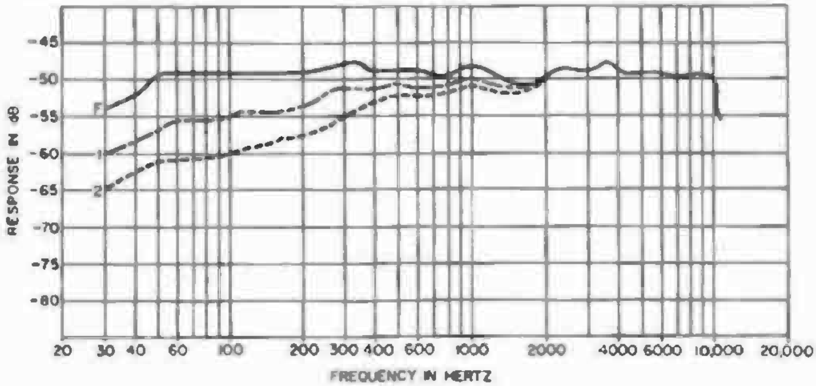


Fig. 4-97F. Frequency response for Electro-Voice Model 642 Cardiline directional in-line microphone.

connected to a line 30 feet long. The loss at 10,000 Hz compared to 1000 Hz is approximately 3 dB. Increasing the line to 100 feet will increase the loss at 10,000 Hz to 16 dB. Decreasing the impedance to 250 ohms or less, the loss at 10,000 Hz is negligible. For long runs an impedance of 50 to 150 ohms maximum is recommended.

4.99 How can microphone cables on a stage be prevented from picking up ac hum?—The microphone cables should be kept as far away as possible from ac circuits. If it is necessary to cross an ac line, do so at right angles, not parallel to it, or support the microphone cable a few feet above the cables carrying the alternating current. Microphone cables should be kept at least 6 feet from arc lights and similar equipment. Arc lights will also create mechanical noise due to the feed mechanism. Therefore, it is desirable that unidirectional microphones be used. The treatment of arc lamp noise and generator ripple is discussed in Question 25.117.

4.100 What is a sound concentrator?—A wooden parabolic reflector used with a microphone to obtain a highly directive pickup response. The microphone is mounted in the center of the reflector, as shown in Fig. 4-100A, forward of the dish center. The microphone is focused by moving it in or out of the reflector for maximum pickup. This type concentrator is often used to pick up a horse race or a group of people in a crowd.

The greatest gain in sound pressure is obtained when the reflector is *large* compared to the wavelength of the incident sound. With the microphone in focus, the gain is the greatest at the midfrequency range. The loss of high frequencies may be improved somewhat by defocusing the microphone a slight amount, which also tends to broaden the sharp directional characteristics at the higher frequencies. A bowl 3 feet in diameter is practically nondirectional below 200 Hz, but very sharp at 8000

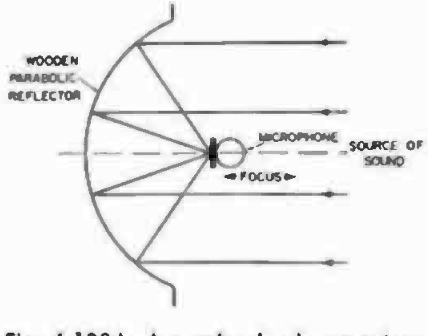


Fig. 4-100A. A wooden bowl concentrator for directional microphone pickup.

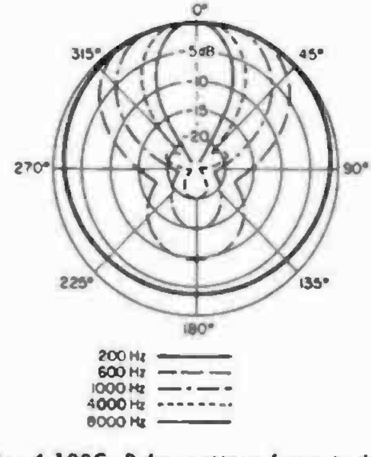


Fig. 4-100C. Polar pattern for a typical parabolic wooden bowl concentrator.

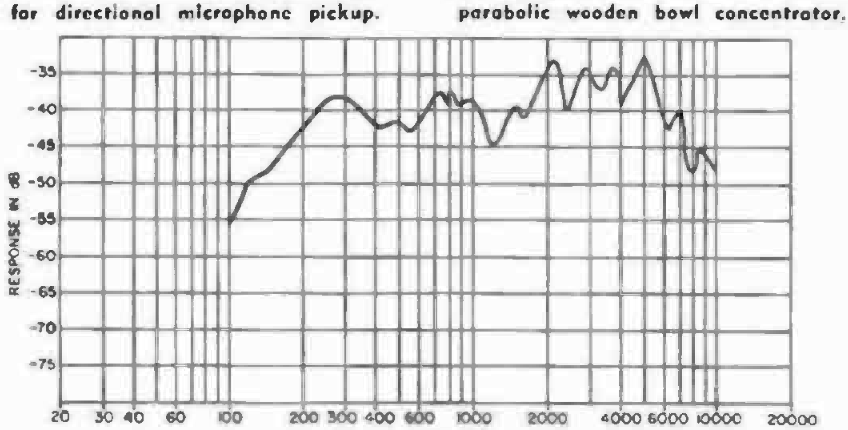


Fig. 4-100B. Frequency response of microphone and parabolic wooden concentrator.

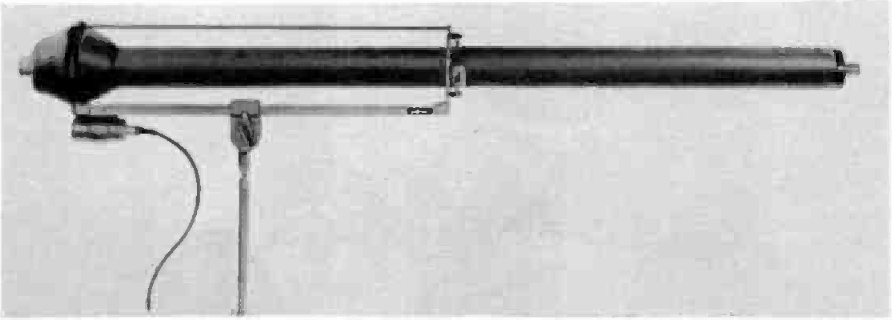


Fig. 4-101A. Electro-Voice Model 643 long range pickup, highly directional Cardiline dynamic microphone, with Acoustifoam wind screen.

Hz. For a diameter of 3 feet, the gain over the microphone without the bowl is about 10 dB, and for a 6-foot diameter, approximately 16 dB. The frequency response for a typical wooden concentrator bowl is shown in Fig. 4-100B, and its polar pattern in Fig. 4-100C.

4.101 Describe a highly directional microphone suitable for long range pickup.—In Question 4.97 the basic principles of an in-line type microphone and a commercial model with such characteristics were discussed. This question will deal with a commercial model of a super in-line microphone used for picking up a group of persons in a crowd from the roof of a nearby building, following a horse around a race track, picking up a band in a parade, and other hard-to-get distant pickups.

The Electro-Voice Model 642 microphone discussed previously has a directivity index ratio of 6:1 with an included pickup angle of 80 degrees. The microphone to be discussed here is the Electro-Voice Model 643 (Fig. 4-101A) which has an included pickup angle of 40 degrees, with a 30:1 directivity index ratio. The basic design is the same as that for the Model 642, except for its much sharper angle of pickup, as it has a longer in-line tube with a greater number of in-line ports.

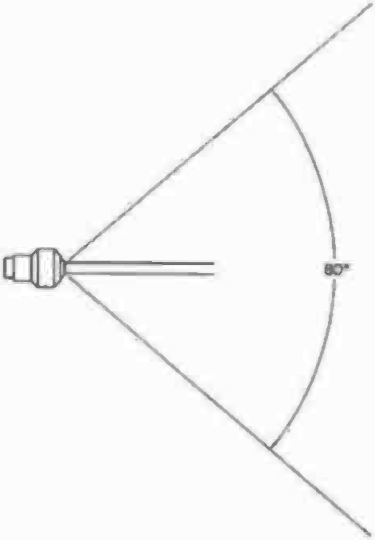


Fig. 4-101C. The included angles of an in-line microphone.

For a better understanding of Model 643, the included angles of pickup for both models, 642 and 643, are compared with that of an omnidirectional microphone having an included angle of pickup of 300 degrees, in Figs. 4-101B, 4-101C and 4-101D. The cone of pickup

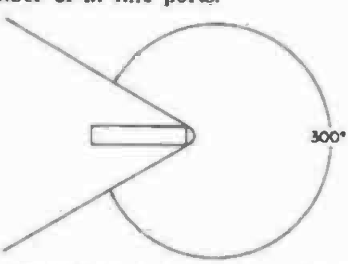


Fig. 4-101B. Included angles of omnidirectional microphone.

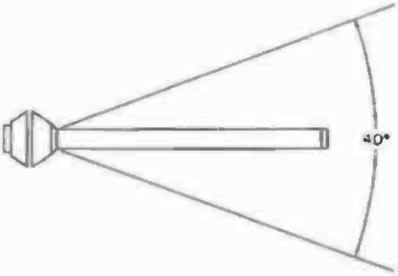


Fig. 4-101D. Included angles of super in-line microphone.

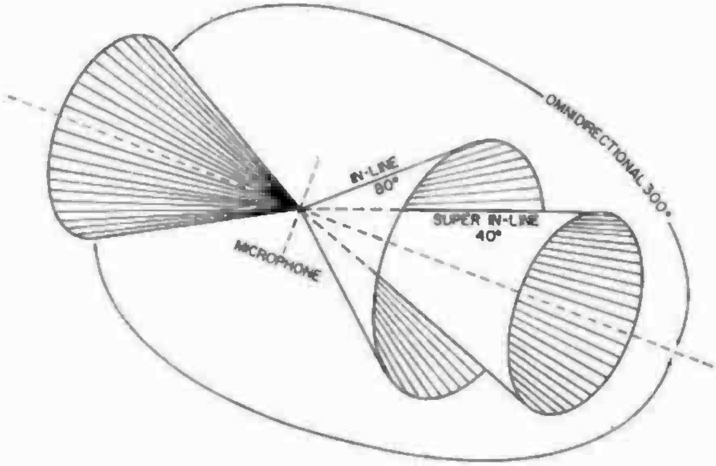


Fig. 4-101E. Included angles in the form of cones showing the range of pickup for an omnidirectional, in line, and super in-line microphone.

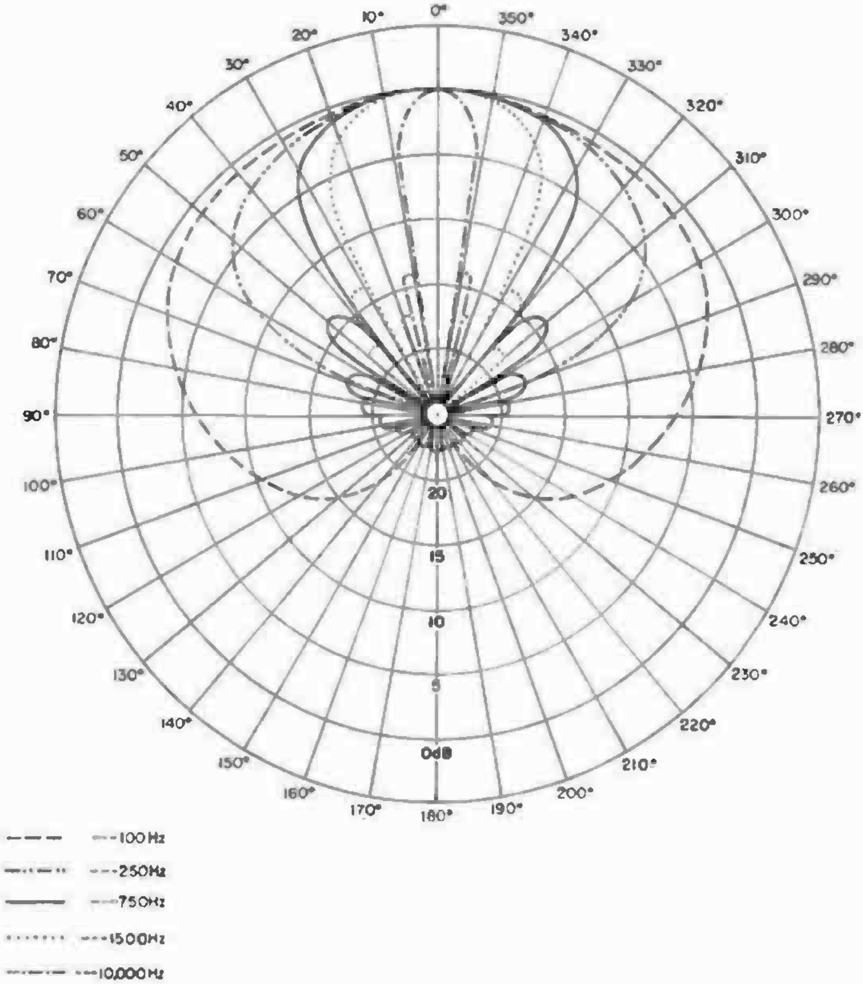


Fig. 4-101F. Polar pattern for Electro-Voice Model 643 Cardiline super-directional microphone.

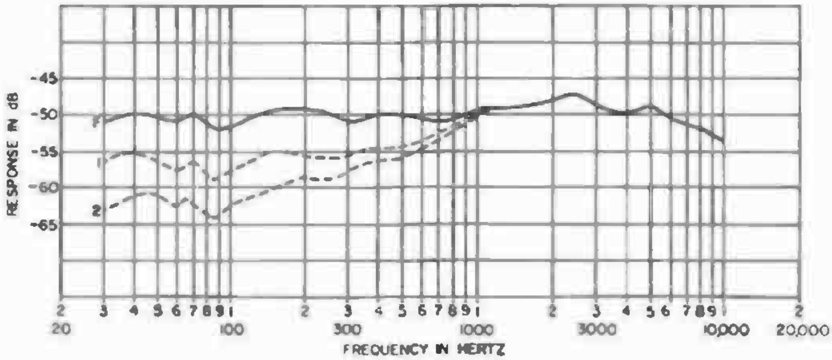


Fig. 4-101G. Frequency response for Electro-Voice Model 643 Cardiline super-directional microphone.

on axis for these microphones is shown in Fig. 4-101E.

The characteristics shown are the ideal and will not always be obtained under practical operating conditions, because of reflections from nearby walls or other objects on axis. It might be mentioned, at the present time there is no known microphone that will exclude completely all sounds originating at the sides or back. The characteristics for the Model 643 provide a cardioid polar pattern up to 100 Hz, and then a highly directional one, as the polar plot of Fig. 4-101F indicates.

The base of the microphone includes an integral two-position, low-frequency rolloff, and a high-pass filter (Figs. 4-101G and 4-101H.) The high-pass filter suppresses room reverberation, while maintaining the presence for dialogue. For comparison with the Model 642 microphone, the overall mechanical dimensions for the Model 643 are given in Fig. 4-101I. Except for the high-pass filter and the slightly different frequency response, the transducer unit, wind screen, output impedances, and sensitivity are similar to the Model 642. This microphone has been used quite successfully by both the television and motion picture industry.

Pictured in Fig. 4-101J is another in-line type microphone with extremely

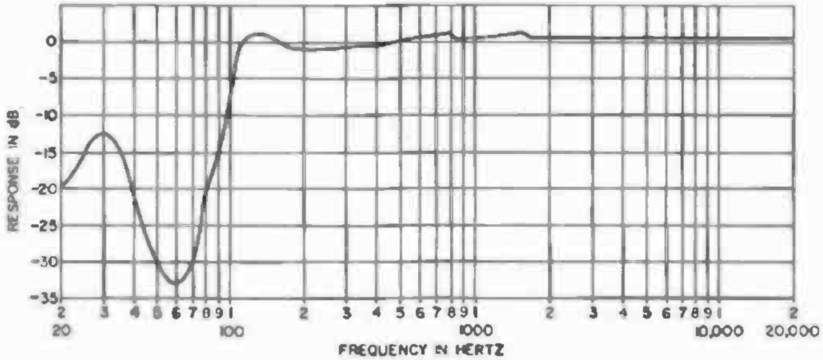


Fig. 4-101H. Frequency response of internal 100-Hz high-pass filter.

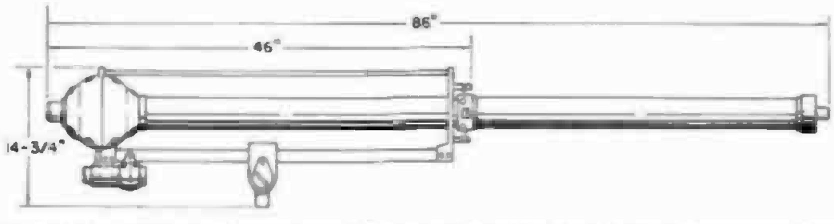


Fig. 4-101I. Cross-sectional view of Electro-Voice Model 643 Cardiline super-directional microphone, which is in reality a 7-foot version of the Model 642.

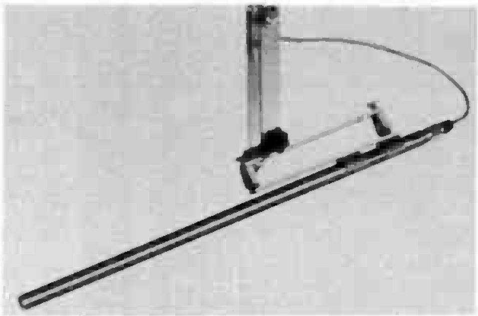


Fig. 4-101J. Sennheiser Electronic, Model MKH-805 capacitor in-line long-range pickup microphone.

narrow pickup characteristics, manufactured by Sennheiser Electronic of West Germany. This microphone uses a radio-frequency oscillator, and a two-stage transistor amplifier mounted in the end of the assembly. The basic principles of operation are discussed in Question 4.71. Insofar as the in-line tube design is concerned, it is quite similar to that described in Question 4.97.

The internal output impedance is 10 ohms and is designed to operate into a load impedance of 150 ohms. Power for the preamplifier is duplexed over the audio lines. Power requirements are 10 volts at 5 milliamperes. Frequency response is 50 to 20,000 Hz. The output level is minus 21.5 dB re: 1 mw/1 dyne/cm², with an EIA rating of minus 115 dB. This microphone like the one previously described, is suitable for long range pickup. (See Question 4.104.)

4.102 What precautions should be taken when using the microphone discussed in Question 4.101?—This particular microphone has an included angle of 40 degrees on axis and appears as a cone, as shown in Fig. 4-101E, which is the starting point for cancellation, but it should not be interpreted that no sound will be picked outside this cone. As the microphone is rotated from an on-axis position to a 180-degree off-axis position, there will be a progressive drop in level. Sounds originating at angles of 90 to 180 degrees off-axis will cancel by 20 dB or more, as indicated by the polar plot of 4-101F. However, the amount of cancellation depends on the level and distance of the microphone from the sound source. As an example, if an on-axis sound originated at a distance of 20 feet, a 90- to 180-degree off-axis sound occurring at the same distance and intensity will be reduced by 20 dB or more, providing none of the off-axis sound is reflected into the front of the microphone by walls, ceilings, etc. On the other hand, should the off-axis sound originate at a distance of 2 feet at the same level, it will be reproduced at the same level as the on-axis sound. The reason for this behavior is that the microphone is still cancelling the unwanted sound as much as 20 dB, but due to the difference in the distances of the two sounds, the off-axis sound is 20 dB louder than the axial sound and they are reproduced at the same level. For a pickup in an area where random noise and reverberation is a problem, the mi-

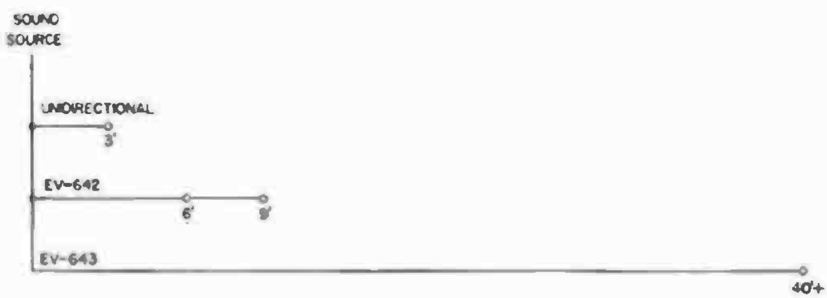


Fig. 4-102. Comparison of standard unidirectional microphone and Electro-Voice Models 642 Cardiline, and 643 long-range pickup.

crophone should be located with the *back end* to the source of unwanted sound and away from the disturbance at the greatest possible distance.

If the microphone is being used inside a truck and pointing out a rear door, poor pickup may be experienced for the following reason. All sounds, both wanted and unwanted arrive at the microphone on-axis. Since the only entrance is through the truck door, no cancellation occurs because the truck walls inhibit the sound from entering the sides of the microphone. In this instance, the microphone will be operating as an omnidirectional microphone. Due to the reflected sound from the walls the same condition will prevail in a room where the microphone is pointed through a window, or when operating in a long hallway. For good pickup, *the microphone should be operated in the open and not in closely confined quarters.*

Because of the narrow included angle of the pickup, random noise is reduced considerably, and the distance to the sound source may be increased without a loss of presence. *It should be understood that this microphone cannot be compared to a zoom lens; the focus does not vary nor does it reach out to gather in the sound.* What the narrow polar pattern and high rate of cancellation does is to reduce the random sound energy and to permit the raising of the amplifier gain *following* the microphone, without seriously decreasing the signal-to-noise ratio. This permits the pickup of voices at a distance that cannot be understood to be reproduced with intelligibility.

Difficulties may also be encountered using this microphone on stage and picking out a speaker in the audience, particularly where the voice is 75 to 100 feet distant and fed back through a reinforcement system for the audience to hear. Under these circumstances, only about 30 to 50 feet is possible without acoustic feedback; even then the system must be balanced very carefully.

If the pickup is being made in a wind of 20 miles per hour or more, the 100-Hz high-pass filter in conjunction with the *Acoustifoam* wind screen will eliminate the low-frequency distortion caused by the wind. Switching in one of the low-frequency rolloffs, as shown in Fig. 4-101G, may also be helpful. If the pickup is only dialogue, the 100-Hz filter may be left in, and the number 1 rolloff used, as most male voices may be attenuated below 120 Hz. (See Question 18.81.) The filter and rolloff characteristics are of exceptional help when reverberation is a problem in a hard-walled room.

A comparison of a standard cardioid microphone with the Electro-Voice 642 and 643 relative to increased distance pickup, is shown in Fig. 4-102. Normally pickups at a distance of 50 feet or greater, using the above precautions, are possible; however, quite satisfactory pickups have been made at distances of 150 feet. Only through *experience* will its full capabilities be realized.

4.103 What is a rifle microphone? —The rifle microphone consists of a microphone transducer with a series of tubes of varied length mounted in front of the transducer diaphragm (Fig. 4-103A). The transducer may be either a capacitor or dynamic type. The tubes are cut in lengths from 2 to 60 inches, and bound together. The bundling of the tubes in front of the transducer diaphragm creates a distributed sound entrance and the omnidirectional transducer becomes highly directional.

Sound originating on the axis of the tubes first enters the longest tube and as the wave front advances, it enters successively shorter tubes in normal progression until the diaphragm is reached. Sounds reaching the diaphragm from the source travel the same distance, regardless of the tube entered; thus, all sounds arriving on-axis are in phase when they reach the diaphragm. However, sounds originating 90 degrees off-axis enter all tubes simultaneously. A sound entering a longer tube may travel 18 inches to reach the diaphragm, while the same sound traveling through the shortest tube will travel only 3 inches, with other differences for the varied length of tubing, thus causing an out of phase signal at the diaphragm. Under these conditions, a large portion of the sound originatiug at 90 degrees is canceled; from 180 degrees an even greater phase difference occurs, and cancellation is increased considerably.

The RCA MI-10006A *Vari-directional* microphone (Fig. 4-103A) consists of nineteen $\frac{5}{16}$-inch plastic tubes, ranging from 3 inches to 18 inches in length. Each tube is damped to prevent reso-

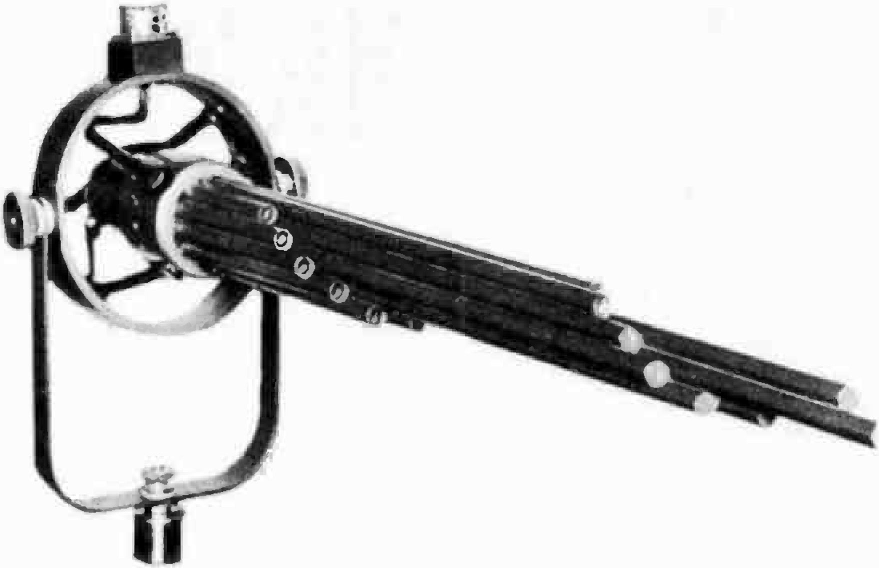


Fig. 4-103A. RCA MI-10006A Vari-directional microphone.

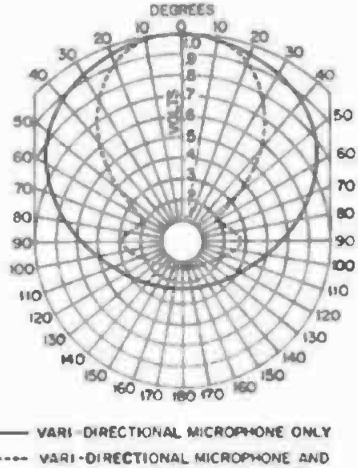


Fig. 4-103B. Polar plot for RCA MI-10006A Vari-directional microphone.

nant effects. The tubes are bundled and mounted in front of an omnidirectional capacitor-microphone head. An amplifier contained in the transducer housing is supplied from a small battery pack, having a life of several months. The polar and frequency response are shown in Figs. 4-103B and 4-103C. Two positions for reducing low-frequency response are provided at the transducer and are indicated by the dotted lines in Fig. 4-103C.

4.104 How is the effective output level of a microphone rate?—The EIA standard defines the system rating (G_m) as the ratio in decibels relative to 0.001 watt per 0.0002 dyne per square centimeter of the maximum electrical output from the microphone to the square of the undisturbed sound field pressure in a plane progressive wave at the microphone. Expressed mathematically:

$$G_m = (20\text{Log}_{10}\frac{E_o}{P} - 10\text{Log}_{10} Z_o) - 50 \text{ dB}$$

where,

E_o is the open circuit voltage of the microphone,

P is the undisturbed sound field pressure,

and Z_o is the microphone rated output impedance.

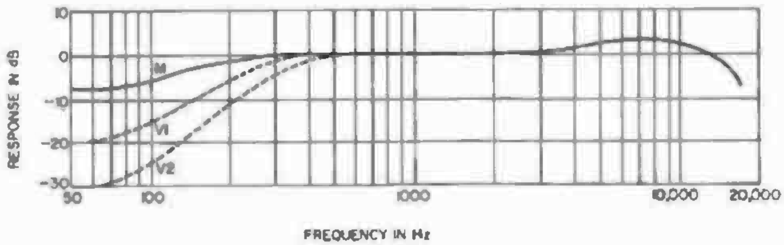


Fig. 4-103C. Frequency response for RCA MI-10006A Vari-directional microphone.

For all practical purposes, the output level of the microphone can be obtained by adding to G_m the sound pressure level, relative to 0.0002 dyne per square centimeter.

The sound pressure level (SPL) is measured with a sound level meter (see Question 22.94). The sensitivity to extraneous magnetic fields is referred to 1 milliwatt and calculated in the same manner as for the effective output level, using as the output voltage, the voltage produced by an arbitrary 60-Hz magnetic field of 0.001 gauss. This subject is discussed in detail in Question 23.81.

4.105 Show the relationship between air-particle amplitude, frequency, and sound pressure in microbars.—By referring to the graph in Fig. 4-105, the air-particle amplitude for frequencies ranging from 10 to 10,000 Hz, may be read directly in microinches for sound pressures between 0.0001 microbar and 100 microbars. As an example: What is the air-particle amplitude for 1000 Hz at an SPL of 0.01 microbar? Enter the graph at 1000 Hz at the left and follow the horizontal line to where it intersects the diagonal line for 0.01 microbar. Follow the intersecting vertical line downward to the lower margin, and read 0.015 microinch. For 10,000 Hz, the amplitude is 0.0015 microinch. Dropping down 10 Hz, the amplitude increases to 1.15 microinches. Taking the amplitude of 10,000 Hz at 0.01 microbar and 100 Hz at 100 microbars (which equals 1500 microinches) the ratio is 1,000,000:1.

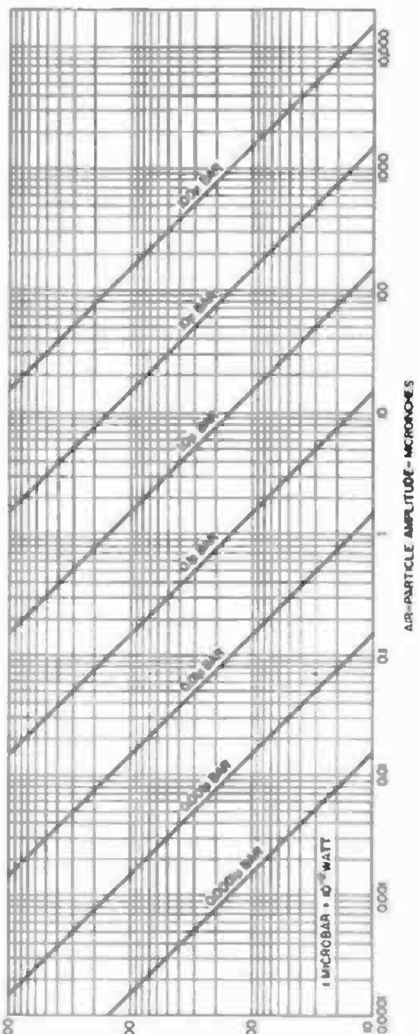


Fig. 4-105. Amplitude of vibrating air-particle versus frequency.

4.106 What is the average output level of different type microphones, taking into consideration the basic design?—The majority of microphones manufactured in the United States rate their microphone level in accordance with the EIA formula, discussed in Question 4.104. However, some European manufacturers rate their output level in volts for a given impedance. Using the EIA standard output rating, the average output levels for several different type microphones are:

Carbon-button (communication type)	−60 to −50 dB, −20 to −10 dB
Crystal	−50 to −41 dB
Ceramic	−50 to −40 dB
Dynamic (moving coil)	−53 to −48 dB
Capacitor (internal amplifier)	−51 to −37 dB
Ribbon-velocity	−51 to −48 dB (up to 0 dBm)
Lavaliers (dynamic moving-coil)	−67 to −60 dB
Transistor (communications)	−60 to −53 dB
Sound power (moving-coil)	−32 to −20 dB

Microphones are also rated by stating the output impedance and output level relative to 1 milliwatt/10 dynes/cm².

4.107 Show a nomograph for determining the output level of a microphone, both open-circuit and terminated.—For low-impedance type microphones (250 ohms or less) the output level is given

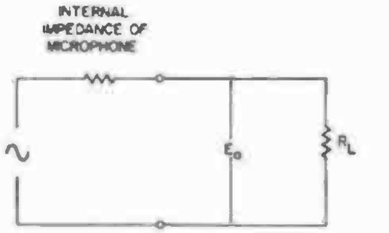


Fig. 4-107A. Equivalent circuit for a microphone output circuit.

in dB, with reference to 1 milliwatt/10 dynes/cm² sound pressure. For carbon and high-impedance microphones (above 250 ohms) the output level is referenced in dB to 1 volt, for 1 dyne/cm² sound pressure. It will be noted that both the acoustical pressure at the face of the microphone and the electrical output are given. Therefore, the microphone may be considered to be an acoustical generator, with a voltage or power output.

The specified output level is taken at 250 Hz for high quality wide-range professional microphones, while for communication types which have considerably less bandwidth, it is taken at 1000 Hz. For low-impedance microphones, the power output is specified. Since the output impedance of these microphones is low, it is customary to transform them to a higher impedance for application to the input circuit of an amplifier. The output is given in power, which does not change with the transformation, except for a small insertion loss of the transformer, which is generally less than 1 dB. As an example: For a low-impedance microphone with a sensitivity rating of minus 60 dB, re: 1 milliwatt/10 dynes/cm², what is the input voltage at the grid of the amplifier? Visualizing the microphone as a generator (Fig. 4-107A), the sensitivity rating of minus 60 dB means that for an acoustic pressure of 10 dynes/cm² at the diaphragm, the microphone will deliver into a resistive load (equal to its own impedance) a power of 10^{-9} watt. Therefore:

$$E_o^2 = 10^{-9} \quad E_o = 3.18 \times 10^{-4} \times R_L$$

For an impedance at the grid of the first stage of 40,000 ohms, the output voltage will be:

$$E_o = 3.18 \times 10^{-4} \times 40{,}000 \text{ ohms}$$
$$= 0.0637 \text{ volts}$$
(for 10 dynes/cm² sound pressure).

If the microphone is operated open-circuit, the output voltage will be doubled. A nomograph for determining the output voltage for various sensitivities is given in Fig. 4-107B. (See Question 12.170.)

High-impedance microphones are not transformed, as they are already 20,000 to 40,000-ohms impedance (internal transformer), and may be directly connected to the input of an amplifier, and

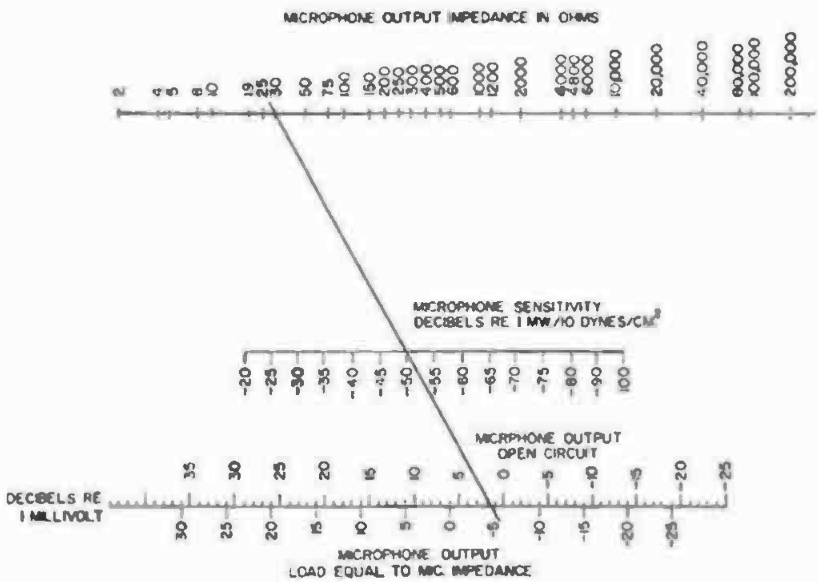


Fig. 4-107B. Nomograph for determining the output level of a microphone in decibels or voltage, both open-circuit and terminated.

the sensitivity is given as an open-circuit voltage, for a given sound pressure. A high-impedance microphone, with a sensitivity of minus 55 dB re: 1 volt/1 dyne/cm² will produce an open-circuit voltage of 0.0018 volt, with a sound pressure of 1 dyne/cm². The sensitivity rating determines the output voltage at only one sound pressure. However, since nearly all microphones are linear over all pressures encountered, the voltage output is a direct function of the sound pressure. If the pressure is increased by a factor of 10, the electrical output will increase by a factor of 10. Thus, knowing the sound pressure and sensitivity, the output voltage may be determined.

Sound pressure is generally given in reference to 0.0002 dyne/cm², and is considered to be zero sound pressure, and is approximately equal to the threshold of hearing. Typical sound pressures are:

Speech, ¼-inch from microphone front, 107 dB (peaks may add another 12 dB)

15-piece orchestra, 72 dB (with another 20 dB on peaks)

Threshold of pain, 130 dB

Acoustic noise in very quiet studio, 10 dB

Thus it can be seen that the first stage of a microphone preamplifier is required to operate over a wide range of input voltage.

To use the nomograph, two factors must be known in advance—the output relative to 1 milliwatt/10 dynes/cm², and the impedance. For example, to find the open-circuit output level for a microphone of 30-ohms impedance, rated minus 50 dB (RCA 77-DX, Question 4.77), a straight edge is laid from the 30-ohm impedance on the upper scale to a minus 50 dB on the center scale. The output level is read on the open-circuit values on the lower scale, which for this example is 1.5 dB above 1 millivolt. Terminating the microphone in its own impedance, the output voltage level drops to 4.5 dB below a reference level of 1 millivolt.

4.108 Give the equation for calculating the ribbon-deflection amplitude in a ribbon-velocity microphone.—Of the various types of microphones developed through the years, the ribbon-velocity microphone has, by far, the greatest deflection amplitude, approaching within a few percent the amplitude of the vibrating air particles. The amplitude of the ribbon deflection may be calculated:

$$d = \frac{e10^{8}}{LB2\pi f}$$

where,

d is the deflection amplitude,
e is the open circuit voltage,
L is the length of the ribbon in centimeters,
B is the flux density in the air-gap in gauss,
and f the frequency in Hz.

As an example, if the ribbon length is 5 centimeters, the flux density 10,000 gauss, the frequency 1000 Hz and the open circuit voltage 0.001 volt, for a sound pressure of 1 microbar, the ribbon-deflection amplitude is 1.2-microinches. As the frequency decreases, the ribbon deflection increases. For an *increase* in frequency, the ribbon deflection decreases.

4.109 What does the term dBV mean?—This term is used with a microphone when the output level is rated with reference to a one-volt reference level. (See Question 10.36.)

4.110 How is the transformer in a microphone designed to prevent pickup from extraneous fields?—The coils are wound hum-bucking and the whole assembly is encased in a triple-shielded case as shown in Fig. 8-50. The efficiency of transformer shields may be measured as described in Question 23.162.

4.111 How are unidirectional ribbon-velocity microphones operated from a boom? — For an example, the RCA MI-10001C microphone (Question 4.59) is suspended from a turret head on the boom, tilted at an angle of 45 degrees relative to the floor, about 4 feet from the sound source. For dialogue, the voice filter is used or dialogue equalization in the mixing panel (one or the other, but not both). For a normal pickup at this angle, the response changes less than 1.5 dB at any frequency. The ribbon should never be placed parallel to the floor, because of the directional nature of the human voice at the higher frequencies. (See Question 4.114.)

4.112 How should microphones be placed on a stage for the best results?—Several microphones spaced about 10 feet apart should be placed in the footlight trough, in an enclosure made of wire screen and covered on the outside with 2 inches of rock wool and a thin layer of felt on the inside surface (Fig. 4-112). This enclosure is designed to reduce or eliminate pickup from the orchestra pit. Unidirectional microphones with rather broad polar patterns should be used; they can be capacitor, dynamic, or ribbon-velocity types. Microphones should also be hung from overhead when possible, in areas where the players are likely to move, to assure an even sound pickup. This is particularly true if the actors turn their backs to the audience.

For small band combinations with a vocalist, it may be necessary to hang a loudspeaker 10 to 15 feet above and slightly behind the performers, and play back the program material at a level that cannot be heard by the audience or fed back to the microphones. This gives the musicians and the vocalist a feeling of the reproduction in the auditorium, and reduces the dead feeling due to the acoustic damping of the stage equipment.

Each microphone output is brought to a separate mixer control at the console. It is of extreme importance that all microphones be in electrical phase with each other, or dead spots may occur between microphones, as the actors walk in front of the microphones. Phasing is discussed in Question 4.85. For the most satisfactory sound pickup, the operator should attempt to follow the action, using only the necessary microphones. Separate microphones are used for the orchestra pickup.

If loudspeakers are used in the auditorium for sound reinforcement, the reproduced level should be just loud enough to get a good coverage, and of such a level that the audience is not aware that reinforcement is being used.

4.113 How is a sound pickup made from an outdoor band shell?—Because of its design, the shell acts as a huge concentrator making it difficult to obtain a pickup with good orchestra balance. This undesirable effect may be overcome by the use of polycylindrical diffusers, as explained in Question 2.76. The use of polycylindrical diffusers will also permit a better balance between the instruments and eliminate resonant peaks that go with a band shell.

4.114 What are the general rules pertaining to the placement of microphones for recording dialogue, music and sound effects, and lecterns?—A single microphone pickup should be used whenever possible. When two or more microphones are used simultaneously for recording either dialogue or music, they should be placed at least 10 feet apart. Also, their levels should not differ by more than 2 dB, have similar frequency characteristics, and be in phase electrically if the final result is to be uniform. An exception to the foregoing occurs when solo microphones are used.

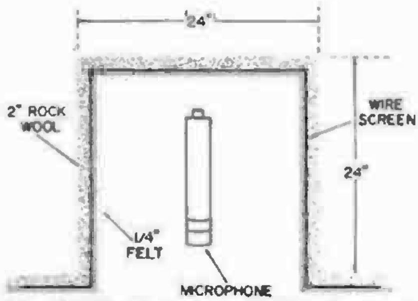


Fig. 4-112. Plan view of microphone enclosure for use in a footlight trough.

When multiple microphones are placed relatively close to the source of the pickup for either dialogue or music, and the sound source is directional, it becomes more important than ever to observe the distance the microphones are separated, their signal levels to one another, and the phasing.

Microphones should not be placed near highly reflective surfaces and must be suspended by resilient supports from either a boom or stand. For television and motion picture recording, the distance the microphone is placed from the source of sound will depend on the scene being photographed. Obviously, long-shot quality with reverberation cannot be used with a close-up and vice versa, as neither would sound natural.

The boom man should not be confused about the position of the camera or its distance from the subject as close-ups and medium-long shots are often made without moving the camera, only the lenses are changed. This is particularly true if a zoom lens is employed.

For a medium shot, the microphone is moved a little farther away from the subject than for a close-up. For a long shot the microphone is left at a distance comparable with the picture composition. However, this is not always done when making motion pictures for television viewing exclusively, as it is important that the dialogue have a high rate of intelligibility.

It is the practice to employ a fixed amount of midrange high-frequency equalization and rolloff at the low frequencies in the production mixer panel. The midrange high-frequency equalization adds presence to the voice, and the low-frequency rolloff attenuates the tubbiness in male voices and reduces low-frequency noise picked up on the set. The same type microphone should be used throughout a production, if possible; if it is not, a noticeable change of quality will be apparent and difficult to correct later in rerecording. After the correct equalization has been selected for a given type microphone, it should not be changed. If different type microphones are employed, *their frequency response must be matched to that of the original microphone*. Any further corrections as to presence and low-frequency attenuation should be left to rerecording. Twelve dB of low-frequency attenuation is maximum, but less than 6 dB (100 Hz) should not be used.

A microphone suspended from a boom should not be panned with too great a rapidity if it can be avoided, as noises may be generated by the movement of the boom mechanism and the swishing of the air as it passes over the face of the microphone.

Recording outdoors is not too different from recording indoors. If an outdoor scene is being shot indoors, the microphone must be kept close to the source of sound to prevent picking up an undue amount of reflected sound and thus lose the outdoor effect. If the indoor scene is being shot outdoors, reverberation can be introduced later during rerecording to simulate the indoor scene.

As most recording systems provide headphones for the boom man, it is his responsibility as well as that of the mixer to maintain a uniform quality of pickup by keeping the microphone at a fixed position in front of the actors although he is required to move the microphone over a considerable area of the set. If the actors are spaced a considerable distance from each other and it is not practicable to pick them up with one microphone, two microphones are used; however, it is highly important that the microphones be phased. The phasing of microphones is discussed in Questions 4.84 and 4.85.

Unidirectional capacitor-dynamic, or ribbon-velocity microphones are used for motion picture recording and television, because of the high ratio of back-to-front pickup. The microphone is maintained within a distance of four feet forward of and not less than three feet above the person speaking. This is illustrated in Fig. 4-114A.

Microphone booms designed for motion picture recording are equipped with a rotatable turret head for gunning the microphone rapidly from one source of pickup to another. This keeps the pickup quality uniform. In some instances a tilting mechanism is also

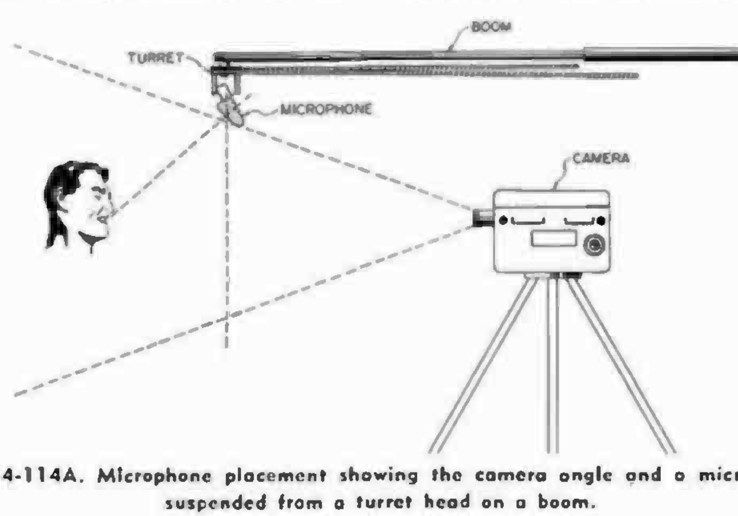


Fig. 4-114A. Microphone placement showing the camera angle and a microphone suspended from a turret head on a boom.

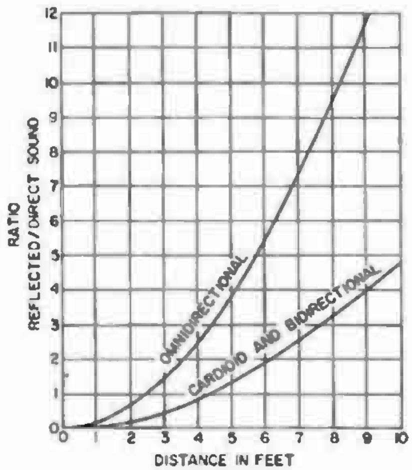


Fig. 4-114B. Ratio of reflected sound to direct sound.

mounted on the turret head to permit the microphone to be moved in a vertical direction. Rotating a unidirectional microphone such as the RCA MI-10001 (Fig. 4-59A), only varies the sensitivity about 1.5 dB when rotated 45 degrees either side of the center plane of the ribbon. The closer these type microphones are placed to the subject, the greater will be the increase of low-frequency response. (See Question 4.127.)

For dialogue recording, low-frequency attenuation is introduced starting at 800 Hz and slowly tapering off to a point where 100 Hz is down 8 to 12 dB with respect to 1000 Hz. This characteristic is discussed in Question 18.81. Dialogue should never be recorded flat because of the increase of tubbiness of the human voice due to set conditions and also to microphone characteristics. If more than one microphone is being used, the voice filters or equalizers must be set to achieve as near similar frequency-response characteristics as possible. If dialogue equalization is used in the mixer, the microphone is set for a flat frequency response. *Dialogue equalization and the voice-filter are never used in combination.*

If a bidirectional ribbon-velocity microphone (Fig. 4-50) is used, it must be kept from 4 to 5 feet away from the subject to prevent a tubby response. If the microphone is rotated, it must be done in such a manner that the dead side of the field pattern (Fig. 4-7C) does not face the sound source. Panning

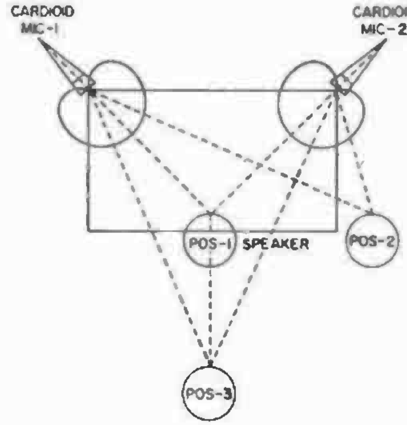


Fig. 4-114C. Usual microphone placement on lectern.

must be smooth without overshooting the subject. Although this microphone is satisfactory for fixed shots, it is not recommended for boom operation.

Microphones are expected to operate under adverse conditions. To facilitate their operation and increase their usefulness, several devices have been devised. Among these are the wind screen and the rain screen. In the early days of sound recording, several layers of thin silk were wrapped around the outside of the microphone to reduce the effects of wind. The many layers of silk also served to reduce the high-frequency response and affected the quality of the recording. Later, manufacturers of microphones developed special wind screens that could be easily attached to the microphones when required. The design of these screens is such that wind noise is reduced without affecting the sound quality. A typical wind screen is shown on the microphone in Fig. 4-31.

Special effects such as creating the illusion of a large auditorium may be obtained by placing the microphone in such a relationship to the sound source that a greater part of the pickup is reflected rather than direct sound. If this is not satisfactory, the sound is reverberated by the use of an echo chamber or reverberation unit, described in Question 2.80. The ratio of direct sound to reflected sound for an omnidirectional microphone, compared to a cardioid and a bidirectional ribbon-velocity microphone is given in Fig. 4-114C.

Microphones used for production recording should have frequent checks of their sensitivity, frequency response, phasing, and overall sound quality to determine whether any internal noise is being generated. These tests may be made using the techniques described in Question 23.80.

Microphone placements for large orchestra setups are discussed in Questions 2.118, and 2.120. Although there are some basic rules for such pickups, each mixer has his own preference as to microphone placement, because he must consider the characteristics of the stage, the number of pieces in the orchestra, and the type of music being recorded, also if it is monophonic or stereophonic. All of these factors affect the placement of the microphones and only actual listening tests will determine the best locations.

In some instances where it is not practical to use a microphone on a boom, a wireless microphone, as described in Question 4.72, may be used. Microphones may also be concealed on the set; however, they are seldom in the correct relationship to the source of sound. If the shot is a fixed one and the characters do not move out of the field of the microphone, it may be hung permanently overhead or concealed on a table or in some other object near the source of the pickup. (See Questions 6.122 and 18.81.) For long range pickups, such as race tracks, parades, groups of people in a crowd, etc., a long range microphone or sound concentrator may be employed, as described in Questions 4.100 and 4.101.

Placement of microphones may appear at first glance to be quite simple; however if certain precautions are not observed, a radical change in the quality of the pickup can occur quite easily. The usual manner of setting up microphones on a lectern is shown in Fig. 4-114B. If the speaker stands at position 1 and does not move, the pickup could be satisfactory. However, speakers do not always stand in one place, and it is not uncommon for them to step completely out of the area of pickup. At position 2, the sound originates off axis from microphone 2, but is fairly on axis of microphone 1. Any movement by the speaker in this area is manifest by a considerable change in the quality of the sound, because of phase differences between the two microphones due to spatial relationship to the sound source. Major dips in the frequency range of 800 Hz and higher will be quite noticeable, resulting in a lack of presence. If the placement of the microphones is changed to where the microphones are rotated 15 to 25 degrees from center (included angle of 30 to 50 degrees) with about 12 to 20 inches between, better results may be expected. Greater angles will permit the speaker more latitude of movement, and the smaller distances between the microphones reduces both the acoustic feedback and background noise. One of the most important points to remember is the *microphones must be in phase electrically* as discussed in Question 4.84. Also, the output levels should be within plus or minus 2 dB, with similar frequency characteristics.

Although omnidirectional or bidirectional microphones can be used for a single microphone pickup, they are generally unsatisfactory because noise picked up from the audience side is mixed with the speaker's voice and when sound reinforcement is in use, serious acoustic feedback can occur. If an omnidirectional microphone is used, the speaker must work as close to the microphone as possible, and the gain of the system must be continuously monitored to prevent feedback. For a single microphone pickup, it is much more satisfactory to use a lavalier type microphone, as discussed in Question 4.81. If a bidirectional microphone is used, (figure-8 polar pattern) the waist of the polar pattern is turned toward the audience at an angle of about 30 to 45 degrees, to reduce the pickup from that direction. Microphone placement for several different type pickups is discussed in Questions 2.124 and 2.127.

Recording of sound effects is rather simple, as only two rules are important. The first is to record as much level on the sound track as practicable, keeping in mind the scene in which it is to be used. The second rule is to keep the background noise to an *absolute minimum*. Equalization should be adjusted to bring out any special effects at the low- and high-frequency end of the spectrum. This will result in good solid sound tracks that can be again equalized during rerecording to suit the scene. Many times a good sound effect is unusable because the level is too low, and in trying to bring it up to a useable level, the background noise becomes excessive. Therefore, the same care must be exercised for recording sound effects as when recording dialogue.

4.115 Describe the construction of a high-intensity capacitor microphone head.—High-intensity capacitor microphone heads are especially designed for measuring high-level sound pressures of jet engines, explosions, acoustic wave shocks, and atomic blasts. A typical example of such design is the Altec-Lansing type 21-BR, shown in Fig. 4-115A. The general theory of its operation is the same as that explained in Question 4.41. Except for the stiffness of the diaphragm, and the internal capacitance being less, the heads follow the same general design.

In the Altec-Lansing capacitor-microphone preamplifier the conventional grid resistance is omitted. Since the microphone head is capacitive, extended low-frequency response dictates that the shunt resistance of its load must be as high as possible. The source impedance viewed from the preamplifier control grid consists of resistance and capacitance in parallel. Also, over the useful frequency range the thermal noise is least when the resistance is the highest. For these reasons no physical grid resistor is employed, the effective resistance between the control grid and cathode being that established by the input of the tube itself. Thus, the load resistance is extremely high and the self-established bias is steadily maintained at approximately 1 volt negative. Consequently, the polarizing voltage applied between the electrodes of the microphone head is approximately 1 volt less than the dc voltage drop across the cathode resistance. The voltages required for operating the amplifier are supplied from an external power supply.

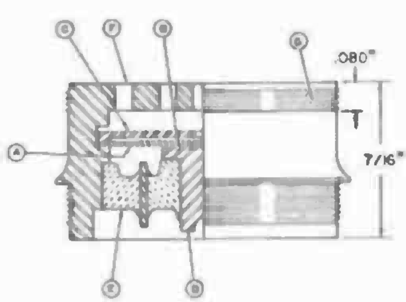


Fig. 4-115A. Cross-sectional view of Altec-Lansing 21-BR high-intensity capacitor microphone head.

The cross-sectional view in Fig. 4-115A is of a typical high-intensity microphone head. Such heads may be obtained for measuring SPL's up to 200 dB, over a frequency range of 5 to 15,000 Hz. The selection of the proper head depends on the device to be measured, its frequency range, and the maximum SPL expected. The diaphragm (A) is 0.5 inch in diameter, made of glass ground optically flat to the thickness of 0.002 to 0.013 inch, depending on the required sensitivity. One surface (B) is gold plated, which makes contact with the outer shell. A damping plate (C) is placed over the diaphragm. The fixed electrode (D) is ungrounded and is a machined part extending through a Mycalex dielectric (E). The dielectric is provided with annular corrugations between the fixed electrode and the outer sleeve to extend the length of the surface leakage paths. To maintain the highest surface resis-

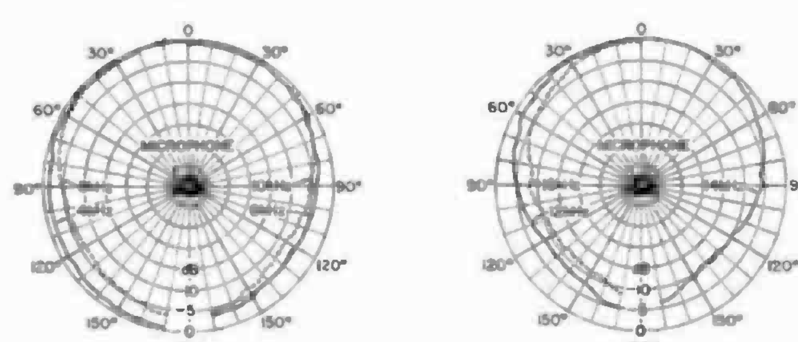


Fig. 4-115B. Polar diagrams for Altec-Lansing 21-BR high-intensity capacitor microphone.

tivity and prevent moisture-film globules, the surface of the dielectric is treated with a silicone compound. The critical elements are made of either glass compounds or stainless steel which have nearly identical coefficients of thermal expansion, to assure the maintenance of the close spacing.

Holes (F) at the top admit the sound pressure to the interior. Threads (G) are for screwing on a protective cap when the microphone is not in use.

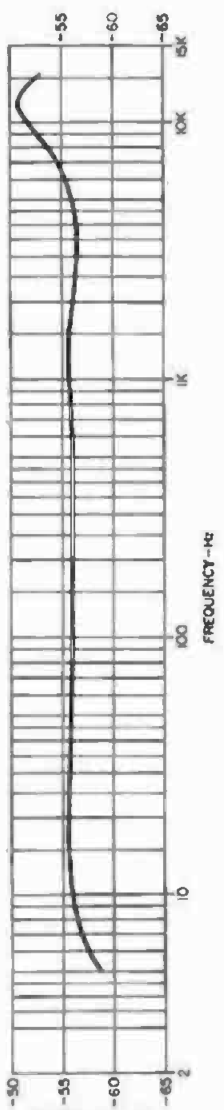


Fig. 4-115C. Typical free-field calibration for Altec-Lansing 21-BR-150 high-intensity microphone, parallel incidence. The output level is the open circuit voltage in dB re: 1 volt/dyne cm².

The axial acceleration of the mass of the diaphragm causes it to deflect to an extent proportional to its density, and inversely proportional to the square of the thickness. Since the diameter of the 21-BR microphone head is $^{11}/_{16}$ inch, or 1 wavelength at 17,000 Hz, variation with the angle of incidence is limited to the highest frequency, as may be seen by inspection of the polar diagrams in Fig. 4-115B. Microphone heads are available for measuring sound pressures ranging from 50 dB to 200 dB re: 0.0002 dyne/cm². The difference in construction is the thickness of the diaphragm. In the low-intensity types, diaphragm resonance is damped principally by means of a narrow air film between the diaphragm and the back-plate. However, in the high-intensity designs the rear diaphragm space is connected to the free air by a vent of sufficiently high impedance to support the frequency response to 5 Hz, and fast enough to prevent the development of static pressure differences when used in aircraft. If the head is exposed to moisture for any length of time, it may be dried out by placing it in a desiccator, or by heating it to 300 degrees Fahrenheit.

A typical free-field calibration for an Altec-Lansing Model 21-BR-150 head capable of linear measurement to 158 dB, made at parallel incidence, is shown in Fig. 4-115C. It will be noted the resonant frequency occurs at 11,000 Hz, with maximum peaks of 4 dB. The polar pattern is omnidirectional. The 165-A base to which the capacitor head attaches contains a preamplifier that uses a 5840 vacuum tube connected as a pentode cathode follower (Fig. 4-115D). The output from the cathode follower and the operating voltages are carried over a 6-conductor cable to the power supply. The polarizing voltage for the head is 200 volts. The cathode current for the 5840 tube is 5 milliamperes. Although the internal output impedance of the cathode follower is 640 ohms, it must not be loaded with less than 50,000 ohms, which is substantially an open circuit. Lower impedances will affect the linearity of the cathode follower. A plug-in transformer may be used for providing a number of output impedances.

Linear limit is the SPL that produces an open circuit output voltage of 20 volts; this is the level that causes the cathode follower to depart 1 dB from a linear input-output response at 10,000 Hz when the output is loaded with its regular cable. The average noise threshold is an SPL that produces a signal-to-noise ratio of unity and a noise output of 0.2 mV. The total noise is primarily the thermal noise of the input circuit of the microphone, with contributions from tube and leakage currents across insulating materials.

4.116 ***What is a filter microphone?*** — A microphone used to simulate a telephone conversation during a radio broadcast. A normal microphone is used with a telephone equalizer or filter in the circuit as described in Question 7.73.

The actors, when required to carry on a conversation over a telephone, step

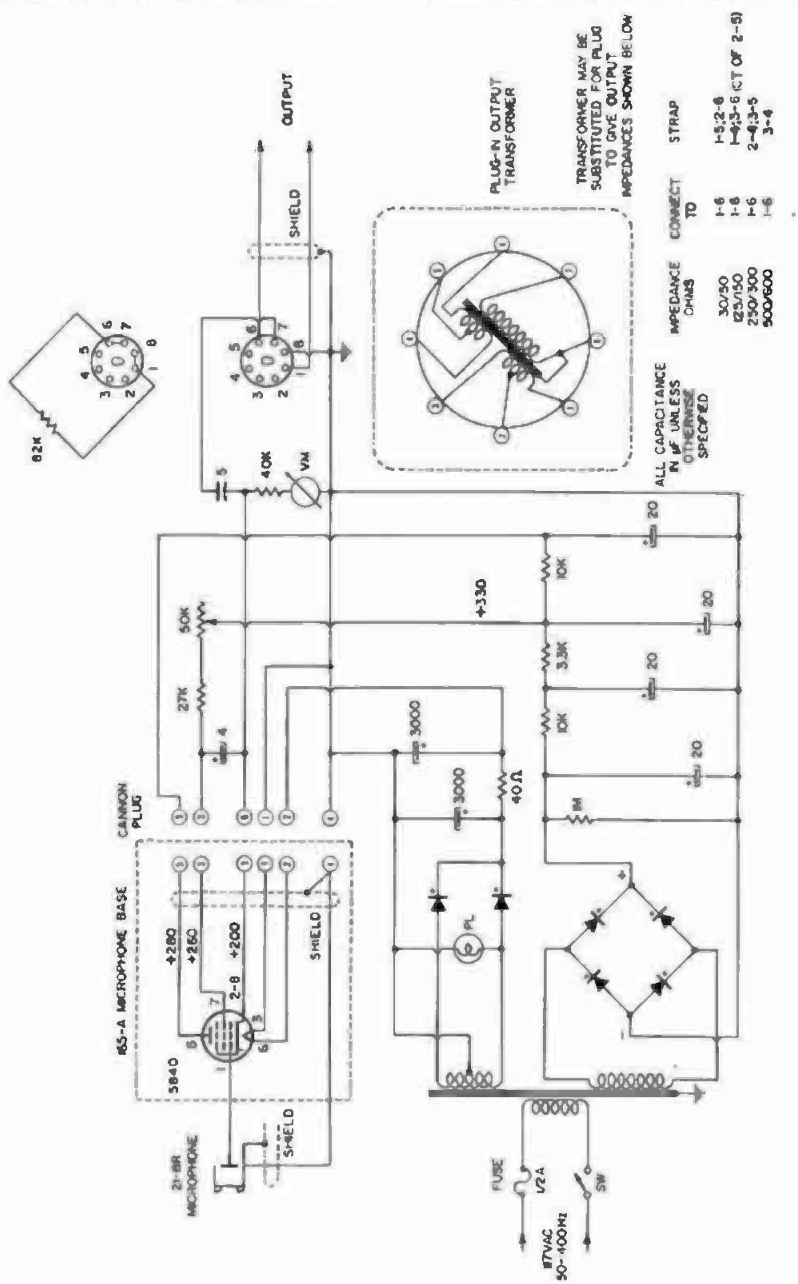


Fig. 4-115D. Schematic diagram for Altec-Lansing 21-BR capacitor microphone, 165-A base, and 526-B power supply.

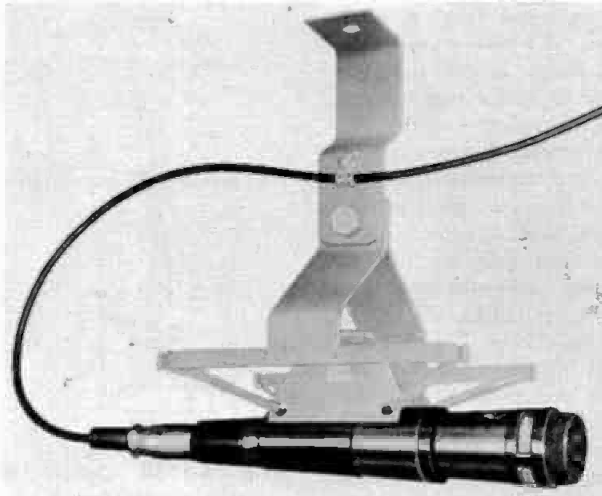


Fig. 4-117A. Altec-Lansing Model 689BX dynamic cardioid microphone and Model 181A hanger.

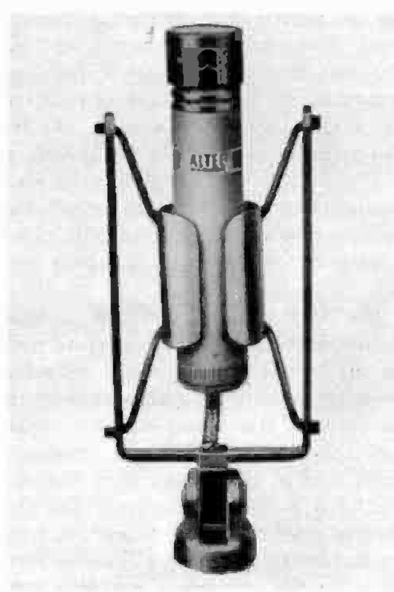


Fig. 4-117B. Altec-Lansing Model M30 capacitor cardioid microphone and Model 169A hanger.

to the filter microphone and use it, rather than having the mixer cut-in a telephone equalizer or filter each time it is required.

4.117 How are microphone hangers constructed?—Each type microphone as a rule requires an individually designed hanger because of the microphone body contour. However, all the microphone hangers have one thing in common, the microphone is supported by a metal holder which is in turn supported by a rubber mounting of some sort.

The purpose of the hanger (besides holding the microphone) is to isolate the microphone from vibrations transmitted by the microphone boom and its mechanism. The hanger should also be designed that when attached to a boom, it can be turned toward the actors for the best sound pickup. (See Question 2.101.)

The hanger used to support the RCA MI-10001C microphone from a boom is shown in Fig. 4-59A. It will be noted in this hanger two metal rings are clamped around the microphone body, and are supported by a group of heavy rubber bands fastened to a larger ring supported by a vertical bracket held by the turret head of the microphone boom. Wing nuts at the lower end of the vertical bracket permit the microphone to be tilted at various angles with respect to the sound source. Several different type hangers may be seen in Figs. 4-31, 4-39, 4-46, 4-59, 4-97, and 4-117A and 4-117B. A special hanger, including a tilting mechanism has been developed by the Fisher Boom Co. for the microphone boom pictured in Fig. 2-101A, which will permit the boom man to change the angle of incidence with respect to the persons speaking, as well as, "gunning" or "panning," from person to person.

4.118 Describe the construction of an optical microphone.—In an optical microphone, the sound waves modulate a light beam reflected from a small mirror mounted on a ribbon or a diaphragm. This design is one that has been under discussion for many years, and to date no microphones of this design have been produced commercially. It has been proven mathematically that it is not feasible to build such a microphone, although several patents have been issued for such a design.

The first design to be discussed was patented in 1938, by Banks, US Patent 2,259,511, and appears in Fig. 4-118A. The assembly is housed in a metal container (A) with an opening on one side. A thin metal diaphragm (B) is mounted below this opening, and the front face is exposed to incident sound, while the rear face is highly polished to function as an optical mirror. The diaphragm is suspended a short distance in front of a fixed reflector (C), with the two mirror surfaces parallel to each other. A suitable light source and slit (D) and lens (E) project a beam of light through the aperture (F) at one end of the second reflector (C) onto the rear reflector side of the diaphragm, at an angle of 45 degrees which, in turn, is rereflected back onto the second diaphragm, and so on for a predetermined number of reflections, in a zigzag pattern.

The light is finally brought to focus on one edge of a second aperture near the other end of the second reflector; however, some light passes through to the cathode (G) of an electron-multiplier tube (H). In the static condition, about half of the light passes into the multiplier cathode; therefore, if the diaphragm is vibrated, the reflections from the diaphragm reflector will cause variations in the amount of incident light falling on the cathode of the electron-multiplier tube. If the diaphragm reflector moves through a distance (D),

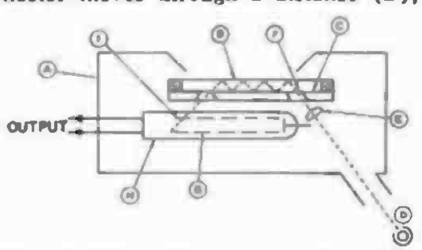


Fig. 4-118A. Optical microphone invented by Banks. A light beam is modulated by a diaphragm and projected on an electron-multiplier tube.

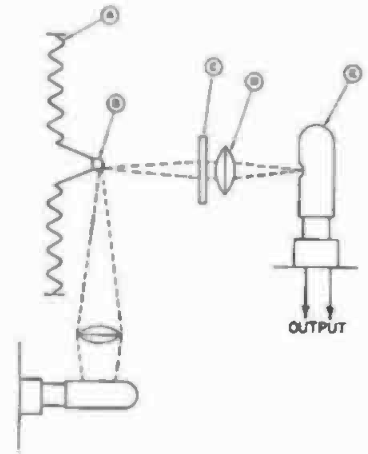


Fig. 4-118B. Optical microphone invented by MacDonell. A light beam is focused on a cylindrical mirror mounted on a diaphragm. The beam is projected through a grating to a photocell.

the lateral movement of the light image at the focus will be (nD), where n is the number of reflections of the light in its passage. The field electrode (I) opposite the cathode must be light permeable (mesh form). The generated signal is taken from the output electrode of the electron multiplier.

A second US Patent (2,666,650) issued to MacDonell is for a different design shown in Fig. 4-118B. This device consists of a diaphragm (A) with a small mirror (B) mounted at the apex of the diaphragm. A beam of light from an exciter lamp is focused on the mirror and is reflected to a grid (C) and lens (D), onto the cathode of a phototube (E). Sound waves striking the diaphragm set it in motion, causing the front-surfaced mirror (B) to vibrate. The condensed light from the exciter lamp focused on the mirror is refracted to the grid (C), where modulation occurs, then to the lens (D) and onto the cathode of the phototube. As the light beam is modulated by the mirror, it is reflected to the grid (C) at an angle which increases and decreases because of the cylindrical shape of the mirror. The grating consists of a piece of quartz or glass, divided into parallel lines equally spaced and equal in width. Each alternate space in the grid is opaque, and the intervening clear spaces are left more or less transmittive as dictated by the type phototube employed. The output signal is taken from

Fig. 4-119A. Euphonics Model C-47M semiconductor close-talking communications microphone.

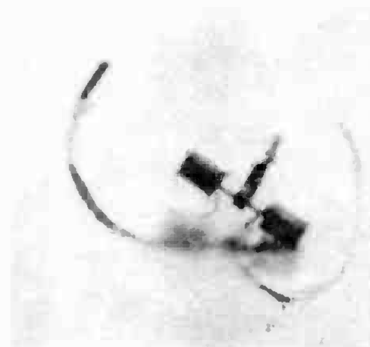


Fig. 4-119C. Semiconductor twister element, twister yoke, and supporting pads with output leads of the Euphonics Model C-47M semiconductor microphone.

the phototube and amplified in the usual manner.

4.119 Describe a semiconductor close-talking communication microphone.—The microphone to be discussed is a Model C-47M, semiconductor microphone (Fig. 4-119A), developed by John F. Wood and George Grover of the Euphonics Corp., Guaynabo, Puerto Rico. The device consists of a highly doped silicon semiconductor element, manufactured by the Endevco Corp., to which a source of dc voltage is applied. The silicon element is coupled to a diaphragm through a flexible yoke. Sound waves impinging on the surface of the diaphragm cause the coupling yoke to transmit a twisting action to the silicon element, changing its internal resistance. This modulates the dc voltage applied to the silicon element; thus, an end audio-frequency signal that is proportional to the sound wave striking the diaphragm is produced. The output signal is then amplified in the conventional manner. It should be understood that the semiconductor does *not* generate a signal, but *modulates* the applied dc voltage, thus producing the audio-frequency signal.

Referring to Fig. 4-119B, here is shown a cross-section view of the semiconductor element and its components. Element (A) is a laminated epoxy beam, with the upper surface (B) gold plated, and a brass strip (C) on the

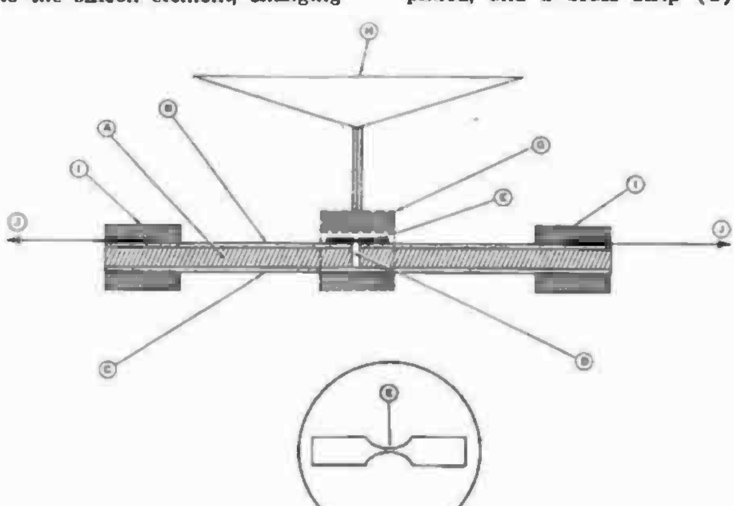


Fig. 4-119B. Cross-sectional view of Euphonics Corp. Model C-47M semiconductor communications microphone.

Fig. 4-119D. Semiconductor element assembly for Euphonics Model C-47M semiconductor close-talking communications microphone.

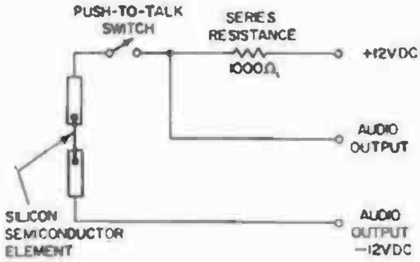


Fig. 4-119F. Internal and external connections for Euphonics Corp. Model C-47M semiconductor close-talking communications microphone.

lower surface. A notch (D) is cut through the epoxy to the inside surface of the lower strip (C), leaving a hinge for stress concentration. There is no electrical connection between the upper surface (B) and the lower surface strip (C). Above the notch is soldered a silicon semiconductor element (E) shaped in the form of a letter H, shown in the enlarged drawing at (F). The silicon fiber (E) is about 0.005 inch in thickness, and connects to the two larger end portions. A flexible yoke (G) is slipped over the silicon element, and connects to a 2-inch conical diaphragm (H) made of *Lexan*, a metal similar to aluminum but more durable. Two resilient pads (I) are used to support the ends of the epoxy beam, and to isolate it from shock and vibration. Contact is made to the two ends of the silicon element by low-resistance leads (J), held in place by the pressure of the pads (I).

The assembled semiconductor element and its modulating yoke is shown in Fig. 4-119C, and mounted in its supporting structure as seen in Fig. 4-119D. The semiconductor element has a frequency range from dc to 50,000 Hz; however, for this application the frequency range is limited from 100 to 6000 Hz, with a peak in the midrange high frequencies (Fig. 4-119E). The output impedance is 500 ohms. With 12 volts dc in series with 1000 ohms applied to the silicon element, the output level is minus 52 dB re: 1 mW/10 dynes/cm^2. The open-circuit voltage for close speech work is approximately 10 volts. A higher output level may be obtained by increasing the dc voltage. However, increasing the dc voltage requires that the series resistance also be increased to limit the current through the silicon element to a maximum current of 13 mA or less. With 20 volts dc and a series resistance of 2000 ohms, the output level is increased to minus 46 dB re: 1 mW/10 dynes/cm^2. A diagram of the internal and external connections is given in Fig. 4-119F.

4.120 What is a differential noise-cancelling microphone? — Differential noise-cancelling microphones are essentially designed for use in automo-

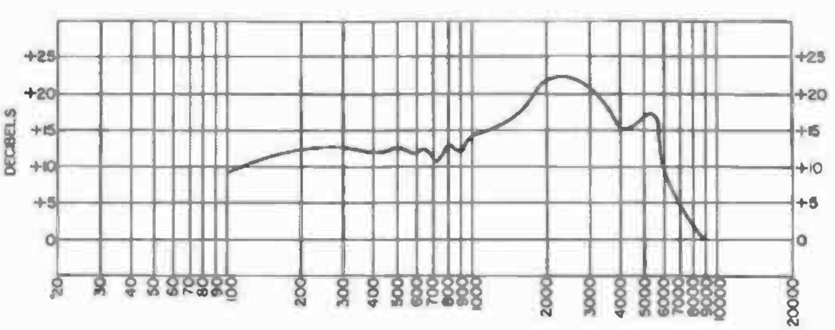


Fig. 4-119E. Frequency response of Euphonics Model C-47M close-talking communications semiconductor microphone.

biles, aircraft, boats, tanks, public-address systems, and industrial plants, or for any service where the ambient noise level is 100 dB or greater, and the microphone is hand-held. Two different designs of noise-cancelling microphones, manufactured by Electro-Voice, are given in Figs. 4-120A and B. The microphone in Fig. 4-120A is the Model 602F, of the dynamic type. Discrimination is afforded against all sounds originating more than ¼-inch from the front of the microphone. The noise-cancelling characteristic is achieved through the use of a balanced port opening, which conducts the unwanted sound to the rear of the dynamic unit diaphragm, and is out-of-phase with the sound arriving at the front of the microphone. The noise cancelling is most effective for frequencies above 2000 Hz. Only the speech originating within ¼ inch of the aperture is fully reproduced. The average discrimination between speech and noise is 20 dB or a ratio of 100:1 in intensity. The frequency response is 200 to 5000 Hz, with an output level of minus 60 dB. The impedance may be either high or 150 ohms. The polar pattern is bidirectional, pressure gradient.

A single-button carbon-granule microphone, Model 205KK is shown in Fig. 4-120B. The ambient noise is fed into two apertures, one above the other in such phase relationship to provide almost complete cancellation. Speech sounds originating ¼ inch from one of the two apertures is fully reproduced. Articulation is 97 percent under quiet conditions and 87 percent in an ambient noise level of 115 dB. The frequency response is 100 to 4000 Hz, with an output level of minus 50 dB re: 1 volt/dyne/cm², or 0.31 volt into 100 ohms. The internal resistance is 100 ohms and requires 10 to 50 mA of button current. The polar pattern is bidirectional, pressure gradient.

4.121 Describe the construction of an electrostatic-velocity microphone.—This microphone was developed about the same time as the first ribbon-velocity microphones. The structure is somewhat similar to an electrostatic loudspeaker. The sensitive elements consist of a flat insulated perforated plate, covered with a group of eight dural ribbons, about ¼ by 4 inches, and approximately 0.0002 inch in thickness. The perforated plate has about 80 holes per square inch. A polarizing voltage of 100 to 350 Vdc is applied between the ribbons and the back plate. Displacement of the ribbons results in the greater or lesser charge across the ribbons and back plate, which is applied to the control grid of an amplifier tube. In general, it may be considered to be a capacitor microphone. The frequency response is surprisingly good, being fairly constant from 40 to 5000 Hz; however, the harmonic distortion is rather high.

4.122 Describe the operating principles of a controlled-reluctance microphone.—This microphone operates on the principle that an electrical current is induced in a coil, located in a changing magnetic field. A magnetic armature is attached to a diaphragm suspended inside a coil. The diaphragm,

Fig. 4-120A. Electro-Voice Model 602F dynamic microphone.

Fig. 4-120B. Electro-Voice Model 205KK carbon-button microphone.

when disturbed by a sound wave, moves the armature and induces a corresponding *varying voltage* in the coil. High output with fairly good frequency response is typical of this type microphone.

4.123 Describe an inductor microphone.—An inductor microphone is a moving-coil type microphone employing a coil consisting of one turn of wire, suspended at the rear of a *parchment* diaphragm. The coil, because of the diaphragm movement, is caused to move in a strong magnetic field generating minute voltages. An impedance-matching transformer matches the low impedance of the coil to the transmission line. The field pattern is omnidirectional. The inductor microphone is now *obsolete*.

4.124 What is the standard for phasing microphones? — Microphones may be phased against a microphone known to be in phase, by the method discussed in Question 4.85. The standard method as recommended by the EIA for manufacturers, specifies that the polarity of the microphone or a transducer element refers to in-phase or out-of-phase conditions of voltage developed at the microphone terminals, with respect to the sound pressure of a sound wave causing the voltage. Exact in-phase relationship can be taken to mean that the phase of the voltage is coincident with the phase of the sound-pressure wave causing the voltage. In practice, this relationship may not always be obtainable.

It is important to keep in mind that if a preamplifier is connected between the transducer element and the output terminals, it may cause a *phase reversal*. For capacitor microphones, the inward motion of the diaphragm, because of a positive increment of pressure by the sound wave, produces an increase in capacitance. Since the charge ($Q = CE$) remains constant, an increase in capacitance will cause a decrease in voltage across the electrodes. Therefore, in a pressure-operated capacitor transducer, the in-phase terminal is the terminal to be connected to the negatively charged electrode.

For a piezoelectric microphone, the phasing is determined by the orientation of the molecules in the element, which may not be known beforehand. This may be ascertained by a slight pressure applied to the sensitive element, and the polarity of the resulting voltage, determined by observing the motion of the trace on an oscilloscope, connected at the output of the microphone. In this instance, the phasing of the oscilloscope must be established beforehand, by use of a battery. (See Question 23.218.)

Magnetic, dynamic, and ribbon microphones may be phased by the application of a low-voltage direct current of known polarity, applied to the termi-

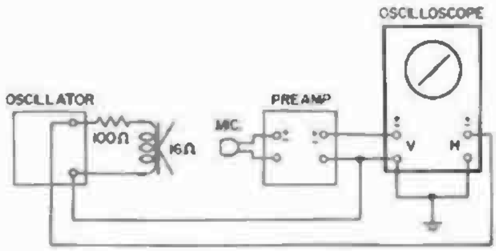


Fig. 4-124A. Test setup for checking the phase of a microphone per the EIA standards.

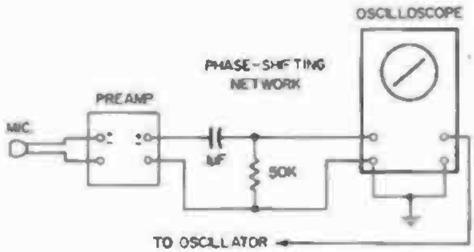


Fig. 4-124B. Phase-shift network for use with a velocity microphone.

nals of the transducer element, which will cause a motion of the element. The in-phase terminal is the terminal which, when the positive polarity is applied to the transducer element, moves *away* from the observer facing the front of the microphone. Great care must be exercised when performing the described measurements, as the moving element can be very easily damaged.

A microphone-phasing test setup is shown in Fig. 4-124A, using a loudspeaker for the sound source, and an oscilloscope for observing the phase conditions. To establish the proper phase at the loudspeaker, the speaker is mounted in an infinite baffle and connected to an oscillator through a 100-ohm resistor. With the resistor in series with the voice coil, the current through the voice coil is then in phase with the voltage at the terminals of the oscillator. If the applied frequency of the oscillator is above the resonant frequency of the speaker, the pressure will be in phase with the applied voltage. For an 8-inch speaker having a resonant frequency of 60 Hz, this condition may be attained at frequencies between 100 and 400 Hz.

To ascertain the in-phase terminal of the speaker, a battery is connected across the voice-coil terminals in such a manner that the cone moves *toward* the microphone; the terminal connected to the positive terminal of the battery is the in-phase terminal of the speaker. With the oscilloscope connected as in Fig. 4-124A, microphones are tested by adjusting the oscillator output for a suitable acoustic output of the speaker. The microphone is positioned at a distance of about 2 inches from the speaker, and the orientation of the oscilloscope trace checked. If the trace is in the form of a slanted line or ellipse, with its major axis orientated from lower left to upper right, then the in-phase terminal of the microphone is the terminal connected to the high potential (±) terminal of the amplifier. This relationship should remain over a range of 100 to 400 Hz.

When this procedure is applied to a gradient microphone (velocity), the trace will form a circle, because the out-of-phase relationship between the pressure and velocity is a spreading waveform. This may be corrected by the use of a phase-shift network, consisting of a 50K resistor and a 0.1-uF capacitor, as shown in Fig. 4-124B. Except for this detail the measurement procedure is the same.

The in-phase terminal of a microphone is the number-1 terminal. The in-phase terminal of a loudspeaker and other equipment is generally marked with a plus-minus sign.

4.125 Describe the basic design of an ultrasonic microphone.—Ultrasonic microphones are used in television receiver remote-control systems, opening of garage doors, control of air-conditioning equipment, slide projectors, gas-leak detectors, and for equipment designed for the assistance in directing blind persons. These microphones are sensitive to frequencies only in the range of 25,000 to 40,000 Hz. For the most part, they are manufactured using a two-plate element of lead-zirconium titanate formed into thin, square plates. The size of the plate determines the response to frequency.

The plate is clamped at four nodal points of minimum motion, to ensure a maximum output voltage. The center of the element is masked so that only the corners are driven directly by the ultrasonic sound waves. The spacing between the center mask and the element is such that sound reaching the center of the plate is changed in phase to correspond with the different phase relationships of the center and corners of the plate. The space behind the plate creates a resonate cavity, and the air compliance assists in control of the plate motion.

The output level of an ultrasonic microphone is on the order of minus 65 dB re: 1 V/dyne/cm^2. The resonant frequency is controlled in manufactnre to within 500 Hz by adjusting the size of each individual ceramic element. This is quite necessary since the elements must respond to several different frequencies for control purposes. The remote exciter unit generally consists of a transistor oscillator with a similar transducer generating frequencies between 25,000 and 40,000 Hz.

Ultrasonic microphones may also be designed to use electrostatic elements similar to an electrostatic loudspeaker, as they are broadly nonresonant, and can be operated over a large number of individual frequencies for circuit-control functions.

4.126 Describe an electret capacitor microphone.—This is a relatively new development in capacitor microphones by Sessler and West of the Bell Telephone Laboratories. The device utilizes a foil electret (see Question 25.51) and therefore requires no polarizing voltage. The frequency response is within plus-minus 3 dB, 50 to 15,000 Hz. Sensitivities of 50 μV re: 1 microbar have been achieved. Because of its relatively low impedance, it can be connected directly to a conventional transistor amplifier.

Like any capacitor microphone, the electret capacitor microphone depends for its operation on minute variations of capacitance produced as sound vibrations impinge on one flexible plate of the capacitor head. However, in this instance, the flexible plate is a *foil electret*, constructed of a thinly metalized sheet of fluorocarbon or polycarbonate. The foil contains a permanent static charge. Since its spacing is varied from the fixed element, the electrostatic field is varied, thus producing a varying voltage at the output terminals. Because of the static charge no polarizing voltage is required. The thin film has a thickness of 0.00012 to 0.001 inch, therefore the capacitance of an electret microphone is about three times that of a capacitor microphone of comparable dimensions, with lower impedance.

The foil is polarized by heating it to about 200 degrees Centigrade while it is held between a pair of charged metal plates (spaced 2-mm) which create an electrostatic field of between 10 and 100V/cm. Charges identical in sign to the adjacent plates migrate to the electret, where they remain after cooling.

The sensitivity of the microphone remains essentially constant over very long periods of time. Measurements indicate that the sensitivity will fall 50 percent in about 100 years. Because of its simple, rugged construction and because it is immune to wide temperature changes, it will no doubt find a wide range of uses in the sound industry, as well as in the recording industry.

For sound pressures of 120 to 140 dB (re: 0.0002 dyne/cm^2), the distortion remains about 1 percent, rising to around 5 percent at 150 dB.

4.127 What is the proximity effect in a microphone?—It is the increase in low-frequency response noted in most pressure-gradient microphones, when the distance from the sound source is decreased, and is most noticeable at distances of less than two feet. This effect is not encountered in omnidirectional microphones. The proximity effect for a bidirectional microphone is shown graphically in Fig. 4-57.

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Section 5

Attenuators

Attenuator networks have been in use since the inception of the telephone, for controlling sound levels and the matching of impedances. Many of the present day configurations are the work of Otto J. Zobel, W. H. Bode, R. L. Diezold, Sallie Pero Mead, and T. E. Shay, all of the Bell Telephone Laboratories. Tables of constants developed by P. K. McElroy (also of Bell Telephone Laboratories) for various values of expression and substitution in equations that have long been a time-saver for the design engineer and have resulted in greater accuracy are included.

In this section, networks both balanced and unbalanced, fixed and variable, impedance matching, combining, bridging, mixer controls, their design and use are discussed. The combining of the various configurations is explained, with tables of constants to reduce computation to simple mathematics.

5.1 *What does the term attenuation mean?*—The reduction of a sound wave or electrical energy in an electrical circuit.

5.2 *What is an attenuator?*—An arrangement of noninductive resistors in an electrical circuit to reduce the strength of an audio or radio frequency signal without introducing appreciable distortion. Attenuators may be fixed in value or they may be variable. If they are variable, they are generally designed to reduce the signal in a logarithmic manner. The configuration may take one of several forms. Two commercially manufactured fixed attenuators are shown in Fig. 5-2. Similar networks are available in steps of 0.5 dB up to 40 dB for any impedance up to 600 ohms.

5.3 *What is a pad?*—It is another name for an attenuator.

5.4 *What does the term network mean?*—A configuration of circuit ele-

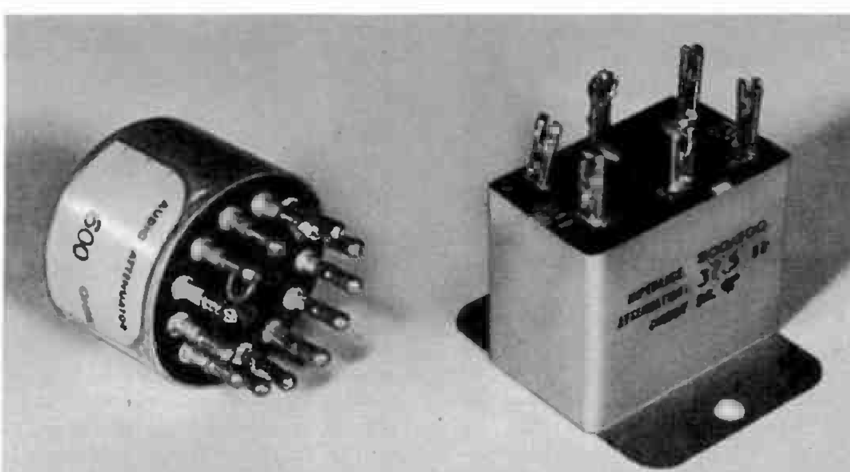


Fig. 5-2. Fixed attenuators. At the left is a "T" -type network that can be varied in steps of 5, 10, and 20 dB, with a total loss of 35 dB. At the right is an "H" network, with a fixed loss of 37.5 dB.

ments such as those used in filters, attenuators, or equalizers.

5.5 ***What does the term configuration mean?***—A circuit or network of circuit elements.

5.6 ***Define the term loss.***—A decrease in the power, voltage, or current at the output of a device compared to the power, voltage, or current at the input of the device. The loss in decibels may be calculated by means of one of the following equations:

$$\text{dB loss} = 10 \operatorname{Log}_{10} \frac{P_1}{P_2} \text{ or,}$$

$$= 20 \operatorname{Log}_{10} \frac{V_1}{V_2} \text{ or,}$$

$$= 20 \operatorname{Log}_{10} \frac{I_1}{I_2}$$

where,

P_1 is the power at the input,
P_2 is the power at the output,
V_1 is the voltage at the input,
V_2 is the voltage at the output,
I_1 is the current at the input,
I_2 is the current at the output.

5.7 ***What is insertion loss?***—The loss created by the insertion of a device in an electrical circuit. The resulting loss is generally expressed in decibels.

5.8 ***What is an impedance-matching network?***—A noninductive, resistive network designed for insertion between two or more circuits of equal or unequal impedance. When properly designed, the network reflects correct impedance to each branch of the circuit.

5.9 ***What is an artificial line?***—A configuration of resistance, capacitance, and inductance representing the electrical characteristics of a transmission line. Attenuators used for reducing the transmission level of a line consist of pure resistance and a uniform frequency characteristic over the whole transmission band. Unless an artificial line is designed as described in Question 25.182, it will not have the electrical characteristics of the line.

5.10 ***What is an attenuation characteristic?***—A graphical presentation of the loss in decibels of a device versus frequency. Attenuators constructed using noninductive resistors offer no discrimination in the normal transmission band. The loss is the same for all frequencies.

5.11 ***What is the difference between the impedance and the dc resistance of an attenuator?***—None, if the resistors are noninductive and correctly terminated. The terms resistance and impedance are used interchangeably with attenuators.

5.12 ***Under what conditions does an attenuator show its correct loss?***—Only when terminated in its characteristic impedance at both the input and output.

5.13 ***What is a line pad and its purpose?***—A line pad is a resistive network placed between the output of a line and the device fed by the line to effect an impedance match. A line pad also supplies a definite termination to the line and to the equipment connected to the line. Line pads are also used to secure a given amount of isolation between the line and the equipment terminating the line. Such a circuit is shown in Fig. 5-13.

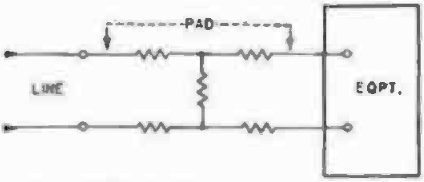


Fig. 5-13. A line-terminating pad.

5.14 ***What does the term "so many decibels of isolation" mean?***—It is a term used to indicate that an attenuator has been connected between the line and a piece of equipment. Isolation, in the form of a pad, is often necessary to supply the correct terminating impedance and to isolate equipment from a line which has been equalized. Isolation pads are also used between filters, equalizers, and amplifiers to isolate them from each other and prevent interaction between them which might upset their frequency characteristics. Pads are also used with impedance-sensitive devices. A typical circuit showing an isolation pad between a low-pass filter and an equalizer is shown in Fig. 5-14.

5.15 ***What is a fader?***—A continuously variable attenuator designed to pass a signal from one signal source to another without interruption, and to provide a smooth transition from one to the other. A typical fader-control circuit is shown in Fig. 5-15.

5.16 ***What is a noninductive resistor?***—A resistor having little or no self-inductance. Such resistors are wound using special winding techniques. A

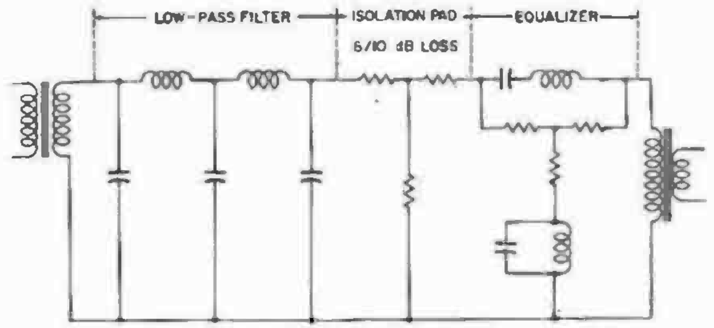


Fig. 5-14. An isolation pad connected between a low-pass filter and an equalizer.

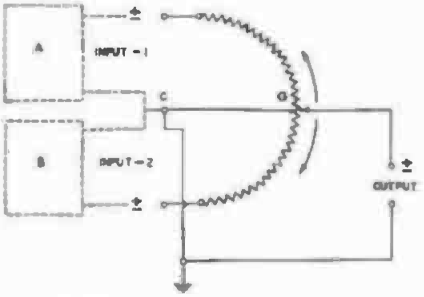


Fig. 5-15. A two-position continuously variable fader. Both signals are at zero when the arm is at the center position.

slotted-form, pi-wound resistor is shown in Fig. 5-16A. In Fig. 5-16B is shown a parallel (bifilar) wirewound resistor. The resistance wire is formed into a loop, then wound with the two wires parallel to each other and in layers. Winding them in this manner cancels the self-inductance, for all practical purposes. Fig. 5-16C shows a group of noninductive wirewound resistors.

5.17 ***May carbon-composition resistors be used for attenuators?***—Yes. They are noninductive at audio frequencies and are convenient to use. (See Question 5.87.) However, some resistors which appear to be carbon are actually wirewound. This may be determined by checking the manufacturer's specifications.

5.18 ***What are the different type windings used for noninductive resistors called?***—Mica-card, Ayrton-Perry, reversed-loop, figure-8, hairpin, fishline, woven-tape, bifilar, bifilar-series, and the slotted-form reversed winding. Slotted-form resistors (shown in Fig. 5-16C) may be used for grid, cathode, plate, and screen resistors in amplifiers, and many other places where noninductive resistors are required.

5.19 ***What tolerance value resistors should be used in attenuators?***—For precision attenuators, one percent; for general test work, five percent.

5.20 ***What effect does an attenuator have on the phase relationships in an electrical circuit?***—None, if the attenuator is constructed of noninductive resistors and the distributed capacity of the circuit is held to a minimum.

5.21 ***What does the term minimum loss mean when applied to an attenuator?***—The minimum loss value for which an attenuator may be designed using a given configuration for a given value of source and load impedance. The minimum loss for attenuators of unequal impedance may be read from the graph of Fig. 5-21.

The graph is entered at the bottom at the desired impedance ratio, then followed vertically until it intersects the diagonal line. The minimum loss is then read at the left margin in decibels. As an example: assume an impedance of 600 ohms is to be matched to an impedance of 150 ohms; this is an impedance

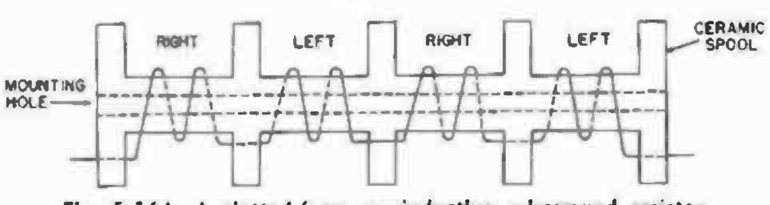


Fig. 5-16A. A slotted-form, noninductive, wirewound resistor.

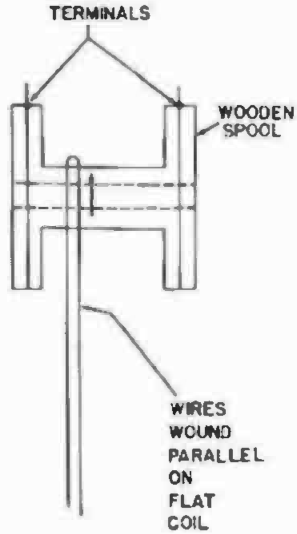


Fig. 5-16B. A parallel-wound (bifilar) noninductive, wirewound resistor.

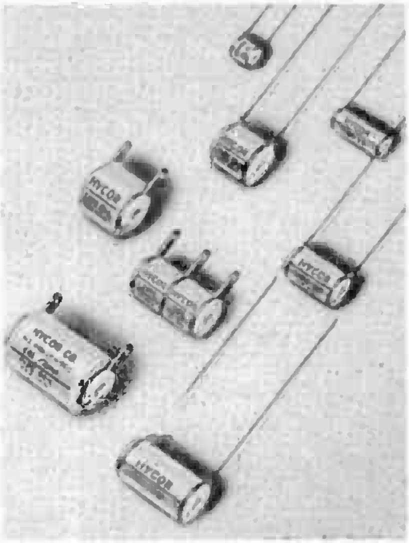


Fig. 5-16C. Noninductive, wirewound resistors that are manufactured by Kelvin Associates.

ratio of four. For this ratio, the graph indicates a minimum loss of 11.5 dB. This is the lowest value for which the attenuator can be designed. In actual practice the network would be designed for a loss of 12 to 15 dB.

5.22 What is the "K" factor used in attenuator equations?—It is the ratio of current, voltage, or power corresponding to a given value of attenuation expressed in decibels. To simplify the calculation of attenuator networks, the values of the most frequently used expressions as tabulated by P. K. McElroy are given in Fig. 5-22. The various values of the expressions are substituted in the equations, thus saving much time and resulting in greater accuracy.

5.23 If attenuators are connected in tandem (series) how is the loss computed?—The internal connections for an

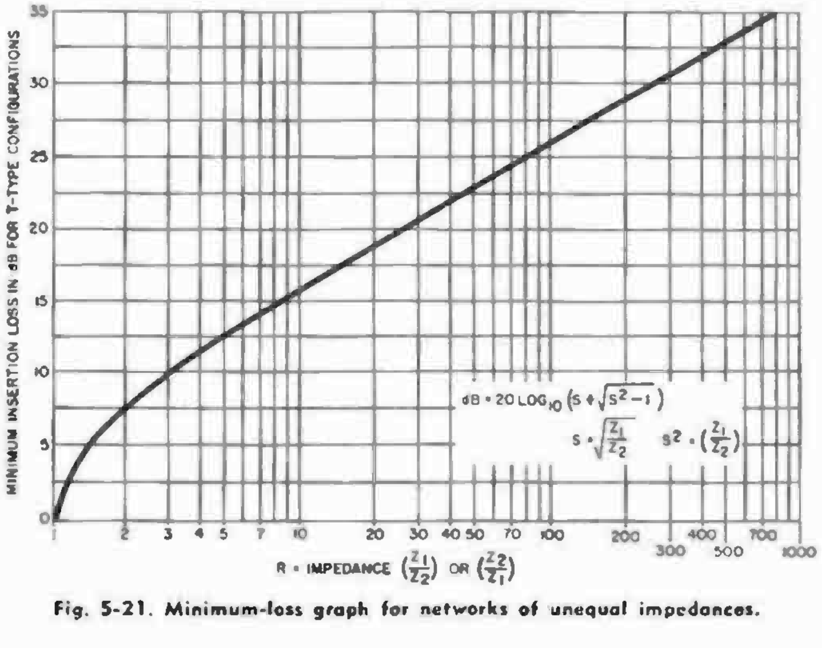


Fig. 5-21. Minimum-loss graph for networks of unequal impedances.

$n(dB)$	$r = \frac{1}{K}$	K	K^2	$\frac{K-1}{K+1}$	$\frac{K+1}{K-1}$	$\frac{K}{K^2-1}$	$\frac{K^2-1}{K} = K - r$	$\frac{K^2+1}{K^2-1}$	$\frac{K-1}{K} = 1 - r$	$\frac{K}{K-1} = \frac{1}{1-r}$	$\frac{1}{K-1}$	$n(dB)$
0.05	0.994260	1.0057731	1.011579	0.0028783	347.43	86.8618	0.011513	173.73	0.0057395	174.22	173.22	0.05
0.1	0.98855	1.011579	1.023292	0.0057562	173.73	43.4303	0.023029	86.866	0.011448	87.363	86.363	0.1
0.2	0.97724	1.023292	1.047128	0.011512	86.866	21.713	0.046052	43.437	0.022762	33.933	32.933	0.2
0.3	0.96605	1.035143	1.071520	0.017268	57.911	14.473	0.069093	28.965	0.033949	19.455	18.455	0.3
0.4	0.95499	1.047128	1.096477	0.023022	43.437	10.854	0.092138	21.730	0.045007	22.219	21.219	0.4
0.5	0.94406	1.059254	1.12202	0.028774	34.754	8.6810	0.11519	17.391	0.055939	17.877	16.877	0.5
0.6	0.93325	1.071520	1.14815	0.034525	28.965	7.2327	0.13827	14.499	0.066746	14.982	13.982	0.6
0.7	0.92257	1.083928	1.17490	0.040274	24.830	6.1974	0.16136	12.435	0.077428	12.915	11.915	0.7
0.8	0.91201	1.096477	1.20227	0.046019	21.730	5.4209	0.18447	10.888	0.087988	11.365	10.365	0.8
0.9	0.90157	1.10917	1.23027	0.051763	19.319	4.8168	0.20760	9.6853	0.098429	10.160	9.1600	0.9
1.0	0.89125	1.12202	1.25893	0.057502	17.391	4.3335	0.23077	8.7237	0.10875	9.1934	8.1954	1.0
1.1	0.88105	1.13501	1.28825	0.063237	15.814	3.9376	0.25396	7.9384	0.11895	8.4069	7.4069	1.1
1.2	0.87096	1.14815	1.31826	0.068968	14.499	3.6076	0.27719	7.2842	0.12904	7.7499	6.7499	1.2
1.3	0.86099	1.16145	1.34896	0.074695	13.388	3.3283	0.30046	6.7313	0.13901	7.1939	6.1939	1.3
1.4	0.85114	1.17490	1.38038	0.080418	12.435	3.0888	0.32378	6.2579	0.14886	6.7176	5.7176	1.4
1.5	0.84139	1.18850	1.41254	0.086132	11.610	2.8809	0.34711	5.8480	0.15861	6.3050	5.3050	1.5
1.6	0.83176	1.20227	1.44544	0.091846	10.888	2.6991	0.37051	5.4899	0.16824	5.9439	4.9439	1.6
1.7	0.82224	1.21618	1.47911	0.097551	10.251	2.5384	0.39394	5.1744	0.17776	5.6258	4.6258	1.7
1.8	0.81283	1.23027	1.51356	0.103249	9.6853	2.3956	0.41744	4.8944	0.18717	5.3427	4.3427	1.8
1.9	0.80353	1.24452	1.54882	0.108939	9.1794	2.2676	0.44099	4.6442	0.19647	5.0897	4.0897	1.9
2.0	0.79433	1.25893	1.58489	0.11463	8.7241	2.1523	0.46460	4.4195	0.20567	4.8620	3.8620	2.0
2.2	0.77625	1.28825	1.65959	0.12597	7.9384	1.9531	0.51200	4.0322	0.22375	4.4692	3.4692	2.2
2.4	0.75858	1.31826	1.73780	0.13728	7.2842	1.7867	0.55968	3.7108	0.24142	4.1421	3.1421	2.4
2.5	0.74989	1.33352	1.77828	0.14293	6.9966	1.7133	0.58363	3.5698	0.25011	3.9983	2.9983	2.5
2.6	0.74131	1.34896	1.81970	0.14856	6.7313	1.6457	0.60765	3.4399	0.25829	3.8657	2.8657	2.6
2.8	0.72444	1.38038	1.90546	0.15980	6.2579	1.5245	0.65594	3.2088	0.27556	3.6289	2.6289	2.8
3.0	0.70795	1.41254	1.99526	0.17100	5.8480	1.4192	0.70459	3.0095	0.29205	3.4240	2.4240	3.0
3.2	0.69183	1.44544	2.08930	0.18215	5.4899	1.3269	0.75361	2.8360	0.30817	3.2450	2.2450	3.2
3.4	0.67608	1.47911	2.18776	0.19326	5.1744	1.2453	0.80302	2.6838	0.32392	3.0872	2.0872	3.4
3.5	0.66834	1.49623	2.2387	0.19879	5.0304	1.2079	0.82789	2.6147	0.33166	3.0152	2.0152	3.5
3.6	0.66069	1.51356	2.2909	0.20432	4.8944	1.1725	0.85289	2.5493	0.33931	2.9472	1.9472	3.6
3.8	0.64565	1.54882	2.3988	0.21532	4.6442	1.1072	0.90314	2.4298	0.35435	2.8221	1.8221	3.8
4.0	0.63096	1.58489	2.5119	0.22627	4.4194	1.0483	0.95393	2.3229	0.36904	2.7097	1.7097	4.0
4.5	0.59566	1.67880	2.8184	0.25340	3.9464	0.92323	1.08314	2.0999	0.40434	2.4732	1.4732	4.5
5.0	0.56234	1.77828	3.1623	0.28013	3.5698	0.82241	1.21594	1.9249	0.43766	2.2849	1.2849	5.0
5.5	0.53088	1.88365	3.5481	0.30643	3.2633	0.73922	1.35277	1.7849	0.46912	2.1317	1.1317	5.5
6.0	0.50119	1.99526	3.9811	0.33228	3.0095	0.66932	1.49407	1.6709	0.49881	2.0048	1.0048	6.0
6.5	0.47315	2.1135	4.4668	0.35764	2.7961	0.60964	1.6403	1.5769	0.52685	1.89807	0.89807	6.5
7.0	0.44668	2.2387	5.0119	0.38246	2.6146	0.55801	1.7920	1.4985	0.55332	1.80730	0.80730	7.0
7.5	0.42170	2.3714	5.6234	0.40677	2.4584	0.51291	1.9497	1.4326	0.57830	1.72918	0.72918	7.5

Fig. 5-22. "K" factors for calculating attenuator loss values.

$n(dB)$	$r=\frac{1}{K}$	K	K^2	$\frac{K-1}{K+1}$	$\frac{K+1}{K-1}$	$\frac{K}{K^2-1}$	$\frac{K^2-1}{K}=K-r$	$\frac{K^2+1}{K^2-1}$	$\frac{K-1}{K}=1-r$	$\frac{K}{K-1}=\frac{1}{1-r}$	$\frac{1}{K-1}$	$n(dB)$
8.0	0.39811	2.5119	6.3096	0.43051	2.3228	0.47309	2.1138	1.3767	0.60189	1.66142	0.66142	8.0
8.5	0.37584	2.6607	7.0795	0.45366	2.2043	0.43765	2.2849	1.3290	0.62416	1.60216	0.60216	8.5
9.0	0.35481	2.8184	7.9433	0.47622	2.0999	0.40592	2.4636	1.2880	0.64519	1.54993	0.54993	9.0
9.5	0.33497	2.9854	8.9125	0.49817	2.0074	0.37730	2.6504	1.2528	0.66503	1.50368	0.50368	9.5
10.0	0.31623	3.1623	10.000	0.51950	1.9249	0.35137	2.8561	1.2222	0.68377	1.46247	0.46247	10.0
10.5	0.29854	3.3497	11.220	0.54020	1.8512	0.32775	3.0512	1.1957	0.70146	1.42559	0.42559	10.5
11.0	0.28184	3.5481	12.589	0.56026	1.7849	0.30616	3.2663	1.1726	0.71816	1.39245	0.39245	11.0
11.5	0.26607	3.7584	14.125	0.57969	1.7251	0.28635	3.4923	1.1524	0.73393	1.36253	0.36253	11.5
12.0	0.25119	3.9811	15.849	0.59848	1.6709	0.26811	3.7299	1.1347	0.74881	1.33545	0.33545	12.0
12.5	0.23714	4.2170	17.783	0.61664	1.6217	0.25127	3.9799	1.1192	0.76286	1.31086	0.31086	12.5
13.0	0.22387	4.4668	19.953	0.63416	1.5769	0.23568	4.2429	1.1055	0.77613	1.28845	0.28845	13.0
13.5	0.21135	4.7315	22.387	0.65105	1.5360	0.22123	4.5202	1.0935	0.78865	1.26799	0.26799	13.5
14.0	0.19953	5.0119	25.119	0.66733	1.4983	0.20780	4.8124	1.0829	0.80047	1.24926	0.24926	14.0
14.5	0.18836	5.3088	28.184	0.68298	1.4642	0.19529	5.1204	1.0736	0.81164	1.23208	0.23208	14.5
15.0	0.17783	5.6234	31.623	0.69804	1.4326	0.18363	5.4456	1.0653	0.82217	1.21629	0.21629	15.0
15.5	0.16788	5.9566	35.481	0.71250	1.4035	0.17275	5.7887	1.0580	0.83212	1.20175	0.20175	15.5
16.0	0.15849	6.3096	39.811	0.72639	1.3767	0.16257	6.1511	1.0615	0.84151	1.18834	0.18834	16.0
16.5	0.14962	6.6834	44.668	0.73970	1.3519	0.15305	6.5338	1.0458	0.85038	1.17595	0.17595	16.5
17.0	0.14125	7.0795	50.119	0.75246	1.3290	0.14413	6.9382	1.04071	0.85875	1.16449	0.16449	17.0
17.5	0.13335	7.4989	56.234	0.76468	1.3077	0.13577	7.3655	1.03621	0.86665	1.15387	0.15387	17.5
18.0	0.12589	7.9433	63.096	0.77637	1.2880	0.12792	7.8174	1.03220	0.87411	1.14402	0.14402	18.0
18.5	0.118850	8.4139	70.796	0.78755	1.2698	0.12055	8.2950	1.02866	0.88115	1.13488	0.13488	18.5
19.0	0.112202	8.9125	79.433	0.79823	1.2528	0.11363	8.8003	1.02550	0.88780	1.12638	0.12638	19.0
19.5	0.105925	9.4406	89.125	0.80844	1.2369	0.10713	9.3347	1.02269	0.89407	1.11847	0.11847	19.5
20.0	0.100000	10.0000	100.000	0.81818	1.2222	0.10101	9.9000	1.02020	0.90000	1.11111	0.11111	20.0
20.5	0.094406	10.5925	112.202	0.82747	1.2085	0.095255	10.498	1.01799	0.90559	1.10425	0.10425	20.5
21.0	0.089125	11.2202	125.893	0.83634	1.1957	0.089841	11.131	1.01601	0.91087	1.09785	0.097845	21.0
21.5	0.084139	11.8850	141.254	0.84478	1.1837	0.084739	11.801	1.01426	0.91586	1.09187	0.091870	21.5
22.0	0.079433	12.589	158.49	0.85282	1.1726	0.079935	12.510	1.01270	0.92057	1.08629	0.086291	22.0
22.5	0.074989	13.335	177.83	0.86048	1.1621	0.075411	13.260	1.01126	0.92501	1.08107	0.081070	22.5
23.0	0.070795	14.125	199.53	0.86777	1.1524	0.071148	14.054	1.01007	0.92921	1.07619	0.076190	23.0
23.5	0.066834	14.962	223.87	0.87470	1.1432	0.067133	14.895	1.00897	0.93317	1.07162	0.071623	23.5
24.0	0.063096	15.849	251.19	0.88130	1.1347	0.063348	15.786	1.00799	0.93690	1.06734	0.067345	24.0
24.5	0.059566	16.788	281.84	0.88756	1.1267	0.059778	16.728	1.00712	0.94043	1.06334	0.063339	24.5
25.0	0.056234	17.783	316.23	0.89352	1.1192	0.056413	17.727	1.00634	0.94377	1.05958	0.059584	25.0
25.5	0.053088	18.836	354.81	0.89917	1.1121	0.053238	18.783	1.00565	0.94691	1.05607	0.056066	25.5
26.0	0.050119	19.953	398.11	0.90455	1.1055	0.050246	19.903	1.00504	0.94988	1.05276	0.052762	26.0
26.5	0.047315	21.135	446.68	0.90965	1.0993	0.047422	21.088	1.00449	0.95268	1.04966	0.049665	26.5
27.0	0.044668	22.387	501.19	0.91448	1.0935	0.044757	22.342	1.00400	0.95533	1.04676	0.046757	27.0
27.5	0.042170	23.714	562.34	0.91907	1.0881	0.042245	23.672	1.00356	0.95783	1.04403	0.044026	27.5
28.0	0.039811	25.119	630.96	0.92343	1.0829	0.039874	25.079	1.00317	0.96019	1.04146	0.041461	28.0

"K" factors—(Continued)

28.5	0.037584	26.607	707.95	0.92755	1.0781	0.037636	26.569	1.00283	0.96242	1.03905	0.039052	28.5
29.0	0.035481	28.184	794.33	0.93147	1.0736	0.035526	28.149	1.00252	0.96452	1.03679	0.036786	29.0
29.5	0.033497	29.854	891.25	0.93518	1.0693	0.033534	29.821	1.00225	0.96650	1.03468	0.034657	29.5
30.0	0.031623	31.623	1,000.0	0.93869	1.0653	0.031655	31.591	1.00200	0.96836	1.03266	0.032655	30.0
31.0	0.028184	35.481	1,258.9	0.94518	1.0580	0.028207	35.453	1.00159	0.97182	1.02900	0.029001	31.0
31.5	0.026607	37.584	1,412.5	0.94817	1.0547	0.026627	37.558	1.00142	0.97339	1.02733	0.027334	31.5
32.0	0.025119	39.811	1,584.9	0.95099	1.0515	0.025135	39.786	1.00126	0.97488	1.02577	0.025766	32.0
33.0	0.022387	44.668	1,995.3	0.95621	1.0458	0.022398	44.646	1.00100	0.97761	1.02290	0.022900	33.0
34.0	0.019953	50.119	2,511.9	0.96088	1.04072	0.019961	50.099	1.00080	0.98005	1.02036	0.020359	34.0
34.5	0.018836	53.088	2,818.4	0.96302	1.03840	0.018843	53.069	1.00071	0.98116	1.01920	0.019198	34.5
35.0	0.017783	56.234	3,162.3	0.96506	1.03621	0.017789	56.216	1.00063	0.98222	1.01810	0.018105	35.0
36.0	0.015849	63.096	3,981.1	0.96880	1.03221	0.015853	63.080	1.00050	0.98415	1.01610	0.016104	36.0
37.0	0.014125	70.795	5,011.9	0.97214	1.02866	0.014128	70.781	1.00040	0.98588	1.01433	0.014328	37.0
37.5	0.013335	74.989	5,623.4	0.97368	1.02703	0.013338	74.976	1.00036	0.98666	1.01352	0.013516	37.5
38.0	0.012589	79.433	6,309.6	0.97513	1.02550	0.012591	79.420	1.00032	0.98741	1.01275	0.012750	38.0
39.0	0.0112202	89.125	7,943.3	0.97781	1.02270	0.0112216	89.114	1.00025	0.98878	1.01135	0.011348	39.0
40.0	0.0100000	100.000	10,000.	0.98020	1.02020	0.0100010	99.990	1.00020	0.99000	1.01010	0.010101	40.0
40.5	0.0094406	105.925	11,220.	0.98130	1.01906	0.0094414	105.916	1.00018	0.99056	1.00953	0.0095306	40.5
41.0	0.0089125	112.202	12,589.	0.98233	0.01799	0.0089134	112.193	1.00016	0.99109	1.00899	0.0089926	41.0
42.0	0.0079433	125.89	15,849.	0.98424	1.01601	0.0079436	125.88	1.00013	0.99206	1.00801	0.0080070	42.0
43.0	0.0070795	141.25	19,953.	0.98594	1.01426	0.0070795	141.24	1.00010	0.99292	1.00713	0.0071301	43.0
43.5	0.0066834	149.62	22,387.	0.98672	1.01346	0.0066834	149.61	1.00009	0.99332	1.00673	0.0067286	43.5
44.0	0.0063096	158.49	25,119.	0.98746	1.01270	0.0063096	158.49	1.00008	0.99369	1.00635	0.0063496	44.0
45.0	0.0056234	177.83	31,623.	0.98887	1.01131	0.0056234	177.83	1.00006	0.99438	1.00566	0.0056551	45.0
46.0	0.0050119	199.53	39,811.	0.99003	1.01007	0.0050119	199.53	1.00005	0.99499	1.00504	0.0050370	46.0
46.5	0.0047315	211.35	44,668.	0.99058	1.00951	0.0047315	211.35	1.000045	0.99527	1.00475	0.0047540	46.5
47.0	0.0044668	223.87	50,119.	0.99111	1.00897	0.0044668	223.87	1.000040	0.99553	1.00449	0.0044869	47.0
48.0	0.0039811	251.19	63,096.	0.99207	1.00799	0.0039811	251.19	1.000032	0.99602	1.00400	0.0039970	48.0
49.0	0.0035481	281.84	79,433.	0.99293	1.00712	0.0035481	281.84	1.000025	0.99645	1.00356	0.0035607	49.0
50.0	0.0031623	316.23	100,000.	0.99370	1.00634	0.0031623	316.23	1.000020	0.99684	1.00317	0.0031723	50.0
51.0	0.0028184	354.81	125,890.	0.99438	1.00565	0.0028184	354.81	1.000016	0.99718	1.00283	0.0028264	51.0
52.0	0.0025119	398.11	158,490.	0.99499	1.00504	0.0025119	398.11	1.000013	0.99749	1.00252	0.0025182	52.0
54.0	0.0019953	501.19	251,190.	0.99602	1.00400	0.0019953	501.19	1.000008	0.99801	1.00200	0.0019992	54.0
55.0	0.0017783	562.34	316,230.	0.99645	1.00356	0.0017783	562.34	1.000006	0.99822	1.00178	0.0017815	55.0
56.0	0.0015849	630.96	398,110.	0.99684	1.00317	0.0015849	630.96	1.000005	0.99842	1.00159	0.0015874	56.0
57.0	0.0014125	707.95	501,190.	0.99718	1.00283	0.0014125	707.95	1.000004	0.99859	1.00141	0.0014145	57.0
58.0	0.0012589	794.33	630,960.	0.99749	1.00252	0.0012589	794.33	1.000003	0.99874	1.00126	0.0012605	58.0
60.0	0.0010000	1,000.0	10^6	0.99800	1.00200	0.0010000	1,000.0	1.000002	0.99900	1.00100	0.0010010	60.0
65.0	0.00056234	1,778.3	3.1623×10^6	0.99888	1.00112	0.00056234	1,778.3	1.000001	0.99944	1.00056	0.00056265	65.0
70.0	0.00031623	3,162.3	10^7	0.99937	1.00063	0.00031623	3,162.3	1.000000	0.99968	1.00032	0.00031633	70.0
75.0	0.00017783	5,623.4	3.1623×10^7	0.99964	1.00036	0.00017783	5,623.4	1.000000	0.99982	1.00018	0.00017786	75.0
80.0	0.00010000	10,000.	10^8	0.99980	1.00020	0.00010000	10,000	1.000000	0.99990	1.00010	0.00010001	80.0
85.0	0.000056234	17,783.	3.1623×10^8	0.99989	1.00011	0.000056234	17,783.	1.000000	0.99994	1.00006	0.000056237	85.0
90.0	0.000031623	31,623.	10^9	0.99994	1.00006	0.000031623	31,623.	1.000000	0.99997	1.00003	0.000031624	90.0
95.0	0.000017783	56,234.	3.1623×10^9	0.99996	1.00004	0.000017783	56,234.	1.000000	0.99998	1.00002	0.000017783	95.0
100.0	0.000010000	10^5	10^{10}	0.99998	1.00002	0.000010000	10^5	1.000000	0.99999	1.00001	0.000010000	100.0

"K" factors—(Continued)

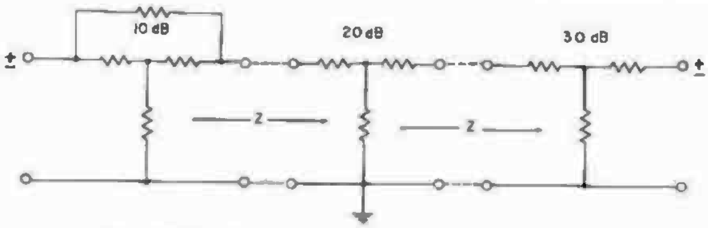


Fig. 5-23A. Three attenuators connected in tandem.

attenuator of the T-type shown in Fig. 5-23A are given in Fig. 5-23B. Contained in the metal housing are *six individual pads* of 1-, 2-, 3-, 4-, 10-, and 20-dB loss. The total loss is 40 dB adjustable in steps of 1 dB, by connecting the appropriate sections in series or tandem. Terminals 1 and 14 are common to all sections and are connected to the low potential side of the circuit in which the attenuator is used. Other types of attenuators have sections with value of 0.5-, 1.5-, 3-, 6-, and 12-dB loss per section. The resistors are noninductively

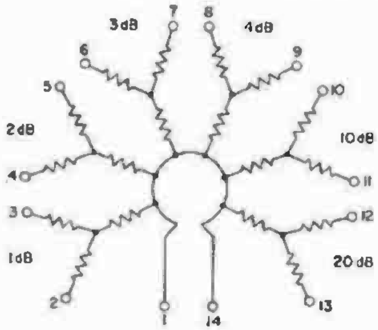


Fig. 5-23B. Internal circuitry for adjustable attenuator having 6 fixed steps of attenuation.

wound, in a metal case that is designed to be mounted with a single screw at the center. They may be obtained for any of the standard impedances.

5.24 ***What is an unbalanced attenuator?***—A configuration with the resistance elements in one side of the line only, as shown in Fig. 5-24.

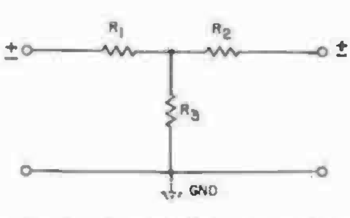


Fig. 5-24. An unbalanced "T" type attenuator.

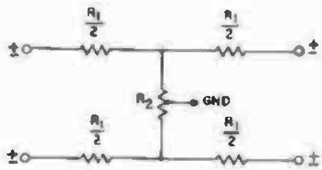


Fig. 5-25. An "H" type or balanced attenuator.

5.25 ***What is a balanced attenuator?***—A configuration with resistance elements in both sides of the line, as shown in Fig. 5-25.

5.26 ***Is it necessary to ground an attenuator?***—If the configuration is unbalanced, the attenuator should be grounded to prevent leakage at the higher frequencies.

5.27 ***Where is a ground connected to an unbalanced pad?***—To the side having no resistance in the line. (See Fig. 5-24.)

5.28 ***Where is a ground connected to a balanced pad?***—To the center of the shunt resistor as shown in Fig. 5-25.

5.29 ***What is the effect if a balanced attenuator is not grounded?***—If the circuits terminating the network are balanced to ground, no difficulty should be encountered. However, if the circuits are unbalanced to ground (unsymmetrical), leakage at the higher frequencies may occur. The amount of leakage may be measured by applying a freqnency of 6000 to 10,000 Hz to the input and observing if the output voltage increases or decreases when the output circuit is reversed. Assuming the transmission characteristics of the circuit are uniform over the above frequency range, no turnover should be noted. Any difference noted with reference to 1000 Hz is caused by leakage. A maximum of 0.25 dB may be tolerated. If possible, a balanced attenuator network should be grounded either by feeding it from, or terminating it by, a balanced to ground circuit, or by connecting a ground directly to the center tap of the shunt resistor.

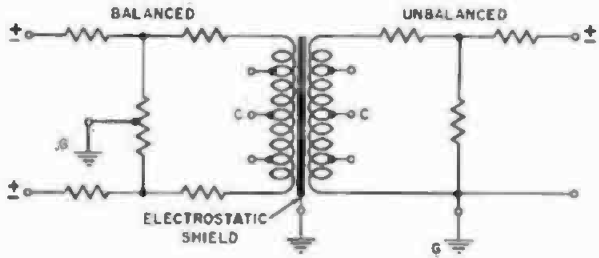


Fig. 5-30A. Connecting balanced and unbalanced networks with a repeat coil.

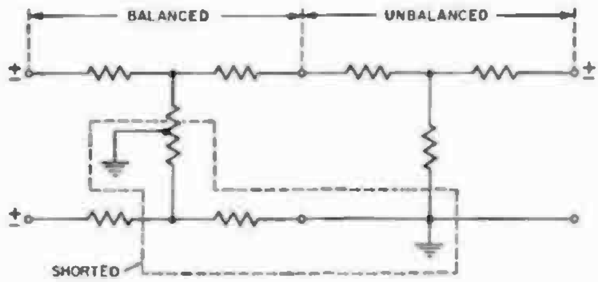


Fig. 5-30B. Two networks connected in tandem.

5.30 ***May a balanced network be directly connected to an unbalanced network?***—Balanced and unbalanced configurations *cannot* be directly connected; however, they may be connected by the use of an *isolation or repeat coil*, Fig. 5-30A. If the networks are not separated electrically, severe instability and leakage at the high frequencies can result. How this occurs is illustrated in Fig. 5-30B. If the networks are connected without the coil, half of the balanced circuit will be shorted to the ground, as indicated by the broken line. The repeat coil will permit the transfer of the audio signal inductively, while separating the grounds of the two networks. Even if the balanced network is not grounded, it should be isolated by a coil. Repeat coils are usually designed for a 1:1 impedance ratio. However, they generally have taps for other impedance ratios.

5.31 ***What is a combining or dividing network?***—A resistive network designed to combine several devices or circuits, each having the same impedance, as shown in Fig. 5-31A. The resistors may be calculated as follows:

$$R_B = \left(\frac{N-1}{N+1}\right) Z$$

where,

R_B is the building-out resistor,
N is the number of circuits fed by the source impedance,
Z is the circuit impedance.

The loss of the network through any two branches is:

$$dB = 20 \operatorname{Log}_{10} (N-1)$$

where,

N is the total number of circuits.

Thus for a three-branch circuit the loss is approximately 6.02 dB.

Unused circuits of a dividing or combining network must be terminated in a resistive load equal to the normal load impedance.

This circuit is often used in the design of sound mixers. This subject is discussed in detail in Section 9. Combining or branching networks may also be designed as a series configuration (Fig. 5-31B). Here are shown three branch circuits combined into one output or input circuit. For equal impedances the equation is:

$$R_t = \left(\frac{N+1}{N-1}\right) Z$$

where,

R_t is the terminating resistor,
N is the number of branch circuits.

The insertion loss may be calculated:

$$dB = 10 \operatorname{Log}_{10} (2N-1)$$

where,

N is the number of branch circuits.

A typical four-circuit combining network, mounted in a metal container, is shown in Fig. 5-31C. The insertion loss of a combining network may be avoided by the use of an *active* combining network, as described in Question 5.99.

5.32 What is a "T" type attenuator?—An attenuator network consisting of three resistors connected in the form

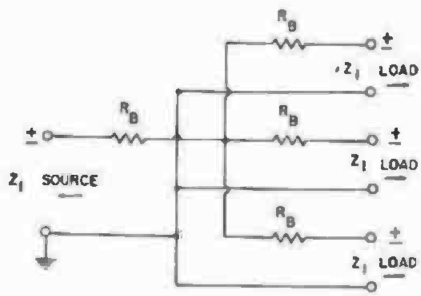


Fig. 5-31A. A combining or dividing network for matching a single circuit to three others.

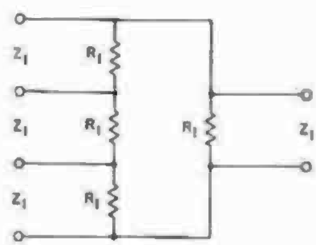


Fig. 5-31B. Series combining network.

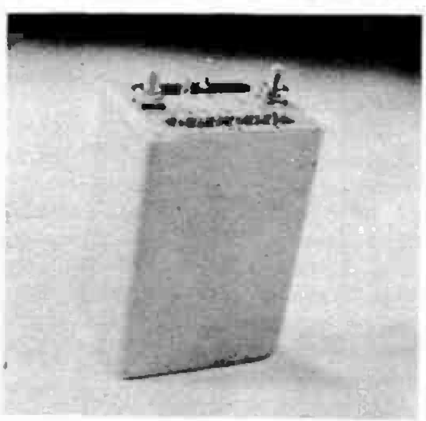


Fig. 5-31C. A four-branch combining network.

of a "T" as shown in Fig. 5-24. The network may be designed to supply an impedance match between circuits of equal or unequal impedance. When designed for use between circuits of unequal impedance, it is often referred to as a taper pad.

5.33 What is the equation used for calculating the element values for a "T" type attenuator, to work between equal impedances?

$$R_1, R_2 = \left(\frac{K-1}{K+1}\right) Z$$

$$R_3 = \left(\frac{K}{K^2-1}\right) 2Z$$

where,

Z is the input and output impedance,
R_1 and R_2 are the series resistors,
R_3 is the shunt arm.

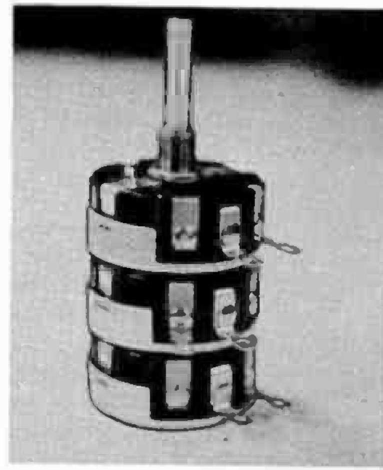


Fig. 5-33. Continuously variable commercial-sound "T" attenuator.

A "T" type attenuator may be designed for any value of loss if designed to operate between equal impedances. The numerical value of the foregoing expressions may be obtained from the table in Fig. 5-22. A commercial-sound, continuously variable attenuator "T" configuration is shown in Fig. 5-33.

5.34 What is a taper pad?—A "T" configuration designed for operation between impedances of unequal value, as shown in Fig. 5-34.

5.35 What is the equation for calculating taper pads?

$$R_1 = Z_1\left(\frac{K^2+1}{K^2-1}\right) - 2\sqrt{Z_1 Z_2}\left(\frac{K}{K^2-1}\right)$$

$$R_2 = Z_2\left(\frac{K^2+1}{K^2-1}\right) - 2\sqrt{Z_1 Z_2}\left(\frac{K}{K^2-1}\right)$$

$$R_3 = 2\sqrt{Z_1 Z_2}\left(\frac{K}{K^2-1}\right)$$

where,

Z_1 is the larger of the two impedances.

The numerical values for these expressions may be taken from the table in Fig. 5-22. Thus, for a network to match 500 ohms to a circuit of 250 ohms with a loss of 20 dB, the resistor values are:

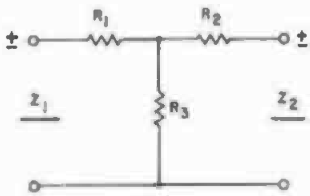


Fig. 5-34. A taper pad used for matching two circuits of unequal impedance.

$$\begin{aligned} R_1 &= 500\,(1.0202) - 2\,(\sqrt{125{,}000})\,(0.10101) \\ &= 510.1 - 2\,(353.55)\,(0.10101) \\ &= 510.1 - 71.42 \\ &= 438.6\Omega \end{aligned}$$

$$\begin{aligned} R_2 &= 250\,(1.0202) - 2\,(\sqrt{125{,}000})\,(0.10101) \\ &= 255.05 - 2\,(353.55)\,(0.10101) \\ &= 255.05 - 71.42 \\ &= 183.63\Omega \end{aligned}$$

$$\begin{aligned} R_3 &= 2\,(\sqrt{125{,}000})\,(0.10101) \\ &= 2\,(353.55)\,(0.10101) \\ &= 71.42\Omega \end{aligned}$$

5.36 ***What is a bridged "T" attenuator?***—An attenuator network containing four resistive elements, as shown in Fig. 5-36. The resistors R_1 are equal in value to the line impedance; therefore, they require no calculation. This network is designed to work between impedances of equal value only, and is the configuration most commonly employed in sound-mixer controls because the configuration requires only two rows of contacts, and two variable resistor groups. The contact arms for resistors R_5 and R_6 are connected mechanically by a common shaft and vary inversely in value with respect to each other.

5.37 ***What is the equation for calculating a bridged-"T" attenuator?***

$$R_1 = Z$$

$$R_5 = (K - 1)Z$$

$$R_6 = \left(\frac{1}{K - 1}\right)Z$$

where,

Z is the line impedance,
R_5 is the bridging resistor,
R_6 is the shunt resistor.

Again the factors may be taken from the tabulations of Fig. 5-22.

5.38 ***What is a balanced bridged-"T" attenuator?***—A configuration similar to the unbalanced bridged-"T" attenuator, except the resistor elements are divided and placed in each side of the line as shown in Fig. 5-38. The principal objection to the use of this configuration, if made variable, is that the shunt resistor R_6 must be divided into two separate arms to provide a ground

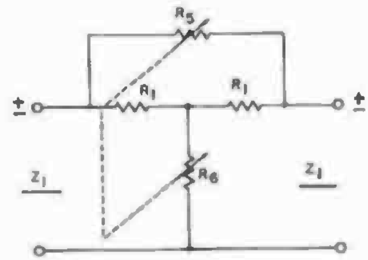


Fig. 5-36. A bridged-"T" attenuator. For variable pads, the arms R_5 and R_6 are made variable.

connection at the exact electrical center. However, if the circuit feeding or terminating the attenuator is balanced to the ground, the ground connection at the attenuator center will not be required.

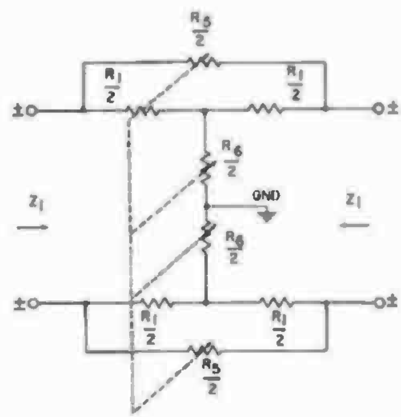


Fig. 5-38. A balanced bridged-"T" attenuator. For a variable configuration four variable arms are required.

5.39 ***What is an "H"-type attenuator?***—A balanced "T" pad. The pad is first calculated as an unbalanced "T" configuration described in Question 5.33. The series resistance elements are then divided and one-half connected in each side of the line as shown in Fig. 5-39. The shunt resistor remains the

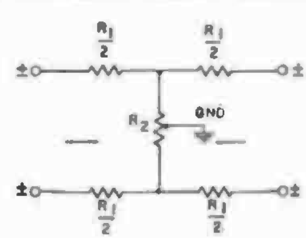


Fig. 5-39. An "H"-type attenuator, also called a balanced "T."

same value as for the unbalanced configuration. A tap is placed at the exact electrical center of the shunt resistor for connection to ground.

5.40 *What is an "L"-type attenuator?*—A configuration consisting of two resistive elements connected in the form of an "L" shown in (a) to (d) of Fig. 5-40. This pad does not reflect the same impedance in both directions. An impedance match is afforded only in the direction of the arrow shown in the figures. If an "L"-type network is employed in a circuit which is sensitive to impedance match, the circuit characteristics may be affected. An "L"-type network should not be used, except where a minimum loss is required and a network of the "T" configuration will not serve because of its minimum loss.

5.41 *What are the equations used for the design of "L"-type attenuators?*—An "L"-type network can only be designed to supply an actual impedance match in a given direction. For unequal impedances, the impedance match may be in the direction of the larger or the smaller impedance but not both.

When an "L"-type network is connected to circuits of equal impedance, the circuits are matched in one direction but not in the other. However, the network may be designed to match either of the terminating impedances.

The arrows in Fig. 5-40 indicate the direction of impedance match. If the network is designed to match the impedance in the direction of the series arm, the mismatch is toward the shunt arm. The mismatch increases with the increase of loss, and at high values of attenuation the value of the shunt resistor may become a fraction of an ohm, which can have a serious effect on the circuit to which it is connected.

The configuration for an "L"-type network operating between impedances of unequal value, Z_1 and Z_2, is shown at (a) in Fig. 5-40. The impedance match is toward the larger of the two impedances Z_1, the values of the resistors are:

$$R_1 = \left(\frac{Z_1}{S}\right)\left(\frac{KS-1}{K}\right)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{1}{K-S}\right)$$

where,

S equals $\sqrt{\frac{Z_1}{Z_2}}$.

For a condition where the impedances are equal, and the impedance match is in the direction of the arrows, as in (b) of Fig. 5-40, the values of the resistors may be calculated by the formula:

$$R_1 = Z\left(\frac{K-1}{K}\right)$$

$$R_2 = Z\left(\frac{1}{K-1}\right).$$

For a condition where the impedances are unequal and the impedance match is toward the smaller of the two impedances, as in (c) of Fig. 5-40, the values of the resistors are determined by the formula:

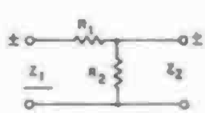
$$R_1 = \left(\frac{Z_1}{S}\right)(K-S)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{K}{KS-1}\right)$$

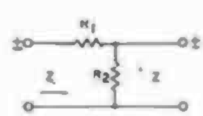
where,

S equals $\sqrt{\frac{Z_1}{Z_2}}$.

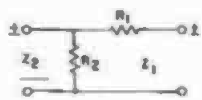
For the conditions shown in (d) of Fig. 5-40, resistors R_1 and R_2 may be calculated by the formula:



(a) Between impedances of unequal value.



(b) Between impedances of equal value.



(c) Impedances unequal, and impedance match is toward the smaller of the two.

(d) Between impedances of equal value, in the direction of the shunt arm.

Fig. 5-40. Configurations that are used for L-type networks.

$$R_1 = Z\,(K-1)$$

$$R_2 = Z\left(\frac{K}{K-1}\right)$$

5.42 What is a pi or delta-type attenuator?—A resistive network resembling the Greek letter pi, or delta, as shown iu Fig. 5-42. Actually there is no difference between the pi or delta configuration, only the manner in which they are drawn. Such networks may be used between impedances of equal or unequal values.

5.43 What are the equations used for designing pi-type attenuators?—For networks operating between impedances of eqnal value:

$$R_1 = Z\left(\frac{K+1}{K-1}\right)$$

$$R_2 = \left(\frac{Z}{2}\right)\left(\frac{K^2-1}{K}\right)$$

where,

R_1 is the input and ontput resistor,
R_2 is the series resistor,
Z is the input and output impedance.

5.44 What are the equations for designing a pi-type attenuator to operate between unequal impedances?

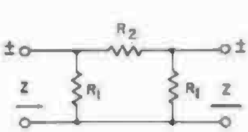
$$R_1 = Z_1\left(\frac{K^2-1}{K^2-2KS+1}\right)$$

$$R_2 = \left(\frac{\sqrt{Z_1 Z_2}}{2}\right)\left(\frac{K^2-1}{K}\right)$$

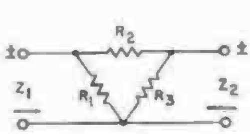
$$R_3 = Z_2\left(\frac{K^2-1}{K^2-2\frac{K}{S}+1}\right)$$

where,

R_1 and R_3 are the shunt resistors,
R_2 is the series resistor,
Z_1 is the input impedance,
Z_2 is the output impedance,
S equals $\sqrt{\frac{Z_1}{Z_2}}$.



(a) Between impedances of equal value.

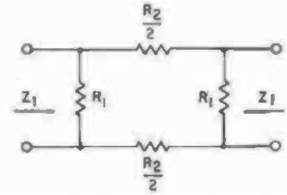


(b) Between impedances of unequal value.

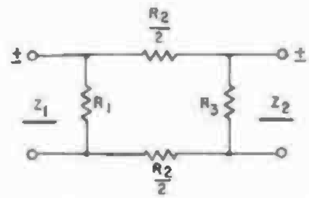
Fig. 5-42. Pi- or delta-type attenuator networks.

To simplify the calculations, the values of the factors may be selected from the table in Fig. 5-22.

5.45 What is an "O"-type attenuator?—A balanced pi-type attenuator configuration. The circuit element values may be obtained by first calculating



(a) Between impedances of equal value.



(b) Between impedances of unequal value.

Fig. 5-45. The "O"-type attenuator.

for a pi-type configuration, described in Question 5.43, then dividing the series resistor and placing half in each side of the line as shown in Fig. 5-45. The shunt resistors remain the same value.

5.46 What is a "U"-type attenuator and where is it used?—"U"-type attenuators may be of a symmetrical or balanced-type configuration, and are quite handy around the laboratory for

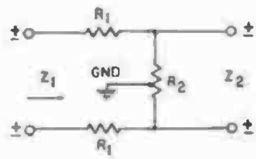


Fig. 5-46. A "U" pad for matching a high impedance circuit to one of very low impedance.

matching a high value of impedance to one of low value. A typical requirement is that of matching the 600-ohm send circuit of a gain set to a microphone preamplifier having a 30-ohm input impedance (Fig. 5-46).

Here the impedance match is of first importance, the loss being secondary. For a symmetrical configuration to work between unequal impedances the resistors may be calculated as follows:

$$R_1 = \left(\frac{Z_1}{2S}\right)\left(\frac{KS-1}{k}\right)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{1}{k-S}\right)$$

where,
- R_1 is the series resistor,
- R_2 is the shunt resistor,
- Z_1 is the larger impedance,
- Z_2 is the smaller impedance,

S equals $\sqrt{\frac{Z_1}{Z_2}}$.

5.47 What is the equation for matching the circuit impedance to the shunt arm of a "U"-type attenuator?—The value of the resistors for the configuration shown in Fig. 5-47 may be calculated as follows:

$$R_1 = \left(\frac{Z_1}{2S}\right)(K-S)$$

$$R_2 = \left(\frac{Z_1}{S}\right)\left(\frac{K}{KS-1}\right)$$

S equals $\sqrt{\frac{Z_1}{Z_2}}$.

The arrow indicates the direction of the impedance match. Any "U" pad may be balanced to ground by connecting a ground to the electrical center of the shunt resistor.

5.48 What is a lattice-type attenuator?—A balanced configuration as shown in Fig. 5-48. Such pads are designed to operate between circuits of equal impedance. They are used in the telephone industry, but are seldom used in normal audio circuits.

5.49 What is the equation for designing a lattice-type attenuator?—The lattice-type attenuator is a truly balanced configuration. Therefore, individual treatment is required for solving the circuit element values. If used in an unbalanced configuration, it degenerates into a pi configuration. The difference between this configuration and the "O" network is that shunt resistors R_2

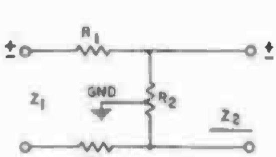
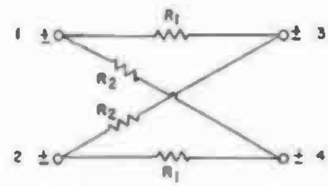
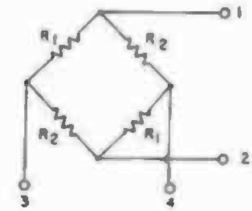


Fig. 5-47. A "U" pad configuration for operation between impedances of unequal value.



(a) *Lattice configuration.*



(b) *Equivalent circuit (bridge).*

Fig. 5-48. A lattice-type attenuator and equivalent bridge circuit.

are *crossed-over* and connected to both the input and output circuits (Fig. 5-48A). The circuit elements may be calculated:

$$R_1 = \left(\frac{K-1}{K+1}\right)Z$$

$$R_2 = \left(\frac{K+1}{K-1}\right)Z$$

where,
- R_1 is the series resistor,
- R_2 is the shunt resistor.
- Z is the line impedance.

Lattice networks can only be employed in ungrounded circuits of *equal* impedance. If an equivalent circuit is drawn for this pad (Fig. 5-48B), it resembles a balanced-bridge circuit. The principal difference is that resistors R_1 are lower in value than R_2. This difference unbalances the bridge and permits the signal to pass unattenuated in an amount controlled by the ratio of resistors R_1 to R_2.

5.50 Show graphically the loss in level because of impedance mismatch.—If two resistive networks are mismatched, generally the frequency characteristics are not affected; only a loss in level occurs. Attenuators, when properly constructed, consist of pure resistance and are nonreactive; therefore, they present a constant impedance or resistance. If the impedance mismatch ratio is known, the loss in level may be directly read from the graph in Fig. 5-50 (see Question 5.11).

5.51 What is a ladder-type attenuator?—A configuration used for sound

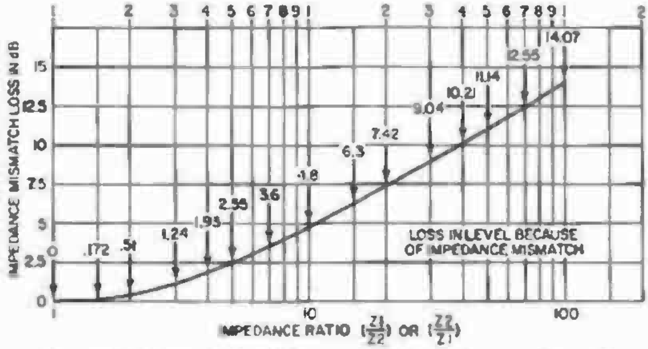


Fig. 5-50. Loss in level because of impedance mismatch.

mixer controls as shown in Fig. 5-51A. Its name is derived from the fact that the configuration is similar to a ladder laid on its side. The configuration of a ladder pot consists of a group of pi-pads connected in tandem. One of the disadvantages of this configuration is the fact that the input and output impedances are not constant throughout its complete range of attenuation. Also, ladder configurations have a fixed insertion loss of 6 dB which is exclusive of the variable setting loss. This loss must be taken into consideration when designing mixer networks.

Ladder pots for mixer-control use, may be obtained in two types of construction—slide-wire and contact types. Ladder pots are designed to operate between circuits of equal impedance.

For motion picture rerecording mixers, the slide-wire type control is generally employed, because it permits a smooth, even attenuation over a wide range. The contact type although not quite as smooth in operation as the slide-wire has only one row of contacts, which reduces the noise and maintenance.

Ladder networks may be designed for balanced operation, also. This is accomplished by connecting two unbal-

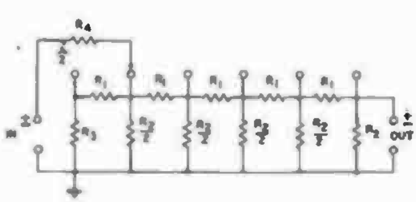


Fig. 5-51A. Configuration for a ladder attenuator for mixer controls. Variable in fixed steps of loss.

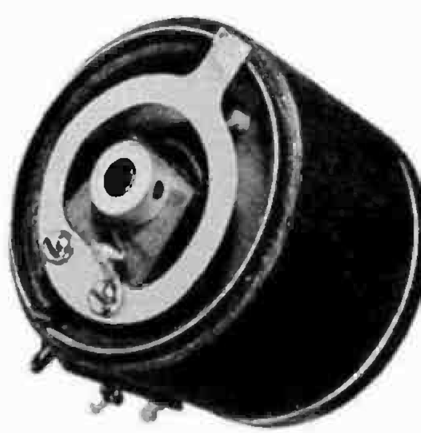


Fig. 5-51C. Unbalanced slide-wire ladder attenuator used in sound mixers.

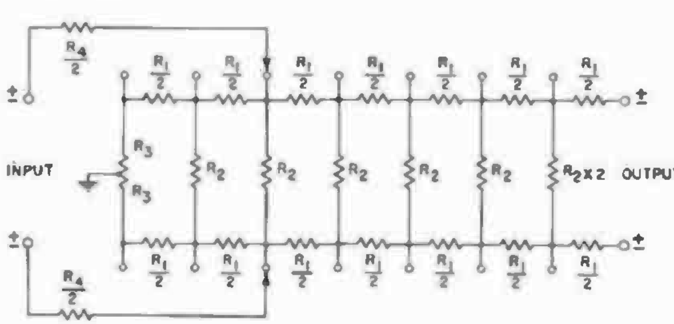


Fig. 5-51B. Balanced ladder network.

anced networks side-by-side (Fig. 5-51B). However, it will be observed, the circuit elements are *not* divided in the same manner as for other type balanced networks. If an unbalanced ladder network is compared with a balanced ladder network, it will be observed that resistors R_1 are divided by *two*, resistors R_1 are also divided by two, and at the output R_4 is now *twice* the value for the unbalanced configuration. Resistor R_3 remains at its *original* value on each side of ground. A typical slide-wire unbalanced ladder attenuator is shown in Fig. 5-51C. The impedance characteristic for a single nnit is given in Fig. 5-83A.

5.52 What is the equation used for calculating a ladder-type attenuator?

$$R_1 = \left(\frac{K^2 - 1}{2K}\right) Z$$

$$R_2 = \left(\frac{K + 1}{K - 1}\right) Z$$

$$R_3 = \frac{R_2 \times Z}{R_3 + Z}$$

$$R_4 = \frac{Z}{2}$$

$$Z_{in} = Z_{out}$$

where,

R_1 is the series resistance,
R_2 is the shunt resistance,
R_3 is the input shunt resistor,
R_4 is the series resistance in the contact arm circuit.

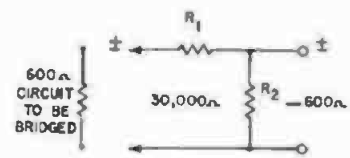
The value of "K" is dependent on the loss per step—not the total loss.

5.53 What is a bridging attenuator? —A pad designed for bridging (paralleling) one circuit while feeding another without disturbing the impedance or frequency characteristics of the circuit being bridged, and absorbing only a small amount of power.

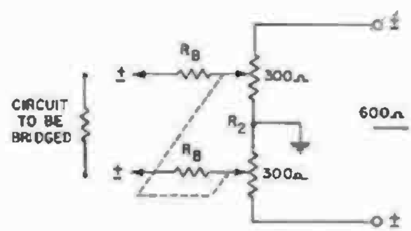
Two such pads are shown in Fig. 5-53, one fixed and the other variable. At (a) is a fixed-loss bridging pad with an input resistance of 30,000 ohms, designed to be terminated in 600 ohms. A pad of this design would have an approximate loss of 40 dB. The resistor R_1 may be changed to any value with a subsequent change of loss. The configuration may be changed to a balanced one by dividing R_1 and placing half in each side of the line.

The pad at (b) may be designed for any bridging resistance by the proper selection of the series resistors R_B, each resistor being one-half the desired bridging resistance. The output resistor R_2 is equal to the impedance of the device across the output terminals of the attenuator. For this illustration, this value is 600 ohms. The two pots in series should total the value of R_2 or, in some instances they can be greater than R_2.

Balanced pads of this design require a ground as shown to prevent leakage at the higher frequencies. The ground to the attenuator is connected to the amplifier ground to prevent the formation of a ground loop.



(a) *Unbalanced (fixed).*



(b) *Balanced (variable).*

Fig. 5-53. Bridging pads.

5.54 What are the resistor values for the more commonly used bridging attenuators?—For the configuration shown at (a) of Fig. 5-53, the following will apply:

R_1	*Loss in dB*	*R_2*
30,000Ω	40.1	600Ω
25,000Ω	38.5	600Ω
20,000Ω	36.6	600Ω
15,000Ω	34.2	600Ω
10,000Ω	30.7	600Ω
7500Ω	28.3	600Ω
5000Ω	24.9	600Ω
3000Ω	20.8	600Ω

The above networks will have the indicated loss only when bridging a circuit of 600 ohms impedance and when terminated in 600 ohms.

The loss for the configuration shown at (a) of Fig. 5-53 may be calculated as follows:

$$dB = 10 \operatorname{Log}_{10} \frac{Z_1}{Z_2} + 20 \operatorname{Log}_{10} \sqrt{\frac{Z_1}{Z_2}} + \sqrt{\frac{Z_1}{Z_2} - 1}$$

where,

Z_1 is the bridging resistance of the pad,

Z_2 is the input impedance of the device to which the pad is connected.

5.55 ***How may a bridging pad be constructed without regard to loss?***—By using the configuration in (a) of Fig. 5-53. Resistor R_1 in the series arm equals the value of the desired bridging input. The shunt resistor R_2 is selected for a value that will properly terminate the input of the device being fed by the pad. The loss may be calculated using the equation given in Question 5.54.

5.56 ***What is the ratio generally employed for bridging pads?***—The bridging impedance should be at least 10 times that of the circuit to be bridged. Standard bridging impedances are: 30,000, 25,000, 10,000, and 7500 ohms.

5.57 ***What is a minimum-loss pad?***—A pad designed to match circuits of unequal impedance with a minimum loss in the matching network. Attenuators designed to work between impedances of unequal value have a minimum loss. This minimum loss is dependent on the ratio of the terminating impedances. As an example: the minimum loss for a pad with an impedance ratio of two is 7.8 dB. Therefore, an attenuator designed to work between 500 and 250 ohms will have a minimum loss of 7.8 dB. The pad may be designed for a greater loss but never less. The minimum loss for various impedance ratios may be determined from the minimum loss chart in Fig. 5-21. Pads designed to work between equal impedances may be designed for any value of loss.

5.58 ***What is the equation used for designing a minimum loss attenuator for matching two impedances of unequal value?***—Using the "L" configuration shown in Fig. 5-58, the resistor values are:

$$R_1 = \sqrt{Z_1 (Z_1 - Z_2)}$$

$$R_2 = \frac{Z_1 Z_2}{R_1}$$

where,

R_1 is the series resistor connected in the side of the larger impedance,

R_2 is the shunt resistor.

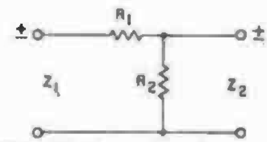


Fig. 5-58. Configuration for a minimum loss attenuator to match two unequal impedances. Match is in the direction of the arrow.

5.59 ***What is the equation used for designing a minimum loss attenuator when only one impedance is to be matched?***—If the larger impedance Z_1 is to be matched, only a series resistor "R" is used as shown in Fig. 5-59. The value of this resistor is:

$$R = Z_1 - Z_2$$

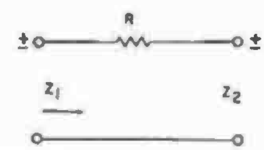


Fig. 5-59. Configuration for a minimum loss attenuator where only the larger impedance (Z) is to be matched. Match is in the direction of the arrow.

5.60 ***What is the equation used for designing a minimum loss attenuator when only the smaller impedance is to be matched?***—In this instance only one resistor is used, connected in shunt with the smaller impedance Z_2, as shown in Fig. 5-60. Its value is:

$$R = \frac{Z_1 Z_2}{Z_1 - Z_2}$$

5.61 ***What are the equations for calculating the loss of a minimum loss attenuator?***—For the "L" type described in Question 5.58:

$$dB = 20 \operatorname{Log}_{10} \sqrt{\frac{Z_1}{Z_2}} + \sqrt{\frac{Z_1}{Z_2} - 1}$$

For Questions 5.59 and 5.60:

$$dB = 20 \operatorname{Log}_{10} \sqrt{\frac{Z_1}{Z_2}}$$

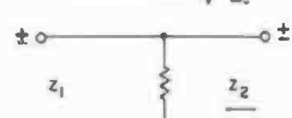


Fig. 5-60. Configuration for a minimum loss attenuator where only the small impedance (Z) is to be matched. Match is in the direction of the arrow.

5.62 What are the resistor element values for the more commonly used minimum loss attenuators?—Values for an "L"-type configuration are as follows:

Z_1/Z_2	R_1	R_2	dB loss
600/500	245	1224	3.8
600/250	458.2	326.8	8.8
600/150	519.6	173.3	11.4
600/30	584.8	30.8	18.9
500/250	253.6	353.6	7.8
500/30	484.8	30.9	18.1
250/30	234.4	32	14.8

This configuration may be converted to a balanced configuration by dividing the series resistor and placing half the value of its resistance in each side of the line.

5.63 What is a grid potentiometer (gain control)? — A potentiometer (pot for short) connected in the control-grid circuit of a vacuum tube for the purpose of controlling the amplification or gain of the stage. The arm of the potentiometer is connected to the control grid of the tube, as shown in Fig. 5-63.

The loss of the pot may be calibrated in decibels by the use of the equation:

$$R_1 = \left(\frac{1}{K}\right) Z$$

$$R = (Z - R_1)$$

where,

- Z is the total resistance of the pot,
- R_1 and R_2 are the upper and lower sections of the resistance, depending on the position of the contact arm.
- K equals a given value of loss in decibels taken from the tables in Fig. 5-22.

The steps may be fixed or continuously variable.

5.64 How may a grid potentiometer be calibrated in voltage?

$$E_2 = E_1 \left(\frac{R_1}{R_1 \times R_2}\right)$$

where,

- E_1 is the input voltage,
- E_2 is the output voltage,
- R_1 and R_2 are the upper and lower sections of the resistance. (See Fig. 5-64.)

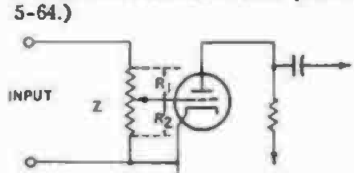


Fig. 5-63. A grid pot.

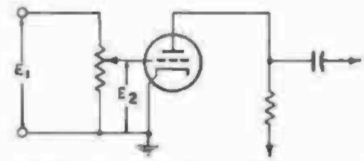


Fig. 5-64. Symbols for calculating the voltage ratio of a grid pot (gain control).

5.65 What is a loudness control?—A ladder-type frequency-compensated gain control consisting of a group of RC circuits. The purpose of this control is to compensate for the human ear characteristic when the sound level is increased or described in a sound reproducing system. The frequency compensation of the control is based on the well-known Fletcher-Munson equal-loudness contours as shown in Fig. 1-76. The configuration of two such controls is shown in Figs. 5-65A and 5-65B. The frequency characteristic of the second control is shown in Fig. 5-65C.

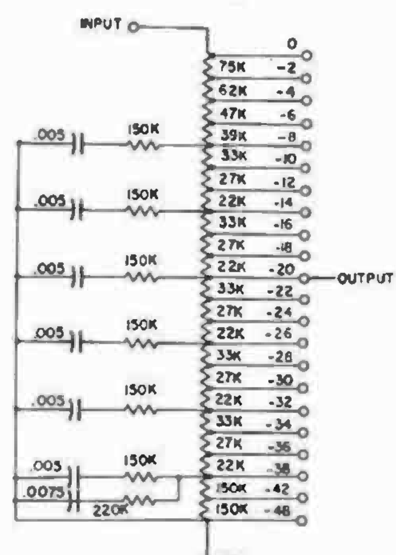


Fig. 5-65A. Configuration for a ladder-type loudness control.

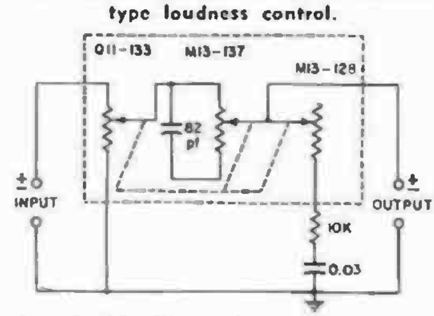


Fig. 5-65B. The IRC loudness control Model LC-1.

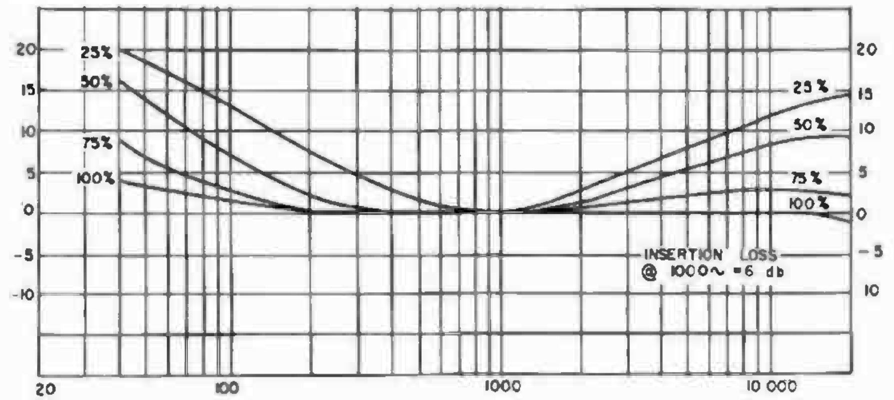


Fig. 5-65C. Frequency characteristics of the IRC loudness control Model LC-1.

5.66 ***Is it practical to construct pads with losses of 60 dB or greater?***—No. It is not good practice to build pads of over 40 dB loss, unless special precautions are taken to reduce the distributed capacity and leakage between the input and output sections. It is more practical to build two or more pads of lower loss and connect them in tandem. The total loss is the sum of the individual losses, assuming that all impedence matches are satisfied between sections.

5.67 ***How may the resistance of an attenuator be measured with an ohmmeter?***—Terminate the output with a resistance equal to the terminating impedance. Measure the input resistance with an ohmmeter. The resistance as measured by the ohmmeter should equal the impedance of the pad. If the attenuator is variable, the dc resistance should be the same for all steps.

5.68 ***If the resistance values of an attenuator are known, how may they be converted to a different impedance?***

$$R_1 = K \times R$$

$$K = \frac{Z_1}{Z}$$

where,
- Z_1 is the new impedance,
- Z is the known impedance,
- K is a multiplying factor,
- R is the known value of resistance,
- R_1 is the new value of resistance.

(See Questions 6.67 and 7.62.)

5.69 ***What is the minimum loss for pads designed to work between equal impedances?***—Pads designed to work between equal impedances may be designed for any value of loss.

5.70 ***What is a compensated L-type attenuator?***—An "L" pad with fixed steps. Each step has a series building-out resistor as shown in Fig. 5-70. The purpose of the resistors in series with the contact arm is to present a constant impedance in both directions. If a constant impedance is required a straight "T" or bridged "T" is recommended.

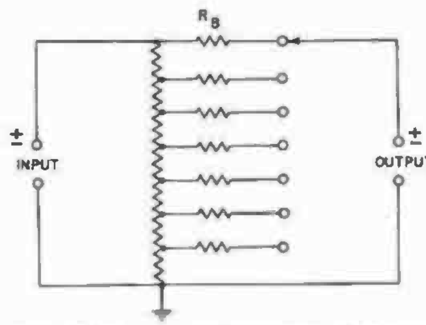


Fig. 5-70. A compensated "L"-type attenuator pad.

5.71 ***How may the impedance of an unknown attenuator be determined?***—By first measuring the resistance looking into one end with the far end open, and then shorted. The impedance is the geometric mean of the two readings:

$$Z = \sqrt{Z_1 \times Z_2}$$

where,
- Z_1 is the resistance measured with the far end open,
- Z_2 is the resistance measured with the far end shorted.

This measurement will hold true only for pads designed to be operated between equal terminations. If the dc resistance of the two ends differs, it may be assumed that the pad is designed to be operated between unequal impedances.

5.72 ***What precautions should be observed when installing attenuators?***—

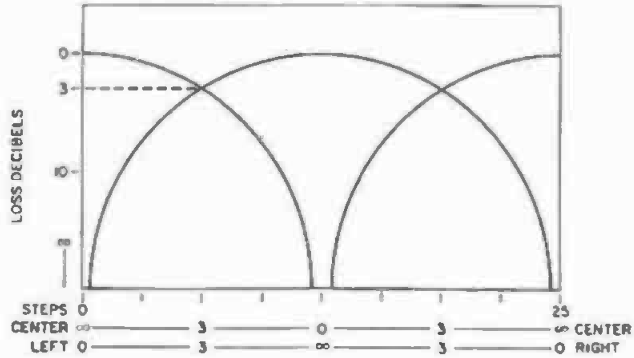


Fig. 5-73B. Characteristics for a three-channel pan-pot.

The input and output circuits must be separated from each other, well shielded, and grounded to prevent leakage at the higher frequencies. As an example: An attenuator of 40 dB loss has a signal voltage reduction of 100:1 between the input and output terminals. Therefore, if coupling between the input and output circuits is permitted, serious leakage can occur at frequencies above 1000 Hz.

5.73 ***Describe a pan-pot (panoramic control).***—Panoramic controls, or pan-pots as they are commonly known, are used in stereophonic rerecording for transferring the *apparent* position of the sound source from one section of a sound field to another. The arrival time of a sound will influence the apparent position of its source, and the intensity offsets the effects of the arrival time. A change of only 3 dB in intensity is enough to displace the apparent source across the sound field. Its effect is controlled by the use of a pan-pot.

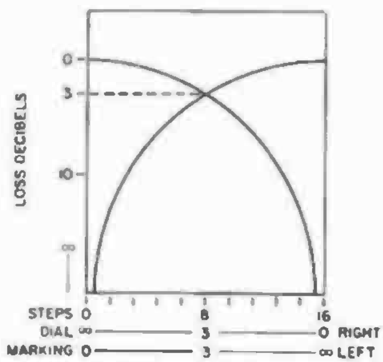


Fig. 5-73A. Characteristics for a two-channel pan-pot.

Either a two- or three-channel pan-pot may be used to pick up a monophonic sound or transfer it to any geometric position desired in the final stereophonic rerecording. Pan-pots designed for stereophonic usage differ from the conventional mixer control in that for a two-channel system, two oppositely wound controls are ganged to-

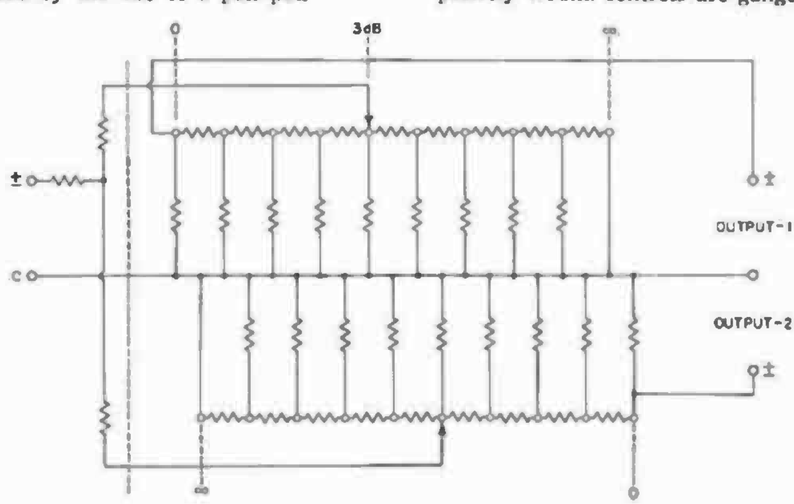


Fig. 5-73C. Two-channel ladder configuration pan-pot for stereophonic recording.

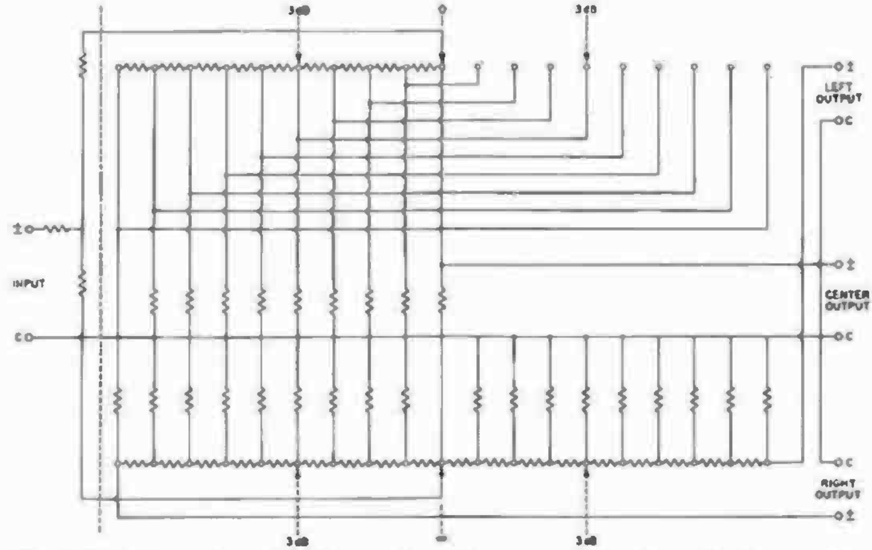


Fig. 5-73D. Three-channel ladder configuration pan-pot for stereophonic recording.

gether in such a manner that the 3-dB down point of each control occurs at zero degrees (Fig. 5-73A). For a three-channel control (Fig. 5-73B), they are so designed that the 3-dB down point occurs at 45 degrees with reference to either side of center. At either of the 90-degree positions, the extreme opposite channel is at infinite attenuation. The usual taper required for such attenuation rate requires at least 16 positions in each control unit to attain the smoothest transition throughout the audible range of control. The attenuation ratio at the extremes is rapid, but slow in the range of overlap from one section to another. The attenuation rate must be precise and conform to the mathematics governing the angular displacement of the sound field with changes of sound level.

Pan-pots lend themselves easily to the ladder configuration, as only two rows of contacts are required for either two- or three-channel control, contrasted to four and six rows for a "T"-type control. The insertion loss is 12 dB for either type. The schematic diagram for a one channel into two is shown at Fig. 5-73C, and a one into three channels in Fig. 5-73D. (See Question 9.22.)

5.74 ***What is a divergence control?*** —A variable network used in stereophonic rerecording for controlling the divergence of the sound field projected by a panometric control over three or more odd-numbered channels. In its narrow position (minimum divergence), a point source of sound is effected. This point lies on the axial position of the sound field under control.

5.75 ***What is a panometric control?*** —A variable network used in stereophonic rerecording to direct a segment of a sound field over a wider field. It is used in conjunction with a divergence control so that the effect may be shaded from a three-speaker width, in a group of five to seven speakers, to a single point sound source. Its effect is similar to a Zoomar optical lens used on motion picture and television cameras.

5.76 ***What is an off-screen fader control?***—A stereophonic control used in rerecording. A single sound source can be made to arrive from a distance at the left off-screen and ride across to the right side of the screen (or vice versa) and then fade away into the distance.

5.77 ***What is a leakage control?***— A control used in stereophonic projection with a built-in leakage to permit a controlled amount of leakage between channels. The leakage is designed for about 10 dB below the normal level on adjacent channels, and 21 dB on second removed channels. This controlled leakage permits the audience in the front rows and side seats of a theater to properly hear the program material.

5.78 ***What is the Hawkins effect?***— J. N. A. Hawkins when exploring the field of stereophonics in 1938 during the development of Fantasound for the Walt

Disney production "Fantasia" observed that if two loudspeakers are energized with the same program material with a level difference of 7 dB, the acoustic level will drop off only 1 dB when the lower-level speaker is cut off.

5.79 *How are mixer controls constructed?*—Mixer controls used for production or rerecording are generally of the ladder or bridged-T configuration. In the bridged-T type, *two* rows of contacts are required, ranging between 20 and 45 per row, in steps of 0.5 dB to 2 dB per step. For ladder configurations only *one* row of contacts is required, and they generally range from 1 dB to 2 dB per step. For the slide-wire ladder control no contacts are used, the loss being continuously variable by the action of the swinger arm on a resistance strip. Straight-T attenuators are rarely used in mixer circuits because three rows of contacts are required to achieve the same amount of attenuation per step. The resistors are generally wound using a bifilar type winding, as discussed in Question 5.16. The important things to look for in a mixer control are internal noise, contact material, ease of operation, and compatibility of the configuration with the design of the mixer network. (See Questions 5.51 and 5.52.)

5.80 *What type mixer controls are recommended for rerecording consoles?*—Mixer controls used for rerecording require a greater number of steps and attenuation to permit a smoother and greater range of control. As a rule, a slide-wire ladder type control will be found the most satisfactory, because of its continuous attenuation without steps (Fig. 5-51C). The next choice is a ladder configuration, using 45 contacts of 1-dB attenuation per step. A third choice is a bridged-T with 45 steps of 0.5-dB attenuation per step. All the above types have rapid cutoff in the last three steps to infinity. It should be remembered the ladder configuration has a *fixed* insertion loss of 6 dB, while the bridged-T has none. If the gain following the mixer network is sufficient, the additional 6-dB loss of the ladder control is of no consequence. This is discussed in detail in Section 9.

There has been over the years a continuing discussion as to the relative merits of the rotary and straight-line attenuators. The selection of a mixer control depends principally on the size of the console and the type rerecording to be done. For small consoles having four to six inputs, the rotary control may be quite satisfactory.

For stereophonic mixer consoles, only straight-line (vertical) controls are used because of space conservation and the necessity of operating several controls simultaneously. It is not uncommon in stereophonic consoles to employ 28 to 34 straight-line controls, plus variable equalizers, filters, and other devices.

Insofar as the electrical characteristics are concerned, the configurations are the same for both the rotary and straight-line controls. Straight-line attenuators lend themselves to simultaneous operation better than the rotary type, although many mixers have mastered the technique of moving two and even three rotary controls with one

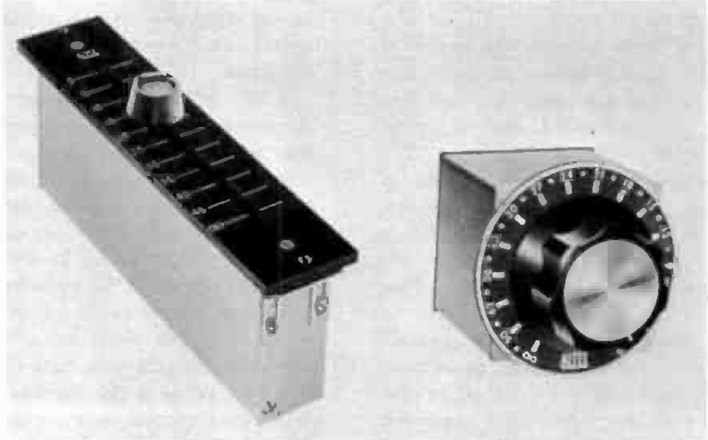


Fig. 5-80. Altec-Lansing straight-line and rotary mixer controls.

Fig. 5-81. Attenuation versus percent rotation for a typical mixer control.

hand. In the straight-line type, three or four can be moved with one hand. For stereophonic rerecording, one straight-line control case may contain up to four individual controls, operated from a single knob at the top. The interior of straight-line multicontrol designs are also shown in Section 9. (See Question 5.79.)

5.81 ***What is the attenuation characteristic used in a typical mixer control?***—The attenuation versus the percent rotation for a typical variable attenuator is shown in Fig. 5-81. It will be noted the attenuation increases quite rapidly after 80 percent of the dial movement is reached. This permits a rapid fade or cutoff.

5.82 ***What are the impedance characteristics of a variable bridged-T attenuator?***—The impedance variations for a typical high quality attenuator

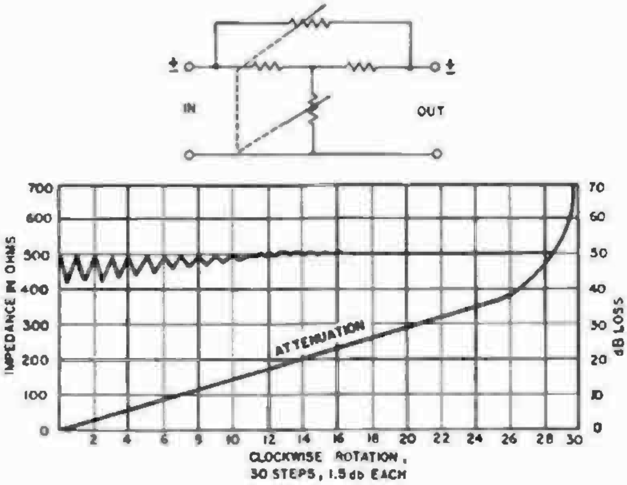


Fig. 5-82. Impedance characteristics for a high-quality 500-ohm variable bridged-"T" mixer control.

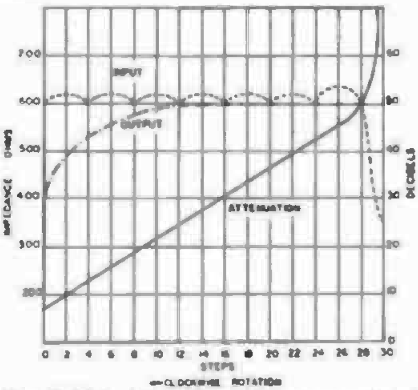


Fig. 5-83A. Impedance characteristics of a 600 ohm ladder-type mixer control.

used in a mixer network are shown in Fig. 5-82. It will be noted the greatest impedance variation occurs as the attenuator arm approaches zero attenuation and amounts to about 80 ohms, reducing the overall impedance to 420 ohms. However, this impedance variation is not too serious, as the mixer-combining network with its building-out resistors isolates this variation to a great extent from associated attenuators. This subject is discussed further in Question 9.24.

5.83 *What are the impedance characteristics for a ladder mixer control?*—This type attenuator will show impedance variations at both the input and output and between steps. However, when used in a combining network with the proper building-out resistors, these variations are of little consequence. A typical impedance curve is shown in Fig. 5-83A.

A plot of the impedance characteristics of a typical four-position mixer network employing four *unbalanced* ladder attenuators is shown in Fig. 5-83B. It will be observed that the principal mismatch occurs at the last three steps, where the attenuation is the least. As a rule, the working range for a typical mixer control is around the 15 to 20-dB loss point. Thus, very little or no mismatch occurs in the operating range.

5.84 *What is a building-out network?*—A resistive combination or combining network used in sound mixers for the purpose of combining the mixer controls in such a manner that the correct impedance match is maintained between the attenuators. Combining networks are also discussed in Question 9.24.

5.85 *What is the cause of noise in a variable attenuator?*—The average noise level for a "T"-type mixer control is on the order of minus 100 dB and is constant; therefore, the signal-to-noise ratio varies with the loss setting. The noise level for a ladder-type attenuator is on the order of minus 120 dB and as the attenuation is increased, the signal-to-noise ratio is *increased*. In high-level mixing circuits, where the mixer control follows a preamplifier, the 6-dB fixed insertion loss of a ladder attenuator increases the signal-to-noise ratio by 6 dB. Also, if the preamplifier employs negative feedback, its internal output impedance will generally be *less* than the impedance of the attenuator. It is not uncommon for a 600-ohm preamplifier to have an internal output impedance of 90 ohms; in this instance building-out resistors should be employed to secure proper impedance

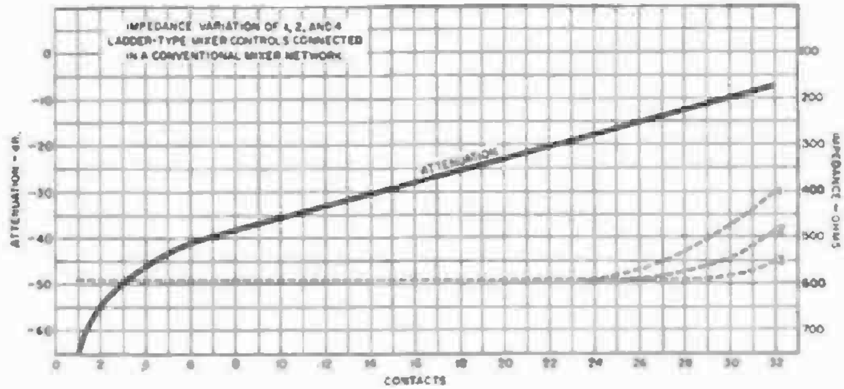


Fig. 5-83B. Impedance variations of 1, 2, and 4 ladder-type mixer controls connected in a conventional mixer network.

match. This subject is discussed in Section 9.

In the case of a slide-wire type attenuator, such as the ladder, the contact noise will be at a minimnm if the material of the contact arm is of the same alloy as the wire. However, noise may be generated when a signal is present if the contact is dirty or corroded because of rectification taking place between the two surfaces. Corrosion will cause a minute current to be generated and this current will be heard every time the arm moves from one wire to the next. As a rule, this type of noise can only be heard when a signal is present. If the contact arm and the wire are of different alloys, minute currents will be generated because of the dissimilar metals and, when amplified, the resulting noise will become quite annoying.

To eliminate the effects of contact noise, manufacturers are now producing variable attenuators with the contacts and arm sealed in *silicone grease*. This design increases the life of the contacts and eliminates noise.

5.86 ***What is an electronic attenuator?***—An amplifier designed to present a loss rather than a gain. Such devices have been constructed with a loss of 75 dB or more.

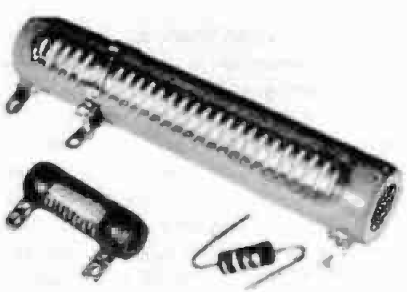


Fig. 5-87. Two variable vitreous resistors and a composition resistor.

5.87 ***What type resistors are used for terminating electronic equipment?***—The type of resistor will depend on the amount of power dissipated in the circuit. For terminating amplifiers and equipment developing a considerable amount of power, wirewound vitreous resistors are used as their self-inductance is quite small. Typical resistors of this type are shown in Fig. 5-87.

For circuits up to one half watt, small composition resistors (shown at the lower right of Fig. 5-87) may be used.

It is the nsual practice to allow a 100 percent safety factor insofar as the wattage rating of the resistor is concerned; that is, if the power dissipated is, say, 5 watts, a 10-watt resistor should be used. (See Question 23.56.)

5.88 ***How are carbon composition resistors constructed?***—By depositing a layer of carbon on a glass or ceramic rod to which wire pigtails are attached. The body of the resistor is completely sealed against moisture. Such resistors are used for all types of circuit elements and, except for certain types of equipment, are quite satisfactory for almost any application.

5.89 ***Describe the type resistance taper curves used in composition control potentiometers.***—Three basic resistance taper curves are shown in Fig. 5-89 for gain controls and other control functions in electronic equipment. Curve 1 is the taper most commonly employed in the manufacture of gain controls (volume controls) for use in amplifiers. The taper is a left-hand logarithmic curve, which provides a small amount of resistance at the beginning of the shaft rotation and a fast increase toward the end. All three curves are plotted for a *clockwise* rotation of the shaft.

The taper of curve 2 is an opposite of curve 1, and is right-hand logarithmic taper. The change in resistance is large for the first half of the shaft rotation and small in the second half. This taper is used for contrast controls in oscilloscopes and bias voltage controls.

The linear taper of curve 3 provides a rate of resistance change that is *proportional* to the shaft rotation. Such tapers are employed for tone control or where a straight-line voltage division is required.

The taper of any control may be measured easily by means of an ohmmeter. First measure the total resistance, then rotate the shaft exactly 50 percent. If the resistance measures 50 percent of its total, it is a curve 3 linear taper; if the measurement is 10 to 20 percent of the total the taper is logarithmic and corresponds to curve 1. But, if the measurement is around 80 percent of the total it is a *reverse logarithmic* taper and corresponds to curve 2.

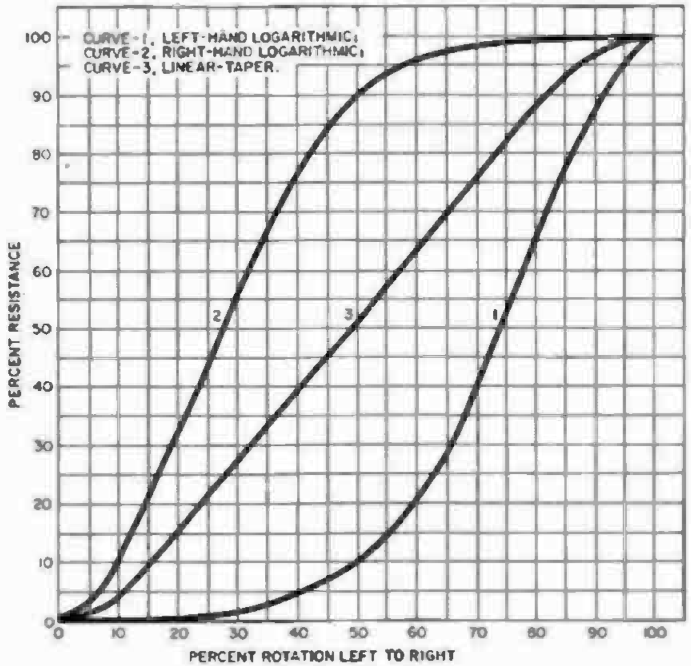


Fig. 5-89. Resistance taper curves for potentiometers. (Courtesy, Mallory Distributor Products, Co.)

5.90 ***What is a dual potentiometer or attenuator?***—Two continuously variable, or step attenuators connected mechanically together by a common shaft. It is used in equipment where the gain must be varied in two circuits simultaneously. Such a control is shown connected at the input of a push-pull amplifier in Fig. 5-90.

5.91 ***What is a compromise network?***—A network used with a hybrid coil for terminating a subscriber's loop (telephone term). The network is not precise but one that will result in the desired isolation between two directions of a hybrid coil. Hybrid coils are discussed in Questions 8.66 and 9.23.

2.92 ***Can a bridged-T attenuator and an L-type attenuator be connected in tandem?***—Yes. Any balanced or unbalanced attenuator may be directly connected to another, provided the impedance match is satisfied and the configurations are of such nature they will not cause an unbalanced condition. At (a) of Fig. 5-92 it is shown how an "L," a bridged-T, and a plain-T pad may be connected in tandem. At (b) is shown the method of connecting balanced attenuator configurations in tandem.

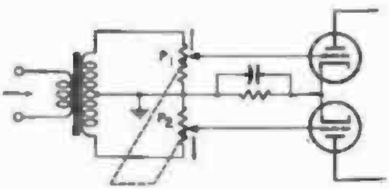
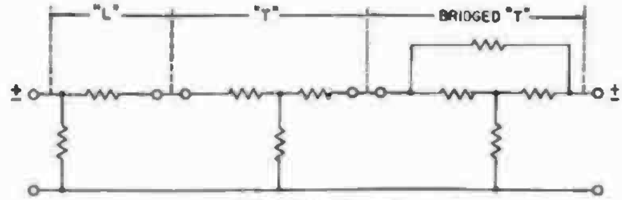


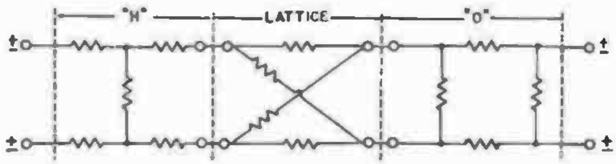
Fig. 5-90. Dual input attenuators that are used for controlling the output of a push-pull amplifier.

5.93 ***Describe a low- or high-frequency attenuator.***—A low- or high-frequency attenuator is in reality a low- or high-frequency equalizer for attenuating a given range of frequencies. An attenuator is a configuration of *noninductive* elements which in a given design have no effect on the frequency characteristics of the circuit in which it is operating. But, in a low- and high-frequency attenuator, *reactive* elements have been added to make it frequency sensitive for achieving a given frequency characteristic. Equalizers of this type are also termed shelf-equalizers and are used in sound mixer consoles. Such devices are discussed in Section 6 and Section 9.

5.94 ***Describe the construction of a precision potentiometer.***—Precision potentiometers and rheostats have been developed, both mechanically and electrically, to a high degree of accuracy, whereby linear accuracies of 0.015 percent are possible. Although such poten-



(a) *Unbalanced.*



(b) *Balanced.*

Fig. 5-92. Attenuators connected in tandem.

tiometers are not designed specifically for audio work, they find many uses for control circuits in a sound-recording activity. The interior construction of three different precision controls, manufactured by the Amphenol Controls Division, are illustrated in Figs. 5-94A, B, and C.

Referring to Fig. 5-94A, in this control the resistance *strip* consists of a single flat winding, with dual wipers for both the resistance strip and the rotor arm. The shaft is supported on a double shielded ball bearing, and can be rotated 354 degrees. At the rear is a clamp to permit the resistance element to be rotated through a few degrees for phasing with other controls. Up to six units may be ganged together. The resistance strip (depending on the total resistance) may contain 290 to 1457 turns, with linear accuracy of 0.5 percent for a resistance strip length of 2.3 inches. The housing diameter is $1\frac{1}{16}$ inches.

A spiral-type potentiometer that may be rotated 10 turns is illustrated in Fig. 5-94B. In this control, the resistance winding is wound around a large supporting wire (about 14 to 16 gauge enamel), then wound in a *spiral* which permits the wiper-arm to be rotated 3600 degrees or 10 turns. The rotor is of the double-wiper type with zero backlash. The number of turns in the winding around the large wire will vary from 1820 to 11,300, again depending on the total resistance of the winding. The winding covers a total length of 17.64 inches, with linear accuracy of 0.25 percent and a dissipation of 3 watts.

A second type spiral-wound control of 10 turns is shown in Fig. 5-94C, with a winding length of 43.5 inches. The resistance element will range from 2500 to 27,000 turns, depending on the total resistance, with linear accuracy of 0.10 percent, capable of dissipating 5 watts. In the latter two types, the bearings are of the sleeve type. Dials may be attached to the shaft to indicate the number of turns within the accuracy of the potentiometer, in both digital and indexing types. The latter type has an accuracy of 1000:1.

5.95 ***Describe how the linearity of a straight-line potentiometer may be altered.***—Very often in the design of electronic eqnipment, the need for a potentiometer with a special taper or characteristic is required. If a straight-line potentiometer is at hand, its linear characteristics may be altered by the shunting of a *fixed* resistance from one end of the resistance to the swinger-arm. Three methods of shunting a straight-line potentiometer are shown in Fig. 5-95. In the first method, the shunt resistor is connected from the swinger-arm to ground. With the correct value shunt resistance, the potentiometer will have a taper relative to the angular rotation, as shown below the schematic diagram. The second method makes use of a second potentiometer ganged with the straight-line potentiometer. In the third method, two *shunt* resistors connected at each side of the swinger results in a taper resembling a sine wave. A fourth method, not shown, use a shunt resistance connected from the swinger to the top of the potentiometer. For a detailed analysis of the above procedures, the reader is referred to the reference.

5.96 *Describe the construction of a photocell attenuator.*—The photocell attenuator is a radical departure from the conventional mixer control, inasmuch as there is no actual connection to the audio circuits in the mixing console. In the photocell-type attenuator the components consist of a source of dc connected to a *rheostat* in the form of a mixer control, a lamp, and a photocell.

The audio signal is amplified in the usual manner; however the gain of the amplifier is controlled by the *brillance* of the lamp placed in front of the photocell, which is caused to go from a dark condition to full brillance by the posi-

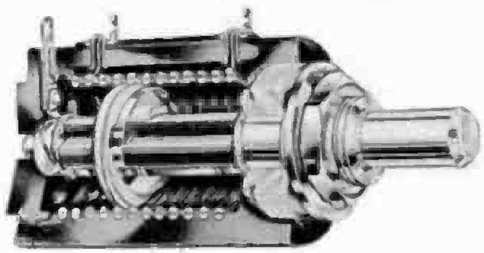


Fig. 5-94A. Amphenol Controls Division, Model 2150 precision potentiometer. Accuracy 0.25 percent.

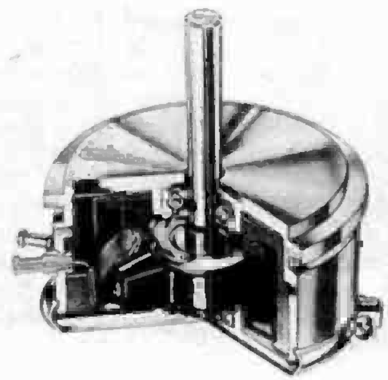


Fig. 5-94B. Amphenol Controls Division, Model 2450 precision potentiometer. Accuracy 0.5 percent.

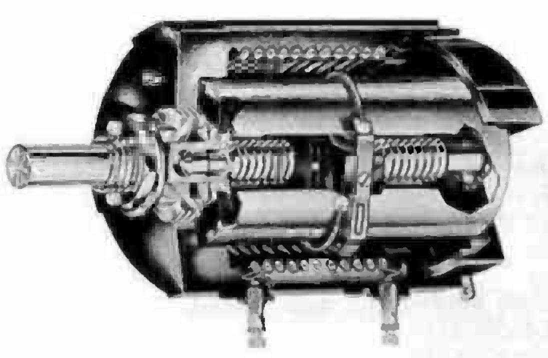


Fig. 5-94C. Amphenol Controls Division, Model 205 precision potentiometer. Accuracy 0.10 percent.

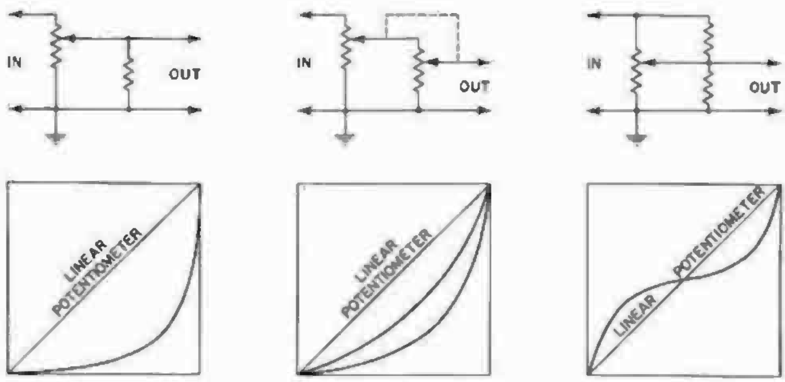


Fig. 5-95. Three methods of applying a shunt resistor to a potentiometer for altering its linear characteristics.

tion of the mixer control (rheostat). In this manner it is claimed that mixer-control contact noise is eliminated. However, there are some drawbacks to this system of mixing.

5.97 ***Describe the construction and operation of an electroluminescence attenuator control.***—Electro-optical attenuators are used in a number of different devices. The one to be discussed is that used in the Teletronix leveling amplifier, described in Question 12.84. Electroluminescence is a method of producing a light source by the passage of current through a thin layer of phosphor. The element, Fig. 5-97, is constructed somewhat similar to a capacitor. The light-producing element consists of a plate of glass or plastic coated with a clear conducting material on one side and a thin layer of phosphor on the

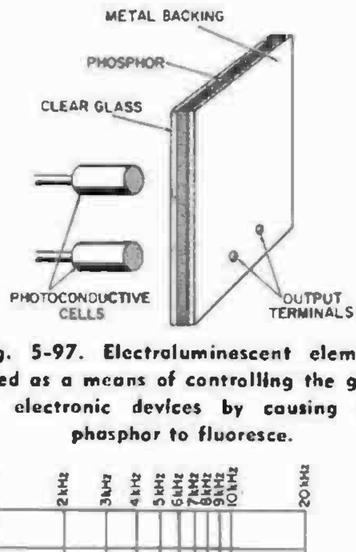


Fig. 5-97. Electroluminescent element used as a means of controlling the gain of electronic devices by causing the phosphor to fluoresce.

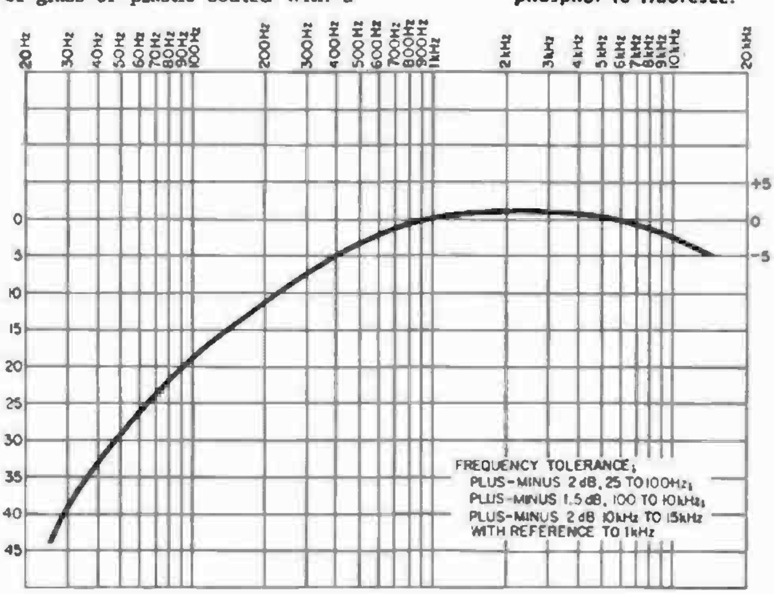


Fig. 5-98A. USASI (ASA) 51.4-1961 Curve "A" weighting network.

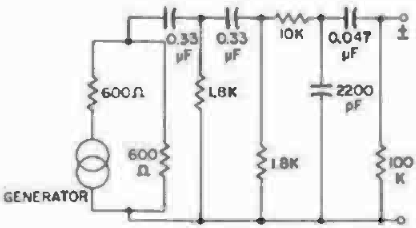


Fig. 5-98B. USASI (ASA) S1.4-1961 curve "A" weighting network for noise measurements, having a frequency characteristic similar to the human ear.

other. The metallic backing contacts the phosphor coating. By applying an alternating current to the conducting plate, the phosphors are excited by the voltage across the dielectric and light is produced. The amount of light emitted depends on the voltage and frequency. The light is picked up by a photoconductive cell whose resistance *decreases* with the impinging light, is amplified, and applied to the control circuitry.

5.98 Describe an RC network for weighted noise measurements.—It is now standard practice for manufactnrers of sound equipment to state the signal-to-noise ratio in terms of a *weighted* curve in accordance with the USASI (ASA) S1-4-1961 Standard Weighted Curve (Fig. 5-98A). This method of measnring signal-to-noise ratios results in a more *realistic* measurement, as the characteristic of the weighting network is similar to the characteristic of the human ear. The network shown in Fig. 5-98B is designed to operate from a 600-ohm circuit, terminated at its output by a vacuum-tube voltmeter or similar device with at least a 1-megohm input impedance, with ballistics and frequency characteristics of a standard volume indicator in accordance with USASI (ASA) Standard C16.5-1961. (See Questions 3.93 and 17.159.)

5.99 Describe an active network and its use.—Active networks are solid-state devices, with or without gain, designed to replace the conventional resistive network used in mixing networks. Fig. 5-99A shows such a network manufactured by Electrodyne having a total of

Fig. 5-99A. Electrodyne Model ACN-1 active combining network.

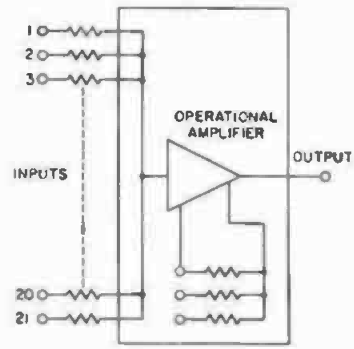


Fig. 5-99B. Block diagram for Electrodyne ACN-1 active combining network.

21 input circuits. Contained within the network is a solid-state integrated circuit (IC) amplifier which may be strapped for a no-gain operating condition, or for a gain of 10 or 20 dB, or varied in steps of 3 dB. Isolation between the input circuits is greater than 70 dB. The use of this device eliminates the need for a booster amplifier to compensate for the insertion loss induced by a resistive type combining network. A block diagram of its internal connections is given in Fig 5-99B.

The frequency response is plus-minus 0.5 dB, 20 to 20,000 Hz, with less than 0.3 total harmonic distortion (THD) for any level up to plus 18 dBm. The input impedance for each input is 10,000 ohms, with an output impedance of 600 ohms. Power requirements are; 24 Vdc at 20 mA. The internal circuitry is protected against accidental reversal.

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Section 6

Equalizers

All sound recording and reproducing equipment have one thing in common—the loss of high-frequency response. A greater part of the basic research work was accomplished in the early stages of development of the telephone by Zobel, Shay, Bode and Sallie Pero Mead and others of the Bell Telephone Laboratories. Many of their designs are in current use. In this section, different types of equalization are discussed, using simplified equations and graphs for their design. Numerous practical circuits are given for both recording and reproducing equipment, both solid state and vacuum tube. Of interest to the audio engineer will be the graphs for designing different types of bridged-T equalizers.

6.1 ***What is an equalizer?***—A device consisting of reactive elements which may be connected into an electrical circuit for the purpose of altering the frequency characteristics of that circuit.

6.2 ***What is a compensator?***—It is another name for an equalizer.

6.3 ***What is a duller?***—A form of equalizer used to reduce the high-frequency response of an electrical circuit. It is so named because the high-frequency response appears to be dull and lacking in presence.

6.4 ***What does the term pre-emphasis mean?***—It is the same as pre-equalization. (See Question 6.7.)

6.5 ***What does the term post-emphasis mean?***—It is the same as post-equalization. (See Question 6.8.)

6.6 ***What does the term de-emphasis mean?***—It is the same as post-emphasis.

6.7 ***Define pre-equalization.***—Equalization which is inserted in the recording circuits. (See Question 6.11.)

6.8 ***Define post-equalization.***—Equalization inserted in the reproducing circuits. (See Question 6.11.)

6.9 ***What is the take-off point of an equalizer?***—The frequency where the equalization starts to become effective.

6.10 ***What is a low- or high-frequency attenuator?***—A circuit used in a mixer network to reduce the low- or high-frequency response. It generally consists of a shunt capacitor or reactor in parallel with an input circuit to reduce the high frequencies, or a series capacitor for reducing the low frequencies. This subject is further discussed in Questions 6.80 and 6.81.

6.11 ***What is the purpose of pre- and post-equalization?***—Pre-equalization is connected in the recording circuits and is used to *increase* the amplitude of frequencies above 1000 Hz, and to obtain a greater signal-to-noise ratio during reproduction. Post-equalization is connected in the reproducing circuit and has an *inverse* frequency characteristic to that of the pre-equalization.

The characteristics of both type equalizers is shown in Fig. 6-11. It will be observed that during recording 10,000 Hz is pre-equalized 15 dB with reference to 1000 Hz. When reproduced, 10,000 Hz is post-equalized (attenuated) 15 dB, thereby reducing the surface noise at these frequencies by the same amount. (This curve does not represent present-day pre- and post-equalization; the values shown are for illustration only.)

Pre- and post-equalization may be applied to any type recording system. This subject is further discussed in Section 13 and Section 18.

6.12 ***Describe a variable equalizer.***—It is an equalizer network in which the amount of equalization may be increased or decreased at will. Equalizers of this type are used for rerecording (and other purposes) and are provided

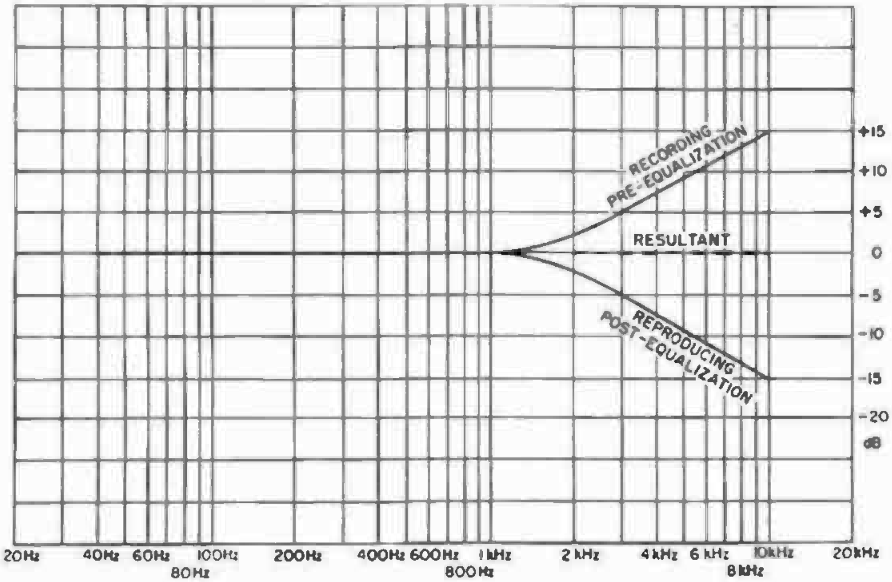


Fig. 6-11. Typical pre- and post-equalization curves.

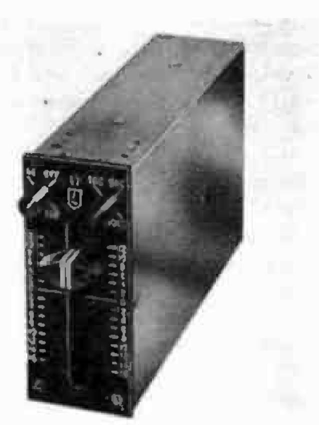


Fig. 6-12A. Program equalizer Model EQ-251A manufactured by Langevin.

with switches for changing the values of the network components. Their design is such they may be changed during actual program operation without affecting the gain or impedance, or introducing noise.

A variable program equalizer manufactured by Langevin is shown in Fig. 6-12A, with its many frequency-response characteristics shown in Fig. 6-12B. Over 1500 different combinations of equalization and attenuation are possible, at frequencies of 40, 100, 3000, 5000, 10,000, and 15,000 Hz. The configuration is a passive constant-B network, employing inductance and capacitance.

Two vertical slide-type switches provide control for a maximum of 12-dB equalization, and a maximum attenua-

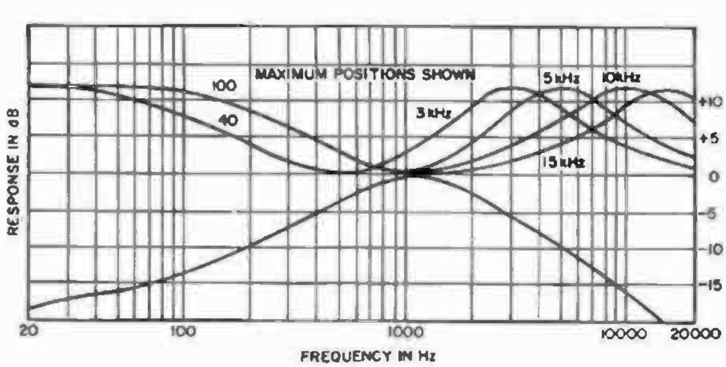


Fig. 6-12B. Frequency-response characteristics of Langevin Model EQ-251A program equalizer.

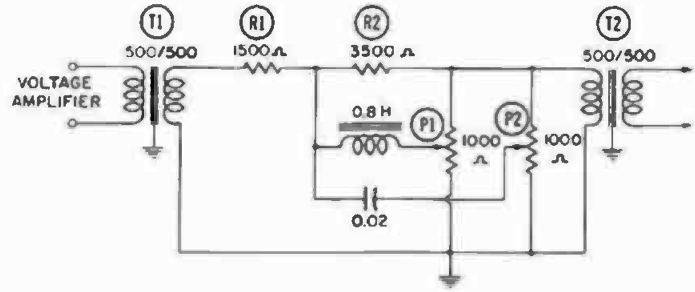


Fig. 6-13A. A simple variable equalizer circuit.

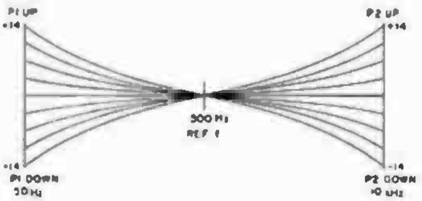


Fig. 6-13B. Frequency characteristics of the variable equalizer circuit given in Fig. 6-13A.

tion of 16 dB. Two rotary switches permit the selection of the peak frequencies. The device presents a *constant impedance* of 600 ohms, with *constant insertion loss* of 14 dB. Because of its physical size, it may be installed above a mixer control The internal wiring consists of printed-circuit boards.

6.13 Show a simple variable equalizer for use in a low-impedance circuit. —A simple variable equalizer consisting of two resistors, an inductance, and a capacitor is shown in Fig. 6-13A. Near the center setting of the variable pots P1 and P2, the frequency response is flat. Moving P1 upward increases the low-frequency response and moving it downward decreases the low-frequency response. When P2 is moved upward the high-frequency response is increased and the reverse takes place when it is moved downward. The total equalization is 28 dB, or 14 dB plus or minus the reference frequency. This equalizer has two disadvantages: (1) the insertion loss is not constant and varies with the amount of equalization and (2) the impedance is not constant. Therefore, the device must be operated from and into circuits presenting a solid termination. The approximate equalization obtained for different settings of P1 and P2 is shown in Fig. 6-13B.

6.14 What effect does an equalizer have on the gain of a circuit?—It presents a loss at a given reference frequency. The amount of loss will depend on the circuit design. If the insertion loss is appreciable, additional amplification will be required to compensate for the insertion loss.

6.15 What is a constant-loss equalizer?—An equalizer having a constant insertion loss regardless of the amount of equalization. Variable equalizers, unless specially designed, have a variable amount of equalization. To overcome this difficulty, equalizers have been designed that will present a constant loss for any setting of their equalization. Such equalizers employ two attenuator pots connected mechanically in such a manner that, as the loss of one pot is increased, the other is reduced in a like amount. Thus, the loss in the circuit remains constant—only the amount of equalization is changed. This equalizer is often erroneously referred to as a constant-gain equalizer. A diagram of a constant-loss equalizer is shown in Fig. 6-15. It will be noted that the loss of the two pots P1 and P2 always remains the same—for this example, 12 dB. Therefore, the insertion loss is 12 dB. (See Question 6.107.)

6.16 What effect does an equalizer have on the impedance relations of a circuit?—If the equalizer is not of constant resistance or impedance, the impedance of the circuit will be disturbed and the expected equalization may not be obtained. Whenever possible, constant-resistance equalizers should be used.

6.17 What is a constant-resistance equalizer?—A configuration using a constant resistance or impedance pad in conjunction with reactive elements. The reactive elements in the shunt and series arms have an inverse relationship; therefore, the impedance at either the input or output is essentially constant.

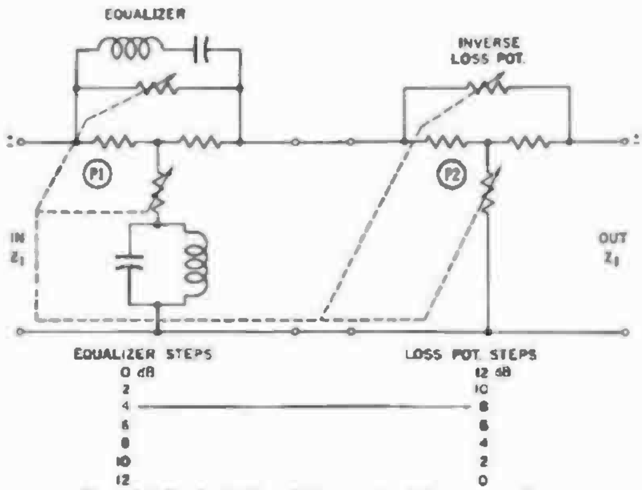


Fig. 6-15. A bridged-T constant-loss equalizer.

6.18 What is the configuration for a constant-resistance equalizer?—Generally, a plain-T pad, as shown in Fig. 6-18. If the equalizer is to be variable, a T pad, variable in steps of 1 dB, is substituted for the fixed pad. As a rule, variable T-type equalizers employ bridged-T attenuators as described in Question 6.48.

6.19 What is the maximum equalization that can be obtained with a constant-resistance equalizer?—About 20 dB is the practical limit. If a greater amount of equalization is required, two or more equalizers may be connected in tandem. The amount of equalization at the lower frequencies is limited by the reactive elements, particularly the inductances. Only coils of the highest Q should be employed in conjunction with high-grade paper or mica capacitors. The coils should have their maximum Q at the resonant frequency.

6.20 What is an L-type equalizer?—One employing a single coil and capacitor as shown in Fig. 6-20. The chief disadvantage of this equalizer is that it does not present a constant resistance to the operating circuits. (See Question 6.95.)

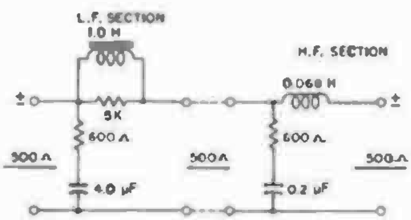


Fig. 6-20. L-type equalizers connected in tandem. Values shown are suitable for the reproduction of 16" transcriptions using the NAB characteristic.

6.21 What does the term "shelving" mean when applied to an equalizer?—It means that the frequency response has a shelflike characteristic at the upper and lower ends of the spectrum as shown in Fig. 6-21.

6.22 What is the phase shift for a bridged-T equalizer?—About 40 degrees.

6.23 What is an attenuator equalizer?—An equalizer using an attenuator

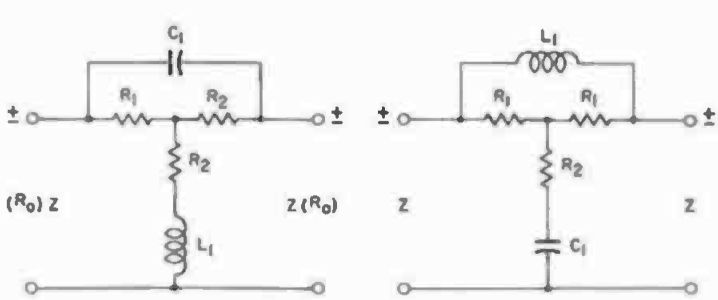


Fig. 6-18. Constant-resistance or impedance equalizers.

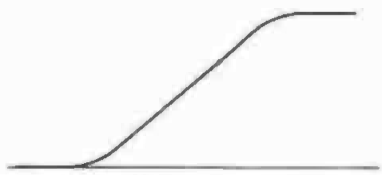


Fig. 6-21. Shelving characteristic of an equalizer.

network for controlling the amount of equalization. They are also known as constant-resistance or constant-impedance equalizers. The equalizers described in Question 6.48 are attenuator equalizers.

6.24 ***Describe an equalizer amplifier.***—Equalizer amplifiers were used in the early days of radio for reproducing 16-inch electrical transcriptions. The required equalization consisted of RC equalizer sections interposed between the first and second stages. The advantage of this design was that the equalizer and preamplifier were one unit. Equalizer amplifiers are now obsolete.

6.25 ***What is a degenerative amplifier equalizer?***—An equalizer employing reactive components in the cathode and plate circuits of a vacuum tube (Fig. 6-25). The amount of equalization is controlled by the two potentiometers, which control the degeneration in the cathode circuit. Many different combinations of both the low and high frequencies are possible.

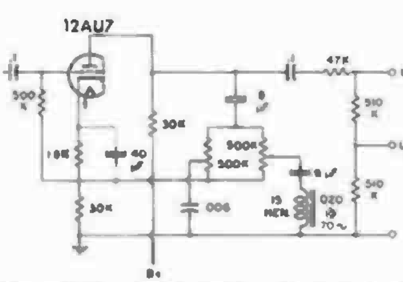


Fig. 6-25. Configuration of a cathode equalization circuit.

This amplifier is often used in public address systems. Because the amount of equalization affects the gain of the amplifier stage, the main gain control of the amplifier system must be changed to compensate for every change in the amount of equalization. The potentiometers are especially designed for this equalizer circuit.

6.26 ***What are the characteristics of a resonant circuit containing capacitance and inductance?***—Referring to the graph in Fig. 6-26, it will be seen that as the frequency is increased from the lowest value to a high value, the capacitive reactance (X_C) decreases, while the inductive reactance (X_L) increases. At the resonant frequency, the circuit behaves as a resistance. Below the resonant frequency it behaves as a capacitor, while above the resonant frequency, the circuit behaves as an inductance. Below the resonant frequency, the current leads the voltage, while above the resonant frequency, it lags the voltage.

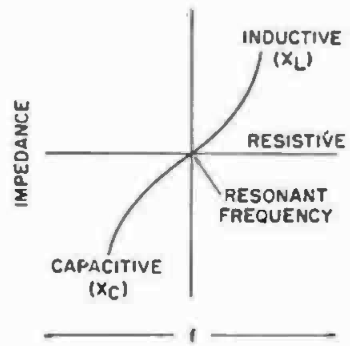


Fig. 6-26. Characteristics of a resonant circuit consisting of inductance and capacitance.

6.27 ***What is a diameter equalizer?***—In disc recording, as the smaller diameters are approached, the groove velocity is lowered, with an accompanying loss in high frequencies, particularly above 3000 Hz. (See Question 13.49.) To overcome this deficiency, diameter equalization is automatically introduced in the recording circuits at diameters previously determined by a family of frequency-response measurements, at a rate of 1 dB per step. The resonant frequency of the equalizer is about 10,000 to 12,000 Hz. Diameter equalization should not be confused with pre-equalization, for it is a form of equalization used *in addition* to regular pre-equalization.

The principal objections to the use of diameter equalization are that it increases the intermodulation distortion at the smaller diameters and adds to the difficulty of tracking the pickup during reprodnction. Also, the power handling capabilities of the recording amplifier must be increased to overcome the additional insertion loss of the diameter equalizer. As a rule, for installations using the standard RIAA pre-equalization and 6 dB of diameter equalization, a recording amplifier of 100 to 120 watts output is required. Since the adoption of the hot-stylus technique for recording disc records, diameter equalization is not used to any great extent. Hot stylus recording techniques are discussed in Questions 15.58 to 15.71.

6.28 ***What type equalizer is used for diameter equalization?***—A bridged-T, constant-loss type. (See Question 13.111.)

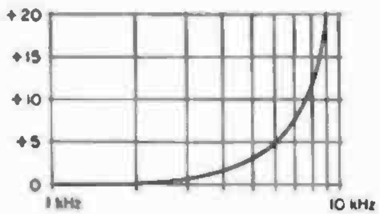


Fig. 6-29A. Transmission characteristics of a high-frequency equalizer.

6.29 ***How are equalizer characteristics plotted?***—Equalizer characteristics may be plotted in two ways, using either the transmission loss or the insertion loss versus frequency. As a rule equalizer plots are made showing their transmission characteristics as it is easier to picture their operation in a recording circuit. An insertion-loss plot is used when designing such devices.

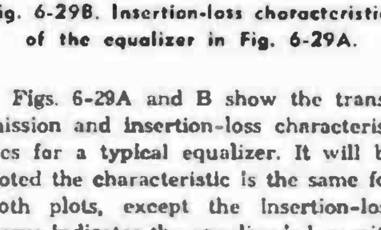


Fig. 6-29B. Insertion-loss characteristics of the equalizer in Fig. 6-29A.

Figs. 6-29A and B show the transmission and insertion-loss characteristics for a typical equalizer. It will be noted the characteristic is the same for both plots, except the insertion-loss curve indicates the equalizer's loss with respect to frequency and the transmission curve the gain of the circuit with respect to frequency.

6.30 ***Can equalization be obtained with simple circuits containing only capacitance and inductance?***—Yes, except that such circuits are not constant-impedance and a given frequency response is difficult to obtain.

6.31 ***If a capacitor is connected in series with a circuit, how is the frequency response affected?***—If the capacitor is small, the low frequencies will be attenuated as shown in Fig. 6.31. The smaller the capacity, the greater will be the attenuation.

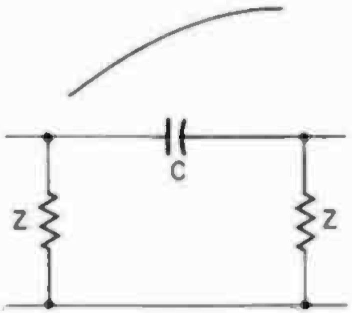


Fig. 6-31. A simple equalizer using one capacitor in series with a circuit.

Although this circuit could be used to increase the high-frequency response or decrease the low-frequency response, it is not entirely satisfactory because the rate of cutoff, frequency response, and impedance match cannot be controlled to any great extent. The circuit functions by virtue of the capacitive reactance. At the low frequencies the reactance is quite high and at the high frequencies it is quite low, permitting the higher frequencies to pass with little attenuation. The capacitor might be looked upon as a variable resistance in series with the circuit controlled by the applied frequencies. Therefore, the impedance varies with frequency.

6.32 ***If a capacitor is connected in parallel with a circuit, how is the frequency response affected?***—The higher frequencies are attenuated. The larger the capacity the greater will be the attenuation, and the reverse of Fig. 6-31. As the higher frequencies are approached, the capacitive reactance decreases, shunting the high frequencies to the low potential side of the circuit. It might be said that the capacitor acts like a *variable short circuit*, controlled by the applied frequencies. The reactance of the capacitor for any frequency may be computed:

$$X_C = \frac{10^6}{2\pi fC}$$

where,

X_C is the capacitive reactance in ohms,
f is the frequency in Hz,
C is the capacitance in microfarads.

At a frequency where the capacitive reactance equals the circuit impedance, the response is down 3 dB.

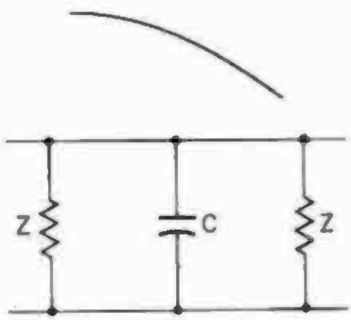


Fig. 6-32. A simple equalizer using one capacitor in parallel with a circuit.

When the capacitive reactance is one-tenth of its original value, and the impedance of the circuit is ten-times the reactance of the capacitor, the change with frequency becomes negligible. To evaluate the change with frequency, it is necessary to refer to a reference frequency, generally 1000 Hz.

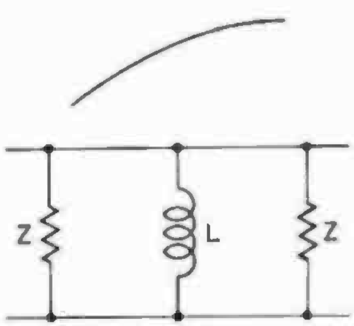


Fig. 6-33. A simple equalizer using one inductor in parallel with a circuit.

6.33 If an inductance is connected in parallel with a circuit, how is the frequency response affected?—If a constant voltage is applied to the input and the frequency varied, the voltage at the output will *increase* with frequency, because of the increase in the inductive reactance of the coil. This is similar to the series capacitor in Fig. 6-31. The

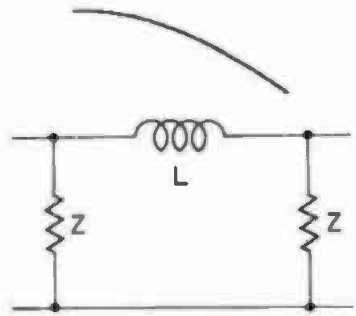


Fig. 6-34. A simple equalizer using one inductor in series with a circuit.

circuit impedance will vary with frequency.

6.34 If an inductance is connected in series with a circuit, how is the frequency response affected?—As the frequency is increased the inductive reactance *increases*, attenuating the higher frequencies, and the response becomes similar to that in Fig. 6-32. The inductive reactance for any frequency may be computed:

$$X_L = 2\pi fL$$

where,

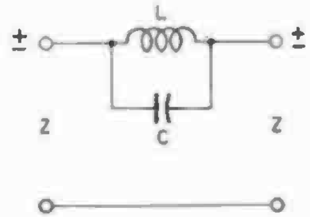
X_L is the inductive reactance in ohms,
f is the frequency in Hz,
L is the inductance in henries.

The inductive reactance of the coil may be looked upon as a series resistance which varies with frequency; thus the circuit impedance changes with frequency.

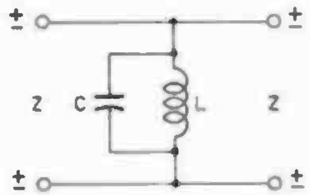
6.35 What is a parallel-resonant equalizer?—A configuration with the circuit elements connected in parallel. The completed network is then connected either in series or parallel with the circuit to be equalized. (See Fig. 6-35.)

6.36 What is a series-resonant equalizer?—A configuration with the circuit elements connected in series. The completed network is then connected either in series or parallel with the circuit to be equalized. (See Fig. 6-36.)

6.37 Why is it necessary to equalize a telephone line?—For ordinary telephone lines, it is not necessary unless they are to be used over long distances. Telephone lines used for radio transmission are equalized to secure a uniform frequency response within limits, depending on the classification of the line. A telephone line may be considered as a low-pass filter, the circuit



(a) Connected in series with a line.



(b) Connected in parallel with a line.

Fig. 6-35. Parallel-resonant equalizer circuits.

elements being series inductance, shunt capacitance, and the dc resistance of the conductors.

To correct for these deficiencies, equalizers, phase-correction networks, and amplifiers are connected in the line at given intervals called repeater stations. The amplifiers compensate for line losses and the insertion loss of the corrective networks. The phase-correction network reduces distortion due to phase shift. (See Questions 7.23 and 7.24.)

6.38 ***Show a simple telephone-line equalizer.***—Fig. 6-38 shows a parallel-resonant equalizer connected in parallel with a line through a variable resistor. The capacitor and coil in parallel resonate at the desired frequency. The resistor R controls the amount of equalization.

The resonant frequency of the equalizer is made slightly higher than the highest frequency to be transmitted by the line. High quality lines are equalized to within plus or minus 1 dB from a reference frequency of 1000 Hz. A typical line-equalizer characteristic is shown in Fig. 6-39.

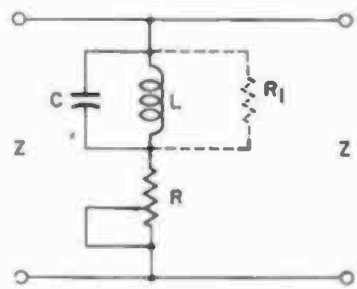


Fig. 6-38. A simple telephone-line equalizer network.

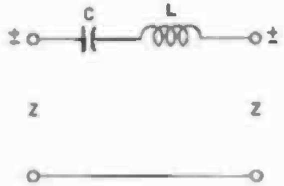
6.39 ***Show the frequency characteristics of a line before and after equalization.***—The plot of a typical telephone line consisting of a No. 19 nonloaded cable pair, 10.8 miles in length is shown in Fig. 6-39. Curve A shows the frequency response of the line before equalization, and C after equalization. Curve B is the equalizer characteristic alone. Although the line characteristic at C is plotted to 10,000 Hz, it is the practice to equalize lines for broadcast use to within 1 dB at 8000 Hz only. For fm use, the frequency response is extended beyond 15,000 Hz.

It will be noted the equalizer reduced the line level from a minus 14 dBm to a minus 30 dBm, after equalization. This loss is caused by the insertion loss of the equalizer.

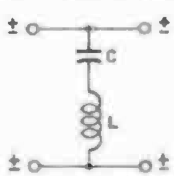
The Western Electric 23A equalizer which is similar to that shown in Fig. 6-38 may be used on longer lines with the following results:

Line Length (Miles)	*Cable Pair Size*	*Equal. at 10 kHz*
10.0	19	1 dB
11.5	19	2 dB
21.5	16	1 dB
25.0	16	2 dB

This is for a nonloaded pair. Cable loading is discussed in Question 25.186.



(a) Connected in series with a line.



(b) Connected in parallel with a line.

Fig. 6-36. Series-resonant equalizer circuits.

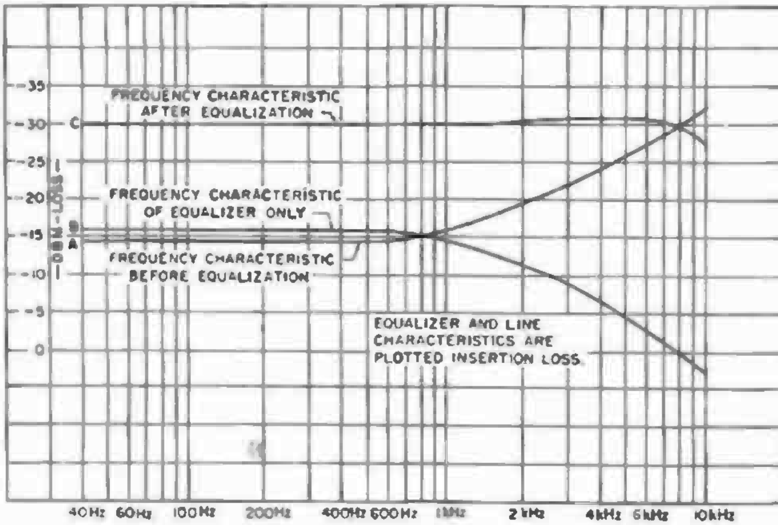


Fig. 6-39. Frequency characteristics of No. 19 nonloaded cable before and after equalization using Western Electric 23A equalizer.

6.40 How is the amount of equalization estimated for a given line?—The amount of equalization may be calculated from the known characteristics of the line. Open pairs require less equalization than cable pairs, because of the greater distributed capacity and series inductance of the cable pairs. The loss in dB per mile for standard telephone cable may be obtained from the cable manufacturer, and the equalizer designed accordingly.

6.41 Explain how the equalizer in Fig. 6-38 functions.—An equalizer may be looked upon as a variable-loss device, controlled by the applied frequencies, which varies the gain of the circuit (at a given frequency) in which it is connected. Frequencies attenuated by a line or device cannot be replaced; however, the predominating frequencies can be reduced to the level of the attenuated frequencies. Thus, a uniform level or response is obtained.

Referring to the equalizer in Fig. 6-38, below the resonant frequency, a parallel-resonant circuit functions as an inductance and offers a low reactance at the low frequencies and a high reactance at the high frequencies. As the frequency is increased, the reactance increases and builds up the voltage across the line, with the greatest voltage occurring at the resonant frequency. This action is opposite to that of the transmission-line characteristic which attenuated the high frequencies.

An equalizer of the type shown in Fig. 6-38 may, therefore, be viewed as a variable short circuit in parallel with the line and controlled by the applied frequencies. At the lower frequencies the reactance is low and at the high frequencies it is high. Thus, at the high frequencies, the shunt effect of the equalizer is reduced to a small amount.

The value of the series resistance R is determined by the length of the transmission line. As the distributed capacity of the line increases, the high-frequency loss increases. Hence, less series resistance is required for a long line than for a short one.

6.42 Where should a telephone equalizer be connected?—At the receiving end of the line. This will result in a maximum signal-to-noise ratio.

6.43 Describe the procedure for designing a telephone-line equalizer.—The first step in the design is to measure the frequency characteristics of the line to be equalized, as described in Question 23.96. After the response has been plotted, the resonant frequency of the equalizer may be determined.

Assume an equalizer is to be designed for a transmission line having a frequency characteristic of that shown in Fig. 6-39. It will be observed that at the reference frequency of 800 Hz, the line has a loss of 15 dB, compared to 8000 Hz. A rising frequency characteristic is required; therefore, a parallel-resonant circuit is employed and connected in shunt with the line (Fig. 6-44). Since the highest frequency to be eqnalized is 8000 Hz, a resonant frequency of 9000 Hz will be used. The line will be terminated in 600 ohms. To obtain an equalization of 15 dB, a *toroidal-wound* coil is required, with a "Q" of at least 50 or 60 at 9000 Hz. The value of the inductance may be found with the assistance of the reactance chart in Fig. 25-201. As this is to be a high-frequency equalizer, a ratio of capacitance to inductance is selected that will result in a low inductive reactance at 100 Hz.

Selecting a value of 100 ohms at 100 Hz, the reactance chart is entered at 100 Hz at the bottom of the chart and followed upward to where an inductive reactance of 100 ohms is encountered on the left-hand margin. At this point, a diagonal line representing a value of 160 millihenries crosses the 100-Hz line. The 160-millihenry line is followed to where it intersects the 9000-Hz line. At this juncture, another diagonal line is encountered. Following this downward to the right-hand margin, the value of capacitance is found to be 0.00195 μF. This value of capacitance in parallel with a 160-millihenry coil will resonate at 9000 Hz. The reactance of the network may now be read at the left-hand margin, where it crosses the 9000-Hz line; for this example this is 9000 ohms.

In a parallel-resonant circuit, the voltage falls off on either side of the resonant peak, attenuating the frequencies on both sides. Below the resonant frequency, the circuit functions as an inductance; hence, it offers a *low* reactance to the low frequencies and a *high* reactance to the high frequencies. As the frequency increases, the inductive reactance increases and the voltage across the line is built-up, the greatest voltage occurring at the resonant frequency. This action is *diametrically opposite* to that of the transmission line, whose attenuation *increases* with frequency.

Thus, the equalizer may be viewed as a variable short circuit in parallel with the line, with the amount of short-circuiting a function of frequency. At 100 Hz the network used for this illustration presents an impedance of 100 ohms in parallel with the line impedance, thereby attenuating the low frequencies. At 9000 Hz the network impedance rises to 9000 ohms; thus, the voltage across the line at 9000 Hz is increased.

6.44 *How is a line equalizer adjusted?*—By sending a series of constant-amplitude frequencies from the sending end of the line to the receiving end where the equalizer is connected. (See Fig. 6-44.) The resistive networks at the send and receive ends supply a solid resistive termination, while isolating the line from the measuring instruments. The series resistance R is adjusted for a uniform frequency response within a given set of limits. This subject is further discussed in Question 23.96.

6.45 *If more equalization is required than can be obtained with one equalizer, how is the equalization increased?*—By connecting two equalizers in tandem. The total equalization is the sum of the two, if they have the same frequency characteristics. If the characteristics are not the same, it will be the algebraic sum of the two.

6.46 *What is a series-resonant equalizer and how is it designed?*—Series-resonant equalizers are designed in a manner similar to the parallel-resonant equalizer, using the reactance chart in Section 25. Series-resonant equalizers are connected in parallel, with a line to remove peaks as shown in Fig. 6-46A. The action of this con-

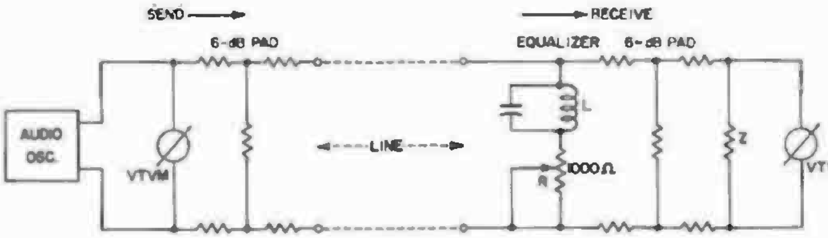


Fig. 6-44. Test setup for measuring the frequency characteristic of a telephone line with equalization.

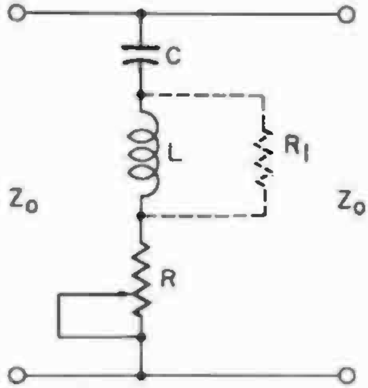


Fig. 6-46A. A simple series-resonant equalizer network.

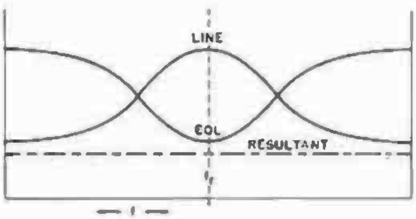


Fig. 6-46B. The effect of connecting a series-resonant equalizer in parallel with a transmission line and the resulting frequency response.

figuration is the reverse of the parallel-resonant type. In Fig. 6-46B is shown a peak removed from a circuit by the use of a series-resonant equalizer connected as shown in Fig. 6-46A. The resistance shown in dotted lines may be connected in parallel with the coil to lower the Q of the circuit, thus, flattening out the response. However, the use of this resistance will also reduce the amount of equalization normally obtained without it.

6.47 ***How are the exact values of capacitance obtained for equalizer construction?***—By connecting capacitors in parallel, in series, or in series-parallel.

6.48 ***Describe the four basic bridged-T equalizers most commonly used in audio work.***—The four basic bridged-T equalizers are shown in Figs. 6-49A and B and 6-50A and B, with their frequency characteristics depicted with their configurations. The configuration in Fig. 6-49A is a low-frequency shelf-type suppressor equalizer employing only two reactive elements, a capacitor and an inductance. The positions of the reactive elements have been reversed in Fig. 6-49B. A high-frequency shelf-type suppressor equalizer is the result, with an inverse characteristic of that in Fig. 6-49A. The configurations shown in Figs. 6-50A and B are similar in nature but employ series and parallel resonant circuits to achieve their characteristics. Although the configurations shown are designed for bridged-T pads, straight-T pads may be substituted, if desired. For variable equalization, a 20-dB bridged-T pad, variable in steps of 1 dB, is used. As these equalizers are of constant impedance or resistance, they must be operated from, and in to, a solid termination. The pads may be designed from the data given in Questions 5.33 and 5.37.

6.49 ***How do the reactive elements in a bridged-T equalizer function?***—They control the pad loss as a function of frequency. In the design of a bridged-T pad, as described in Question 5.37, the two resistors R_1 are equal in value to the circuit impedance. Resistors R_5 and R_6 are varied in their values, inversely to each other, to obtain different values of loss. For a condition of zero loss, R_5 has zero resistance and R_6 infinite resistance. Now, if reactive elements are connected in parallel and in

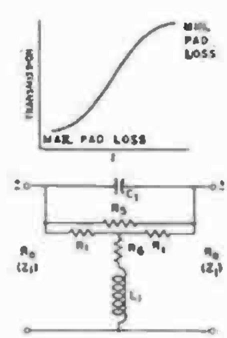


Fig. 6-49A. High-frequency shelf-type suppressor equalizer.

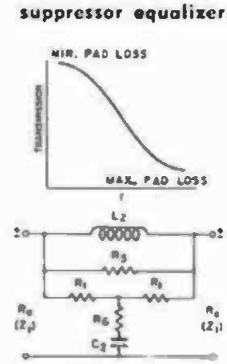


Fig. 6-49B. Low-frequency shelf-type suppressor equalizer.

series with resistors R_5 and R_6, and R_5 and R_6 are selected for a given pad loss, the reactive elements may be employed to control the pad loss as a function of the applied frequencies.

The configuration shown in Fig. 6-49A is a high-frequency equalizer similar to that used in recording circuits for pre-equalization. Capacitor C_5 is connected in parallel with R_5 and the inductance L_6 in series with R_6. At the lower frequencies C_5 has a high reactance and the inductance L_6, a low reactance. Under these conditions, the pad loss is approximately normal.

As the frequency is increased, the reactance of C_5 starts to decrease and the reactance of L_6 starts to increase. When the maximum frequency is reached, C_5 shunts out R_5 and L_6 has, for all intents and purposes, opened up R_6 (because of the high reactance of L_6). Thus, the pad loss is theoretically zero.

Removing the pad loss from the circuit increases the gain at the higher frequencies. When complex waveforms are applied to the equalizer, the pad loss will appear to be different for each frequency component in the waveform. Therefore, the pad loss is controlled by the applied frequencies. The frequency response shown with the configuration is the transmission characteristic plotted for frequencies of constant amplitude applied to the input of the network. Equalizers of this type may be used to increase the frequency response of a circuit at either end of the spectrum. It will be noted that the response at the extreme upper and lower ends is shelved.

Twenty dB of equalization is about all that can be obtained with a single equalizer of this configuration. If a greater amount of equalization is required than can be obtained with a single unit, two or more equalizers may be connected in tandem. The total equalization will be the algebraic sum of the two equalizers. The coils should be of toroidal design with a Q at the lower frequencies of at least 25.

6.50 What is the purpose of using resonant circuits with bridged-T equalizers?—To resonate the equalizer at a predetermined frequency. Equalizers using this type configuration may be designed to attenuate or accentuate certain frequency bands. These designs may also be used to equalize the extreme ends of the spectrum, by selecting the proper circuit elements and using the left or right side of the resonant response curve shown above the configurations in Figs. 6-50A and B.

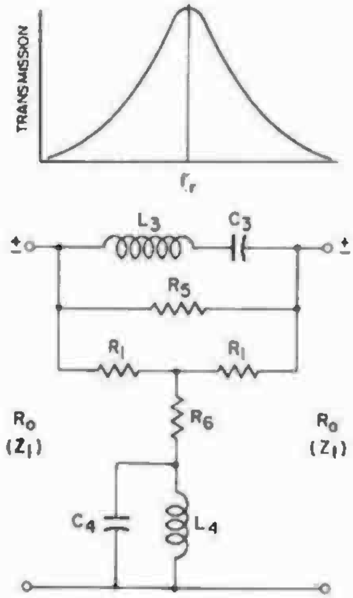


Fig. 6-50A. A resonant circuit peak-type equalizer.

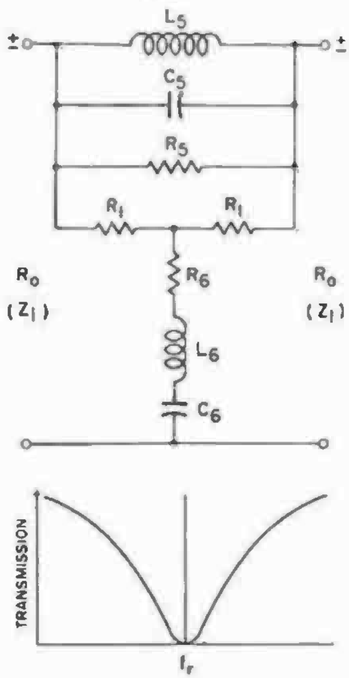


Fig. 6-50B. A resonant circuit dip-type equalizer.

When using the configuration in Fig. 6-50A to increase the high-frequency response, the left side of the response curve is used to provide the lift. For low-frequency equalization, the right side of the curve provides the lift.

The configuration in Fig. 6-50B may be used to attenuate either the high or low frequencies. The right side of the response curve is used for attenuating the lower frequencies and the left side the high frequencies.

The action of the reactive circuits is similar to that described for the configurations in Figs. 6-49A and B. As the frequency of resonance is approached, the impedance of L_3 and C_3 in Fig. 6-50A decreases, shunting out R_3. While this is happening, the reverse is taking place at L_4 and C_4. Thus, the pad loss is removed from the circuit. It will be noted the positions of the resonant circuits are reversed in Fig. 6-50B. Therefore, the actions of the resonant circuits are also reversed and, thus, an inverse frequency characteristic to that in Fig. 6-50A is obtained.

It should be remembered that in a parallel-resonant circuit the impedance increases with an increase in frequency. The reverse is true for a series-resonant circuit. The configurations shown may be designed to induce a peak or dip anywhere in the frequency spectrum, by the selection of the proper configuration and circuit-element values. These equalizers may also be made variable in steps of 1 dB by the use of a 20-dB variable attenuator. The maximum equalization is about 15 dB, although more may be obtained if toroidal coils with the maximum Q occurring at the resonant frequency are used.

6.51 ***How is the equalizer shown in Fig. 6-49A designed?***—For this design, the configuration in Fig. 6-49A will be used. The transmission characteristics for the configurations of Fig. 6-49A and B are given in Fig. 6-51A. The curves are plotted for a 10-dB loss pad. The figure 1.0, at a point where the curves cross is termed the *crossover* frequency and represents the frequency where *one half* the pad loss occurs. For this example, curve 1 will be used, with point 1.0 representing 1500 Hz. The frequency response may now be plotted as in Fig. 6-51B, by first indicating a loss of 5 dB at 1500 Hz. Again referring to Fig. 6-51A, at 3000 Hz (2.0) the loss is 1.25 dB; and at 4500 Hz (3.0), the loss is negligible; at a frequency of 750 Hz (0.5), the loss is 8.75 dB; and at 500 Hz (0.3), it is 10 dB.

To make the curve fit a particular requirement, the frequency at the crossover point (1.0) is shifted up or down on the curve, as required. Three different plots, using *one-half pad loss* frequencies of 100, 1000, and 3000 Hz

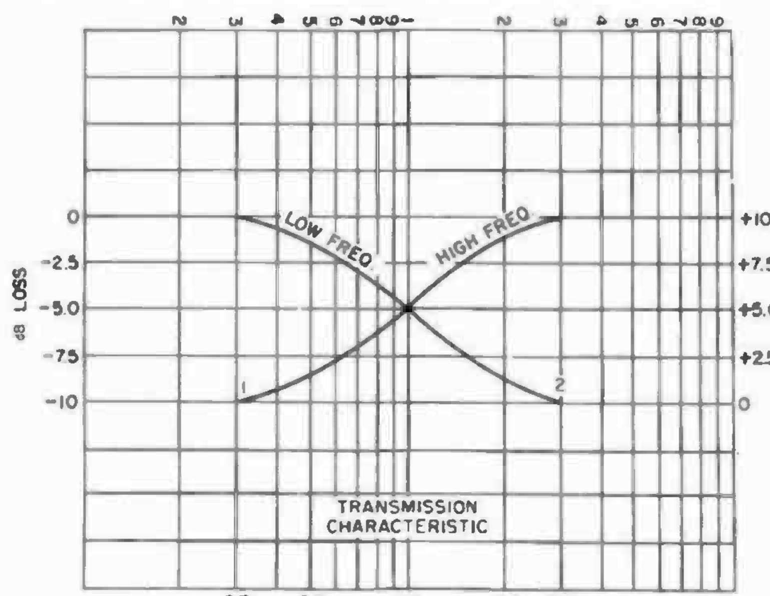


Fig. 6-51A. Equalizer design chart for the shelf-type equalizers given in Fig. 6-49A and Fig. 6-49B.

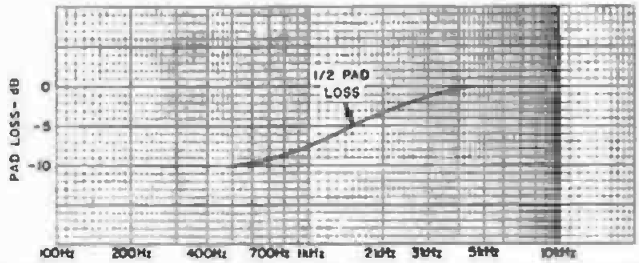


Fig. 6-51B. Frequency response for 1500-Hz, low-frequency, bridged-T, suppressor-type equalizer.

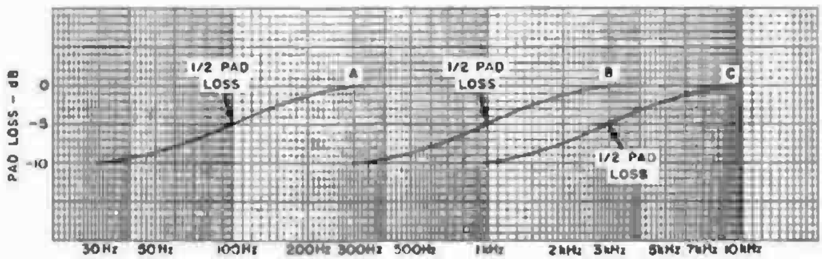


Fig. 6-51C. Frequency response for high-frequency suppressor-type bridged-T equalizer, using three different crossover frequencies.

Frequency of ½ Pad Loss (Hz)	*30*	*40*	*50*	*60*	*80*	*100*	*120*	*150*	*200*
L_1	2.64	1.98	1.58	1.32	0.990	0.792	0.658	0.531	0.396
C_1	7.33	5.50	4.42	3.66	2.750	2.210	1.84	1.470	1.100

Frequency of ½ Pad Loss (Hz)	*300*	*500*	*1k*	*1.5k*	*2k*	*2.5k*	*3k*	*4k*	*5k*
L_1	0.264	0.158	0.079	0.053	0.039	0.0316	0.0264	0.0198	0.0158
C_1	0.650	0.441	0.221	0.147	0.110	0.0884	0.0733	0.0550	0.0441

Frequency of ½ Pad Loss (Hz)	*6k*	*7k*	*8k*	*9k*	*10k*	*12k*	*15k*	*20k*
L_1	0.0132	0.0114	0.0098	0.0086	0.0079	0.0066	0.0052	0.0039
C_1	0.0366	0.0316	0.0275	0.0242	0.0221	0.0183	0.0147	0.0110

Fig. 6-51D. Component values for 600-ohm, shelf-suppressor equalizer of Fig. 6-49A.

are shown in Fig. 6-51C. It may be observed that all the curves have the same shape, and only the frequency of the one half pad loss has been shifted. The affected portion of the frequency spectrum extends from about one third of the half-loss point to three times the half-loss point.

After the frequency response has been plotted and if it is satisfactory, the values of C_1 and L_1 are selected from the table in Fig. 6-51D. Under the 1.5 kHz column, L_1 equals 0.053 henry, and C_1 is 0.147 μF. After having determined the circuit element values and the frequency characteristics, the configuration may be *converted* to an impedance other than 600 ohms and for a pad loss greater than 10 dB by the use of the simple *conversion factors* given in Questions 6.56 and 6.57.

6.52 *How is the equalizer configuration shown in Fig. 6-49B designed?*—It is designed in a manner similar to the example given in Question 6.51, except that *curve-2* of Fig. 6-51A is used. Assume that a low-frequency, shelf-type equalizer with a half-pad loss occurring

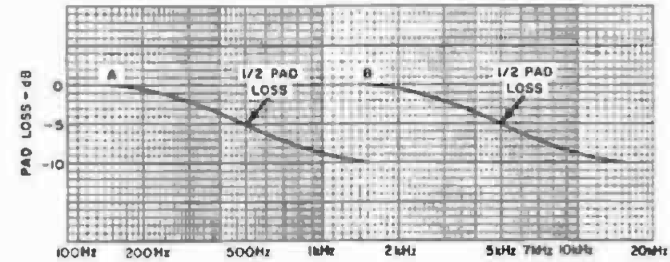


Fig. 6-52A. Frequency response for two low-frequency, suppressor-type bridged-T filters.

Frequency of ½ Pad Loss (Hz)	30	40	50	60	80	100	120	150	200
L_2	3.84	2.88	2.30	1.92	1.44	1.15	0.96	0.767	0.576
C_2	10.07	8.00	6.40	5.34	4.00	3.20	2.67	2.130	1.610

Frequency of ½ Pad Loss (Hz)	300	500	1k	1.5k	2k	2.5k	3k	4k	5k
L_2	0.384	0.231	0.115	0.0768	0.0576	0.0455	0.0384	0.0288	0.0231
C_2	1.065	0.640	0.320	0.2130	0.1610	0.1260	0.1060	0.0800	0.0640

Frequency of ½ Pad Loss (Hz)	6k	7k	8k	9k	10k	12k	15k	20k
L_2	0.0192	0.0168	0.0144	0.0132	0.0115	0.0096	0.0077	0.0058
C_2	0.0530	0.0460	0.0400	0.0367	0.0320	0.0267	0.0213	0.0161

Fig. 6-52B. Component values for 600-ohm, shelf type-suppressor equalizer in Fig. 6-49B.

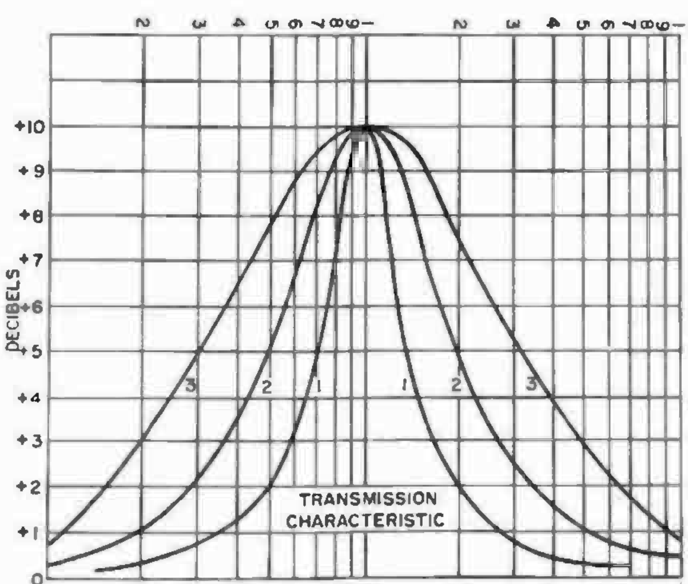


Fig. 6-53A. Equalizer design chart for equalizer shown in Fig. 6-50A.

at 5000 Hz is to be designed. Refer to curve 2, read the insertion loss at the left and plot the frequency characteristics as curve (B) in Fig. 6-52A. At 15,000 Hz (3.0), a loss of 10 dB is indicated: at 10.000 Hz (2.0), the loss is 8.75 dB; at 2500 Hz (0.5), it is 1.25 dB; and for 1500 Hz (0.30), the loss is zero. Similar to the high-frequency configuration, the frequency of one half the pad loss may be *shifted* to fit a particular requirement. The values for the reac-

Resonant Frequency	*30*			*40*			*50*			*60*			*80*		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_a	4.42	1.99	1.05	3.30	1.48	.778	2.660	1.190	.632	2.210	1.01	.537	1.650	.715	.398
C_a	6.38	14.10	26.80	4.84	10.80	20.700	3.800	8.470	16.10	3.200	7.00	13 400	2.420	5.540	10.000
L_b	1.60	3.53	6.70	1.20	2.78	5.080	.948	2.120	4.00	.800	1.75	3.350	.604	1.390	2.540
C_b	17.60	8.00	4.21	13.16	5.74	3.140	10.700	4.770	2.53	8.800	4.04	2.100	6.580	2.870	1.570

Resonant Frequency	*100*			*120*			*150*			*200*			*300*		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_a	1.290	.592	.316	1.100	.505	.265	.868	.396	.210	.666	.296	.158	.442	.199	.105
C_a	1.970	4.300	8.050	1.610	3.540	6.720	1.290	2.850	5.390	.965	2.140	4.030	.638	1.410	2.680
L_b	.487	1.070	2.010	.398	.875	1.680	.325	.712	1.350	.246	.533	1.010	.160	.353	.670
C_b	5.220	2.370	1.300	4.400	2.000	1.050	3.480	1.590	.835	2.620	1.180	.630	1.760	.800	.421

Resonant Frequency	*500*			*1,000*			*1,500*			*2,000*			*2,500*		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_a	.2660	.119	.0632	.129	.0592	.0316	.0868	.0396	.0210	.0666	.0296	.0158	.0532	.0236	.0127
C_a	.3800	.847	1.6000	.197	.4300	.8050	.1290	.2850	.5390	.0965	.2140	.4030	.0757	.1690	.3200
L_b	.0948	.212	.4000	.049	.1070	.2010	.0325	.0712	.1350	.0246	.0533	.1010	.0190	.0424	.0807
C_b	1.0700	.477	.2530	.522	.2370	.1300	.3480	.1590	.0840	.2620	.1180	.0630	.2140	.0960	.0506

Fig. 6-53B. Resonant equalizer component values tive elements are found under the 5000 Hz column in Fig. 6-52B. For purpose of illustration, a second curve half-pad loss occurring at 500 Hz has been plotted in Fig. 6-52A.

6.53 ***How is the equalizer configuration shown in Fig. 6-50A designed?***— By the use of the chart given in Fig. 6-53A. This chart is designed for a pad loss of 1 dB with a circuit impedance of 500 ohms. The left halves of the curves are for equalizers designed to

Resonant Frequency	3,000			4,000			5,000			6,000			7,000		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_4	.0442	.0199	.0105	.0330	.0148	.0078	.0266	.0118	.0063	.0221	.0101	.0053	.0188	.0085	.0044
C_4	.0644	.1410	.2680	.0482	.1080	.2020	.0380	.0850	.1600	.0322	.0700	.1340	.0275	.0610	.1180
L_1	.0160	.0350	.0670	.0121	.0267	.0504	.0095	.0212	.0400	.0080	.0175	.0335	.0069	.0152	.0296
C_1	.1760	.0800	.0420	.1310	.0590	.0320	.1070	.0480	.0253	.0880	.0400	.0210	.0750	.0340	.0180

Resonant Frequency	8,000			9,000			10,000			12,000			15,000		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_4	.0165	.0072	.0040	.0141	.0065	.0035	.0129	.0059	.0032	.0110	.0050	.0027	.0087	.00396	.0021
C_4	.0242	.0540	.1000	.0223	.0480	.0093	.0197	.0430	.0805	.0161	.0350	.0670	.0130	.02860	.0540
L_1	.0080	.0139	.0254	.0053	.0120	.0223	.0049	.0107	.0201	.0040	.0087	.0170	.0033	.00712	.0135
C_1	.0660	.0290	.0160	.0590	.0260	.0140	.0520	.0240	.0130	.0440	.0200	.0110	.0348	.01590	.0085

All Values of Inductance in Henries
All Values of Capacitance in μF

for 500-ohm equalizer shown in Fig. 6-50A.

increase the high-frequency response. The right side is used for increasing the low-frequency response. The complete curve is used when designing a resonant or peaked equalizer. To illustrate how the chart functions, assume an equalizer is required for increasing the high-frequency response at 10,000 Hz 10 dB, for a circuit impedance of 500 ohms.

Referring to the chart in Fig. 6-53A, three sets of transmission curves are given. Because this is to be a high-frequency equalizer, a curve is selected from the left-hand group that will approximate the desired response, with the point (1.0) representing the resonant frequency of 10,000 Hz. The ratio of other frequencies to the resonant frequency may now be plotted.

Assuming the desired frequency response approximates curve 2 at the left, 1000 Hz (0.1) will be up 0.4 dB; 2000 Hz (0.20), up 1 dB; 5000 Hz (0.5); 5 dB, 7000 Hz (0.07), 8 dB; and 10,000 Hz, 10 dB.

While it is true the curves in Fig. 6-53A will not satisfy all requirements, they will be sufficiently close for all practical purposes.

After plotting the frequency response, the circuit element values are selected from the table in Fig. 6-53B under the resonant frequency of 10,000 Hz. For the above example, these values are: L_a 5.9 mH, L_1 10.7 mH, C_3 0.043 μF, and C_1 0.024 μF. The pad loss is 10 dB.

After the unit had been completed and transmission measurements made, the values of the capacitors may be shifted to bring the curve closer to requirements. However, unless extreme accuracy is desired, the indicated values will be close enough.

6.54 How is a low-frequency equalizer similar to the one shown in Fig. 6-50A designed?—By using the same procedure as described for the high-frequency equalizer, with one exception. The right-hand group of curves in Fig. 6-53A is used. The circuit elements are selected from the table in Fig. 6-53B.

6.55 How is the equalizer shown in Fig. 6-50B designed?—By the use of the chart in Fig. 6-55A and using the procedure described in Question 6.53. However, as this equalizer has characteristics reversed from those in Fig. 6-50, the frequency response is plotted as a loss, as the curves approach resonance. The circuit elements are selected from the table in Fig. 6-55B. The left-hand group of curves is used for high-frequency attenuation, and the right-hand group for low-frequency

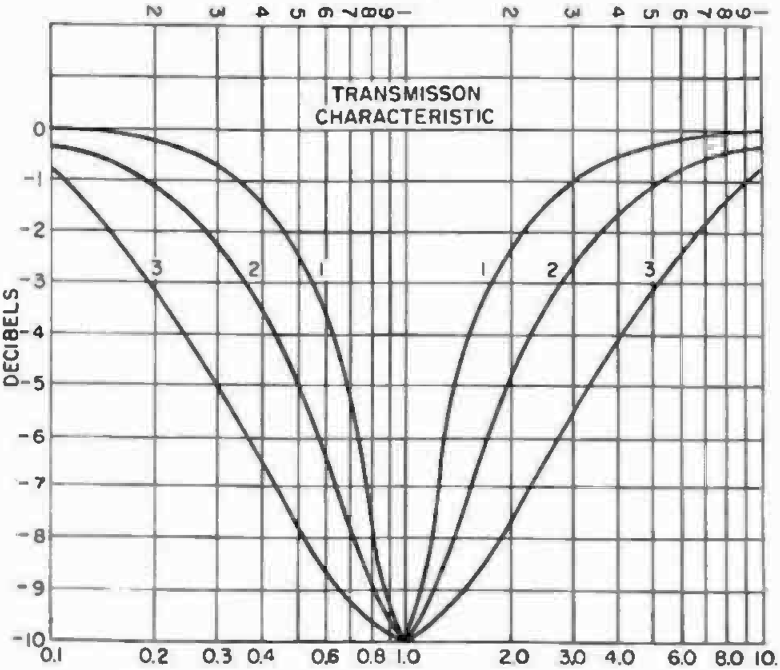


Fig. 6-55A. Equalizer design chart for equalizer shown in Fig. 6-50B.

attenuation. For a dip-type equalizer, the complete curve is used.

6.56 ***How are the foregoing equalizers converted to impedances other than 500 or 600 ohms?***—All of the foregoing design data have been based on either a 500- or 600-ohm circuit impedance. If the equalizer is to be used in a circuit impedance other than that given, *new* circuit component values are required. In this situation the equalizer network is designed for the impedance given,

Resonant Frequency	30			40			50			60			80		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_3	2.30	5.08	9.680	1.74	4.000	7.300	1.40	3.100	5.75	1.15	2.520	4.840	.870	2.000	3.650
C_3	12.20	5.55	2.900	9.14	4.100	2.280	7.40	3.280	1.70	6.15	2.800	1.500	4.570	1.990	1.110
L_4	3.07	1.39	.730	2.28	.100	.550	1.85	.828	.438	1.53	.700	.370	1.140	.497	.273
C_4	9.20	20.30	38.600	6.96	16.400	29.200	5.60	12.200	23.00	4.60	10.100	19.340	3.480	8.000	14.600

Resonant Frequency	100			120			150			200			300		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_3	.702	1.550	2.890	.573	1.270	2.420	.460	1.030	1.940	.354	.767	1.450	.230	.508	.965
C_3	3.620	1.640	.878	3.070	1.380	.750	2.420	1.100	.582	1.800	.829	.439	1.220	.556	.290
L_4	.900	.410	.219	.767	.343	.182	.603	.276	.146	.448	.208	.110	.307	.139	.073
C_4	2.810	6.200	11.600	2.290	5.100	9.670	1.870	4.100	7.750	1.420	3.060	5.800	.920	2.030	3.860

Resonant Frequency	500			1,000			1,500			2,000			2,500		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_3	.137	.3050	.575	.0702	.155	.2890	.0468	.1030	.1940	.0354	.0767	.145	.0273	.0617	.1160
C_3	.738	.3320	.178	.3620	.164	.0880	.2420	.1100	.0580	.1800	.0830	.044	.1490	.0660	.0350
L_4	.185	.0828	.044	.0903	.041	.0219	.0603	.0276	.0146	.0448	.0208	.011	.0374	.0165	.0088
C_4	.548	1.2200	2.300	.2810	.620	1.1600	.1870	.4100	.7750	.1420	.3060	.580	.1100	.2470	.4640

Fig. 6-55B. Component values for 500-ohm resonant equalizer shown in Fig. 6-50B **(continued on pg. 282).**

then converted to the required impedance by means of the conversion factors given in Fig. 6-56. It will be noted in some instances the component values are multiplied, and in others they are divided.

The maximum attenuation (15 to 20 dB) depends largely on the "Q" of the coil used. When the networks shown are used with a variable attenuator, the component values are selected for a pad-loss of the *maximum* value to be

Resonant Frequency	3,000			4,000			5,000			6,000			7,000		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_a	.0230	.0508	.0965	.0174	.0385	.0725	.0137	.0305	.0575	.0115	.0252	.0483	.0099	.0219	.0426
C_a	.1220	.0560	.0300	.0910	.0410	.0220	.0740	.0330	.0180	.0620	.0280	.0150	.0520	.0240	.0120
L_b	.0307	.0140	.0073	.0227	.0103	.0055	.0185	.0083	.0044	.0153	.0070	.0037	.0130	.0059	.0030
C_b	.0920	.2030	.3860	.0700	.1540	.2900	.0550	.1220	.2300	.0460	.1000	.1930	.0400	.0880	.1710

Resonant Frequency	8,000			9,000			10,000			12,000			15,000		
Response	1	2	3	1	2	3	1	2	3	1	2	3	1	2	3
L_a	.0087	.0200	.0363	.0070	.0173	.0322	.0072	.0155	.0289	.0058	.0126	.0250	.0047	.0103	.0194
C_a	.0460	.0200	.0110	.0410	.0180	.0100	.0360	.0160	.0090	.0310	.0140	.0075	.0242	.0110	.0058
L_b	.0114	.0050	.0027	.0103	.0045	.0024	.0090	.0041	.0022	.0077	.0035	.0018	.0060	.0028	.0015
C_b	.0350	.0800	.1460	.0310	.0690	.1290	.0280	.0620	.1160	.0230	.0500	.0970	.0187	.0410	.0775

All Values of Inductance in Henries
All Values of Capacitance in μF

Fig. 6-55B (continued). Component values for 500-ohm resonant equalizer shown in Fig. 6-50B.

500 ohms		600 ohms	
$K = \frac{Z_r}{500}$	K	$K = \frac{Z_r}{600}$	K
$\frac{600}{500}$	1.20	$\frac{500}{600}$	0.833
$\frac{500}{500}$	1.00	$\frac{600}{600}$	1.000
$\frac{250}{500}$	0.50	$\frac{250}{600}$	0.416
$\frac{200}{500}$	0.40	$\frac{200}{600}$	0.333
$\frac{150}{500}$	0.30	$\frac{150}{600}$	0.250

Multiply: C, C, L, L, L, and L.
Divide: C, C, C, C, L, and L.
Z_r = *New Impedance.*

Fig. 6-56. "K" factors for converting equalizer networks to impedances other than their original design impedances.

used. For situations less than maximum, the curves will approximate the same general shape. Higher values of attenuation will cause the curve to become steeper, with the break at the no-loss point remaining around the same point. This will move the design point of the half-pad loss closer to the break from the no-loss portion at the higher values of attenuation, and vice versa. (See Questions 6.61, 6.67, and 6.124.)

6.57 ***How are the foregoing equalizers converted to other pad losses?***—When pad losses other than 10 dB are required, a change in the circuit element values is necessary. The new values are obtained by multiplying or dividing the values given for the 10 dB, 500-ohm equalizer by the factors given in Fig. 6-57.

6.58 ***What is a balanced equalizer?***—One in which the circuit elements are connected in each side of the line in a manner similar to a balanced attenuator. Four such configurations are shown in Figs. 6-58A, B, C, and D. The configurations of Figs. 6-58A and B are derived from those shown in Figs. 6-49A and B, while those in Figs. 6-58C and D are derived from the configurations of Figs. 6-50A and B. The terminating circuits must be balanced or be symmetrical to ground.

6.59 ***Is it permissible to connect a balanced and an unbalanced equalizer***

dB Loss	K	dB Loss	K	dB Loss	K
3	0.30	9	0.89	15	1.62
4	0.38	10	1.00	16	1.75
5	0.48	11	1.12	17	1.87
6	0.57	12	1.25	18	2.00
7	0.67	13	1.38	19	2.12
8	0.78	14	1.50	20	2.25

Multiply L, L, L, C, C, and C.
Divide L, L, L, C, C, and C.

Fig. 6-57. "K" factors for pad losses other than 10 dB.

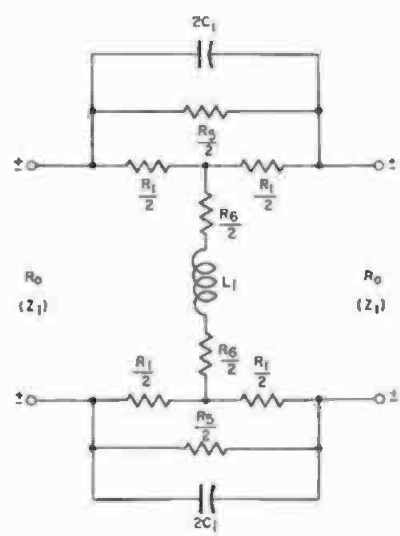


Fig. 6-58A. Balanced configuration, derived from Fig. 6-49A, showing the division of circuit elements.

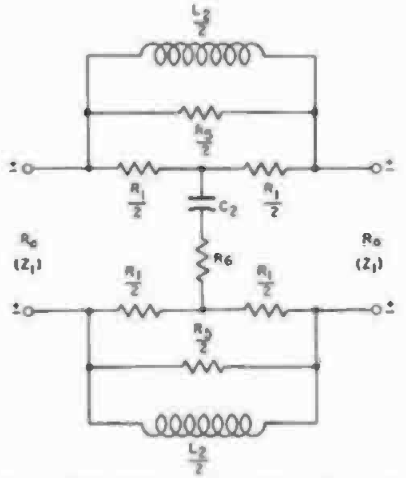


Fig. 6-58B. Balanced configuration, derived from Fig. 6-49B, showing the division of circuit elements.

in tandem?—No. They cannot be directly connected, but must be isolated from each other by a repeat coil as described for attenuators in Question 5.30. Otherwise, serious leakage will occur at freqnencies above 2000 Hz.

6.60 ***What is the designed procedure for a balanced equalizer?***—Balanced equalizers are first designed as an unbalanced configuration and then converted to a balanced configuration, as shown in Figs. 6-58A, B, C, and D.

6.61 ***How is the attenuator loss selected for a variable equalizer?***—The

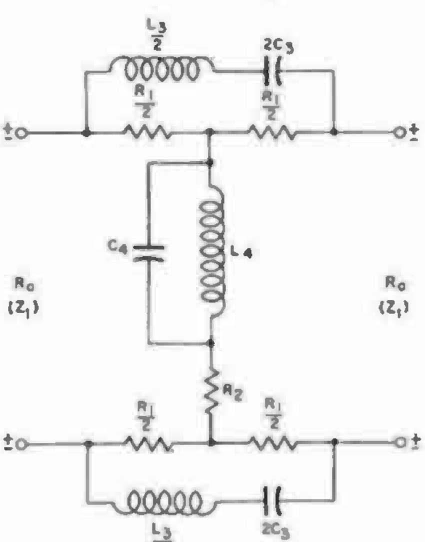


Fig. 6-58C. Balanced configuration, derived from Fig. 6-50A, showing the division of circuit elements.

attenuator loss is selected for the desired maximum value. The transmission curves of the equalizer will remain essentially the same for all loss settings of the attenuator up to 10 dB. For greater values of loss, the curves will become steeper.

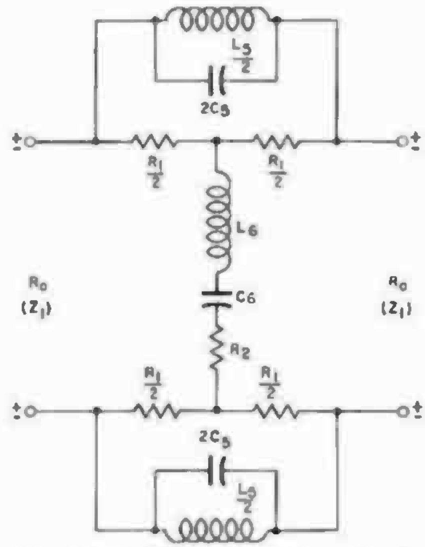


Fig. 6-58D. Balanced configuration derived from Fig. 6-50B, showing the division of circuit elements.

Variable attenuators for use with constant-resistance type equalizers as a rule employ the hridged-T attenuator. However, the plain-T may also be used, except that more noise may be encountered when changing the attenuation becanse of the additional contacts and also because of its configuration. The bridged-T attenuator is less noisy when moved, because of the loading effect of the fixed resistors R_1 in the upper side of the configuration. Generally, the attenuator is so designed that it will increase the loss in steps of 1 dB.

6.62 ***Is a ground connection required with an unbalanced equalizer?***—Yes. If the equalizer is not grounded, leakage will occur at the higher frequencies.

6.63 ***Where is the ground connected to an unbalanced equalizer?***—To the side which has no circuit elements, similar to an unbalanced attenuator.

6.64 ***Is it necessary to ground a balanced equalizer?***—No. If the source and terminating impedances are balanced with respect to ground no equalizer ground is necessary.

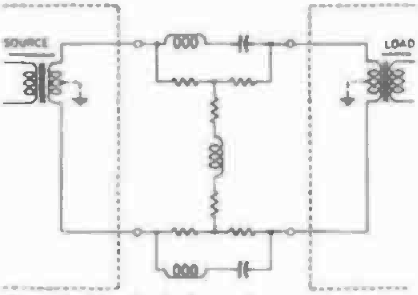


Fig. 6-65. A balanced equalizer connected in a balanced-to-ground source or load circuit. The circuit may be grounded at both the input and output or at one end only.

6.65 Where is a ground connected to a balanced equalizer?—If the circuit is not balanced to ground, the ground is connected to the exact electrical center of the configuration. Because this may be impractical in some configurations, the source and/or terminating circuits are balanced to ground as shown in Fig. 6-65.

6.66 If two equalizers are connected in tandem, what is the overall frequency characteristic?—It is the algebraic sum of the two characteristics as shown in Fig. 6-66. Curves 1 and 2 are the characteristics of the individual equalizers. Curve 3 is the overall characteristic.

6.67 If the frequency characteristics and component values of an equalizer are known, can it be converted to a different frequency?—Yes, by changing the reactive component values inversely with respect to frequency. As an example, an equalizer with a resonant frequency of 1000 Hz may be changed to a resonant frequency of 100 Hz by multiplying the reactive elements by a factor of 10. If this same equalizer were to be converted to 5000 Hz, the circuit elements would be divided by a factor of five. The impedance would remain the same as it was for the original design.

6.68 What type coils are recommended for equalizer construction?—Coils of high Q, such as toroidal-wound coils. The efficiency of a toroidal coil is extremely high compared to other designs. Also, they are not affected by external magnetic fields and may be mounted alongside each other without fear of coupling.

6.69 At what frequency is the maximum Q specified for equalizer coils?—At the resonant frequency. (See Question 8.72.)

6.70 How are toroidal coils constructed?—They are wound on a circular core composed of molybdenum permalloy dust mixed with a plastic binder. The core is compressed into the shape of a doughnut under a pressure of 200 tons per square inch. A typical toroidal coil is shown in Fig. 6-70. By the use of this type winding, coils with a Q of several hundred are possible.

6.71 What is the turnover frequency of a toroid coil?—It is the frequency at which the Q stops rising with frequency and starts downward, as shown by the curves in Fig. 6-71.

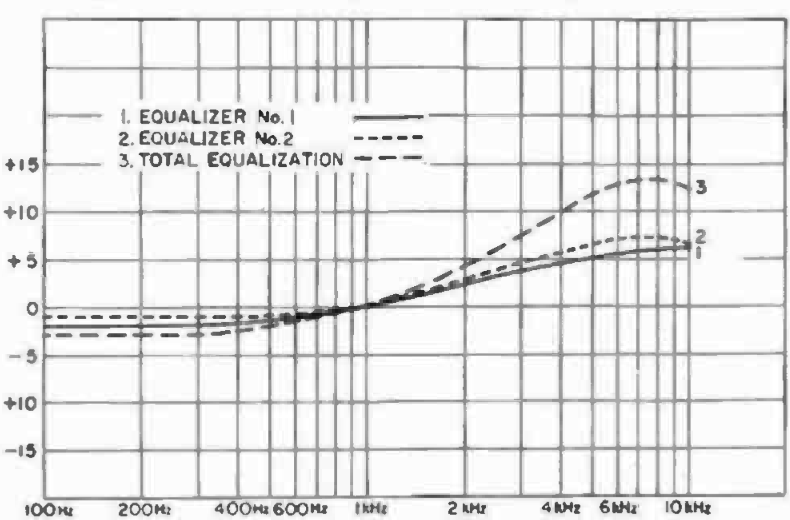


Fig. 6-66. Frequency characteristic of two equalizers connected in tandem.

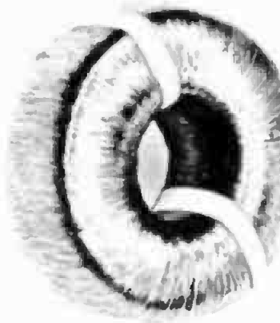


Fig. 6-70. A toroidal-wound coil without the insulating cover.

6.72 ***What is a honeycomb-wound coil?***—An air-core coil wound in such a manner the turns cross over each other and appear as a honeycomb. This design results in low distributed capacity but a low Q.

6.73 ***What is a negative-feedback loop equalizer?***—An equalizer network composed of reactive elements connected in the feedback loop of a negative-feedback amplifier. The equalization may be fixed or variable. However, variable equalization in a feedback loop is not desirable because each time the equalization is changed, the feedback benefits are changed, as is the gain of the amplifier. Typical negative-feedback loop-equalizer circuits are shown in Fig. 6-73. The subject of negative feedback is discussed in Questions 12.136 through 12.168.

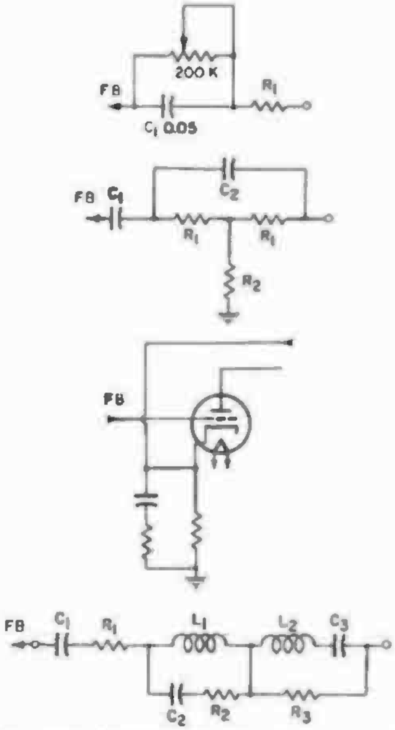


Fig. 6-73. Negative-feedback loop equalizers.

6.74 ***What is a plate-circuit equalizer?***—Reactive elements connected in the plate circuit of an amplifier to produce a given frequency characteristic. A typical circuit is shown in Fig. 6-74. This circuit may be used for securing a low-frequency boost by the change of

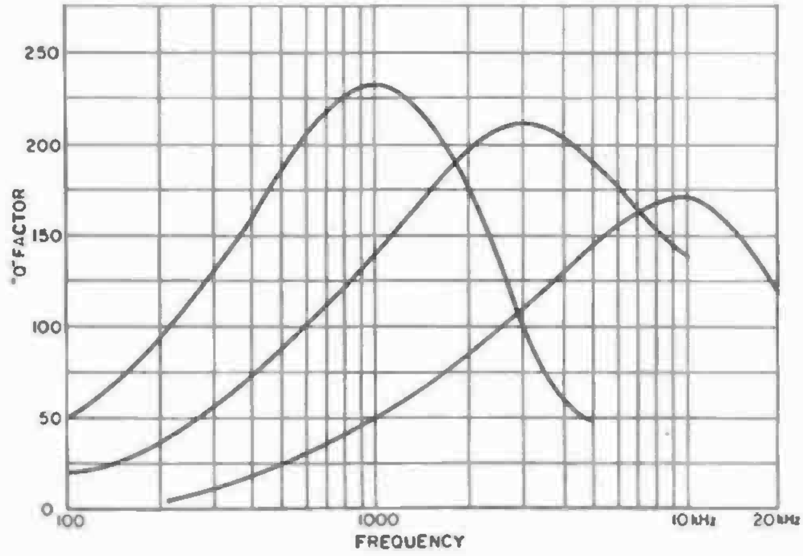


Fig. 6-71. Toroidal coil "Q" versus frequency curves.

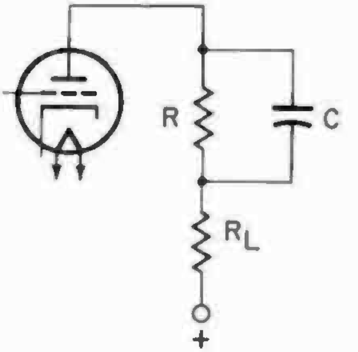


Fig. 6-74. Low frequency plate-circuit equalizer.

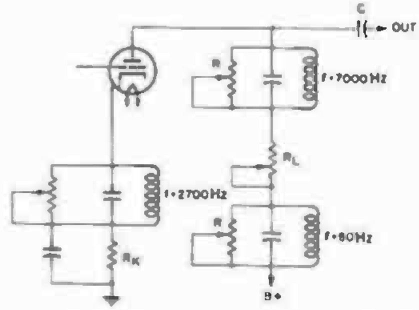


Fig. 6-76. Cathode- and plate-circuit equalizers.

reactance of the capacitor C connected across the load resistor R. The capacitor is selected for a value that will attenuate the high frequencies, beginning at a predetermined frequency. As the frequency is decreased, the reactance of the capacitor increases permitting resistor R to become effective in the circuit. In this way, the gain of the stage is increased at the lower frequencies.

6.75 ***When is a circuit said to be at resonance?***—When the inductive reactance equals the capacitive reactance.

6.76 ***What is a resonant plate-circuit equalizer?***—A vacuum-tube amplifier, with resonant circuits in the plate and cathode, used for equalizing in fixed-frequency bands. A typical circuit is shown in Fig. 6-76. The circuits are designed for a high Q at the resonant frequency. Variable resistors (R) are shunted across the coils to control the amount of equalization. The plate load resistor R_L will vary depending on the type tube used. When the shunt resistors in parallel with the resonant circuits are varied, it will be necessary to change the load resistance R_L to compensate for the loss induced by the networks. As a rule, this type of equalization is only used with triode vacuum tubes.

6.77 ***What tolerance should be specified for equalizer coils?***—Plus or minus two percent.

6.78 ***What tolerance values should be specified for capacitors used in equalizer design?***—Plus or minus three percent. However, if greater accuracy is required, they may be adjusted to the exact value by paralleling. In practice, the equalizer is first completed and measured. The capacitors are then tailored to obtain the desired response.

6.79 ***What tolerance should be specified for resistors used in equalizer design?***—Plus or minus five percent. However, if wirewound resistors are used, they are generally specified plus or minus one percent.

6.80 ***How is low-frequency attenuation achieved in a mixer circuit?***—By connecting a capacitor in series with the hot side of the mixer-input circuit as shown in Fig. 6-80A. This method of obtaining low-frequency attenuation is simple and effective and is often used

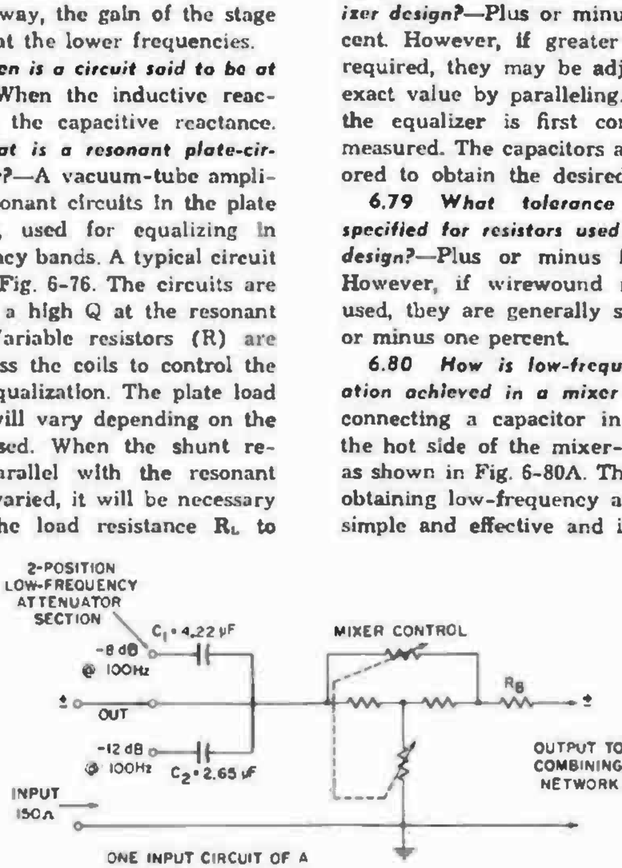


Fig. 6-80A. Low-frequency attenuation in a mixer network achieved by means of series capacitors.

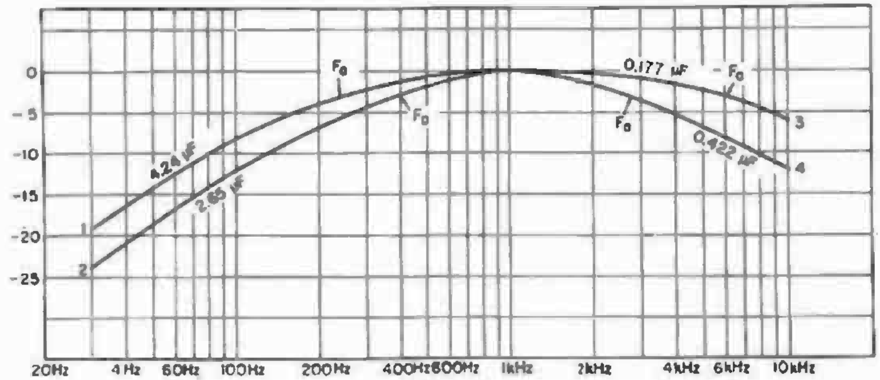


Fig. 6-80B. Typical low- and high-frequency attenuation curves.

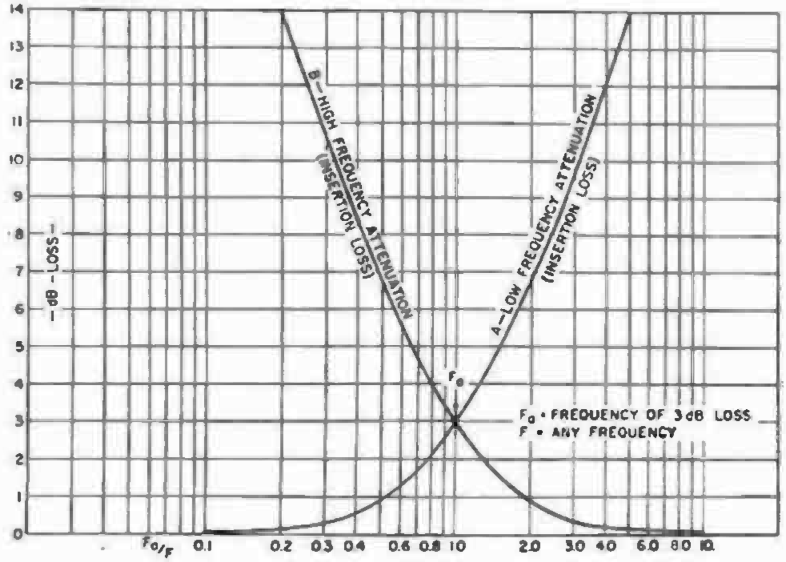


Fig. 6-80C. Design graph for attenuation and L-type equalizers.

in dialogue recording mixers. Generally, two positions of low-frequency attenuation are provided, either one of which may be cut in or out of the circuit by means of a key switch.

The value of the capacitor for a given frequency characteristic is dependent on two factors—the frequency of F_a (the frequency of 3-dB attenuation), and the impedance of the mixer-input circuit. The value of the capacitor may be calculated:

$$C = \frac{10^6}{2\pi F_a R_a}$$

where,

F_a is the frequency of 3-dB attenuation,
R_a is the circuit impedance.
C is the capacitance in microfarads.

To illustrate the design procedure, Fig. 6-80B shows typical low-frequency attenuation curves used in the motion picture industry. Assume a frequency response similar to that of curve 1 is desired. For this characteristic, 100 Hz is to be down 8 dB with respect to 1000 Hz; therefore, a frequency of 250 Hz is selected for F_a. The circuit impedance is 150 ohms. The capacitor is equal to:

$$\frac{10^6}{6.28 \times 250 \times 150} = \frac{10^6}{235{,}500} = 4.24\ \mu F$$

After calculating the value of C, its insertion loss is plotted with respect to frequency, with the aid of the right-hand curve (A) in Fig. 6-80C. Assuming F_a is 250 Hz and is represented by 1.0 on the graph, 250 Hz will have a loss of 3 dB. The response at other frequencies is calculated by dividing F_a by F for each frequency and tabulated as shown in Fig. 6-80D as F_a/F. The

F	$\frac{F_o}{f}$	dB loss
1000 Hz	0.10	− 0.10
500 Hz	0.50	− 0.90
400 Hz	0.62	− 1.40
300 Hz	0.83	− 2.20
250 Hz	1.00	− 3.00
150 Hz	1.66	− 5.70
100 Hz	2.50	− 8.24
80 Hz	3.13	− 10.00
60 Hz	4.16	− 12.40
50 Hz	5.00	− 14.00

Fig. 6-80D. Insertion-loss plot of curve 1, Fig. 6-80B.

loss for each frequency is obtained by entering the graph along the bottom for different values of F_o/F and then reading the loss where F_o/F intersects the right-hand curve (A). The overall frequency response is plotted as shown in Fig. 6-80B.

If the frequency response does not fit a particular response curve, the frequency F_o is shifted to bring the response to the desired shape. It should be kept in mind that such circuits are used only for a slow roll-off type attenuation. If a faster and steeper response is desired, a more elaborate type circuit will be necessary.

Equalizers employing series capacitors in the input circuit are shown in Fig. 6-80A. Although not of constant impedance design they are a simple and economical method of reducing the low frequency response for dialogue recording. However, they should only be employed when the preamplifier offers a solid termination, such as one nsing building-out resistors, or negative feedback, or a combination of both. Because of the solid termination of the preamplifier and the resistive uetwork of the

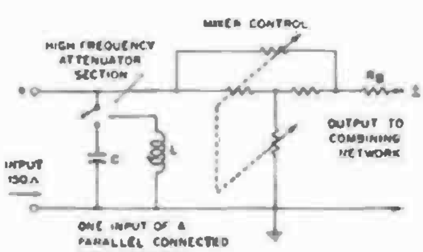


Fig. 6-81. A high-frequency attenuator circuit for a mixer network. Either an inductance or capacitor may be used.

mixer, this method of equalization will operate quite satisfactorily in circuit impedances of 150, 250 and 600 ohms. (See Questions 6.131 and 9.24.)

6.81 ***How is high-frequency attenuation achieved in a mixer circuit?***—By connecting either a capacitor or inductance in parallel with the mixer input circuit, as shown in Fig. 6-81.

In Fig. 6-80B, two high-frequency rolloffs are shown. If a capacitor is to be used, its value is calculated as described in Question 6.80. The frequency response is plotted using the left-hand curve (B) of Fig. 6-80C. If an inductance is used, its value may be calculated:

$$L = \frac{R_o}{2\pi F_o}$$

where,

F_o is the frequency of 3-dB loss,
R_o is the circuit impedance.

The procedure for plotting the insertion loss is the same as described for a capacitor.

6.82 ***Can low-frequency attenuation be obtained using an inductance?***—Yes. After calculating the inductance, the insertion loss is plotted as for a capacitor. Assume a frequency response is desired similar to curve 2 of Fig. 6-80B. The value of the inductance may be calculated:

$$\frac{150}{6.28 \times 400} = \frac{150}{2512} = 0.0597 \text{ henry}$$

where,

F_o equals 400 Hz,
R_o equals 150 ohms.

6.83 ***How is a high-frequency attenuator designed using an inductance?***—The inductance is calculated as described in Question 6.82. The insertion loss is plotted using the left-hand curve (B) of Fig. 6-80C.

6.84 ***How may the frequency response be shifted to meet a given requirement?***—By changing the frequency of F_o, the frequency of 3-dB attenuation.

6.85 ***What is a compensated volume control (loudness control)?***—A volume control used in amplifiers to compensate for the human ear characteristic when the volume level is increased or decreased. The characteristic of the control is based on the well-known Fletcher-Munson or equal-loudness contours. This subject is further discussed in Questiou 5.65.

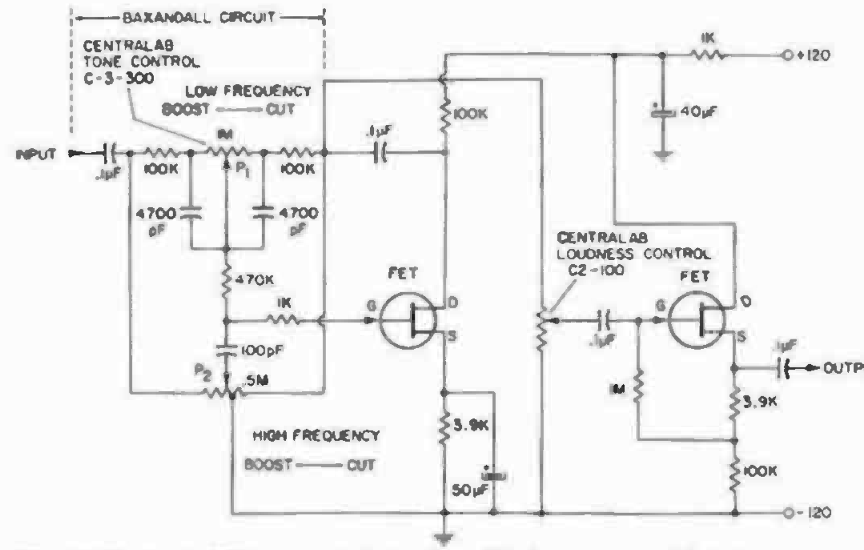


Fig. 6-86A. Baxandall tone-control circuit modified to operate with field effect transistors. (Courtesy, Wireless World)

6.86 *Describe the Baxandall equalizer circuit (tone control)*—The circuit shown in Fig. 6-86A was first described by P. J. Baxandall in 1952 in Wireless World (English publication), and is the basis for many tone controls used in high fidelity sound reproducing equipment. The basic circuit consists of two linear potentiometers, three resistors, and three capacitors. The circuit is generally connected between the plate and grid in vacuum-tube amplifiers, and in transistor circuits, between the emitter and base of the following transistor.

In Fig. 6-86A the circuit is connected between the source and gate elements of two field-effect transistors (FET's). With potentiometers P_1 and P_2 set to their center positions, the circuit has a flat frequency response. Moving the upper control to the *left* increases the low-frequency response, while moving it to the *right* decreases the low-frequency response.

It will be observed, the potentiometers of Fig. 6-86B are of higher value than ordinarily used in transistor circuits (suitable for vacuum tubes also). The equalization at 20 Hz is about plus or minus 18 dB, and for the high frequency control plus and minus 16 dB with reference to 1000 Hz. When the low-frequency control is set for a 10 dB boost or cut, 125 Hz is plus or minus

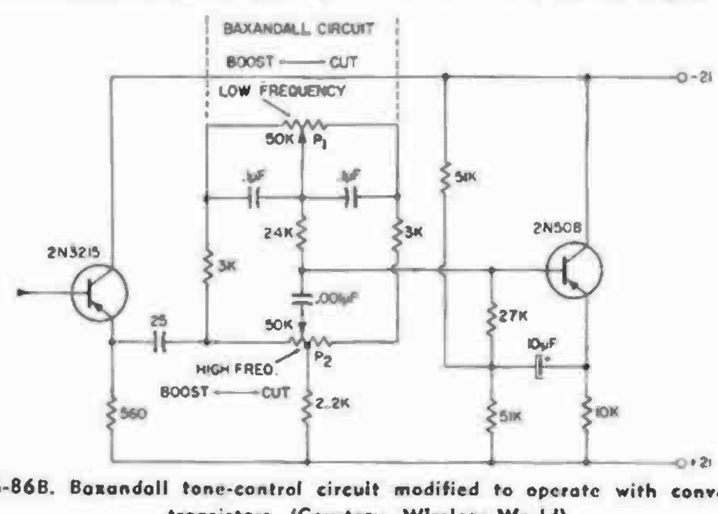


Fig. 6-86B. Baxandall tone-control circuit modified to operate with conventional transistors. (Courtesy, Wireless World)

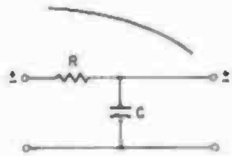


Fig. 6-88. A simple low-pass filter RC equalizer.

3 dB. With the high-frequency control set to a maximum, the boost or cut at 3500 Hz is also 3 dB.

6.87 ***What does the term resonant frequency mean?***—It is the frequency at which the inductive and capacitive reactances in a tuned circuit are equal.

6.88 ***How is a simple low-pass filter characteristic obtained with an RC equalizer?***—As shown in Fig. 6-88. The frequency response is rather broad for a single network; however, the response may be made steeper by cascading two or more sections. When a frequency is reached where the capacitive reactance is one-tenth its original value, little change with frequency is noted. The insertion loss of an RC circuit is rather high; therefore, it must be used with an amplifier. The resistor must be non-inductive.

6.89 ***Show an RC circuit for boosting the low-frequency response.***—Such a circuit is shown in Fig. 6-89. The input signal is applied to terminals 1 and 2, and the output is taken across terminals 3 and 4. With the proper selection of resistance and capacity, the impedance of the capacitor will increase with a decrease of frequency. Thus, the capacitor has less shunting effect across the resistauce R2 and, thereby, the gain at the lower frequencies is increased.

6.90 ***How is a simple high-pass filter characteristic obtained with an RC equalizer?***—As shown in Fig. 6-90.

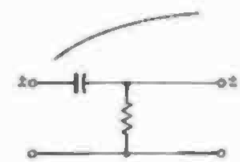


Fig. 6-90. A simple high-pass filter RC equalizer.

6.91 ***Show a simple grid-circuit equalizer, similar to that used in preamplifiers.***—The circuit shown in Fig. 6-91A is rather common in phototube amplifiers. It will be noted the equalizer circuit, C1 and R1, is connected in the grid circuit after the regular grid coupling capacitor C. Generally, the capacitor C1 is of the order of 25 to 150 picofarads. The resistor R1 will range be-

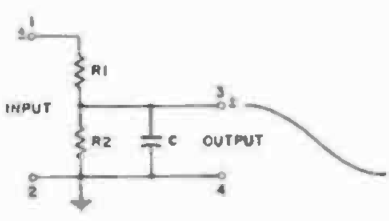


Fig. 6-89. An RC low-frequency boost circuit.

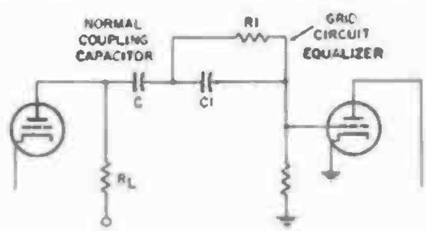


Fig. 6-91A. An RC grid-circuit equalizer.

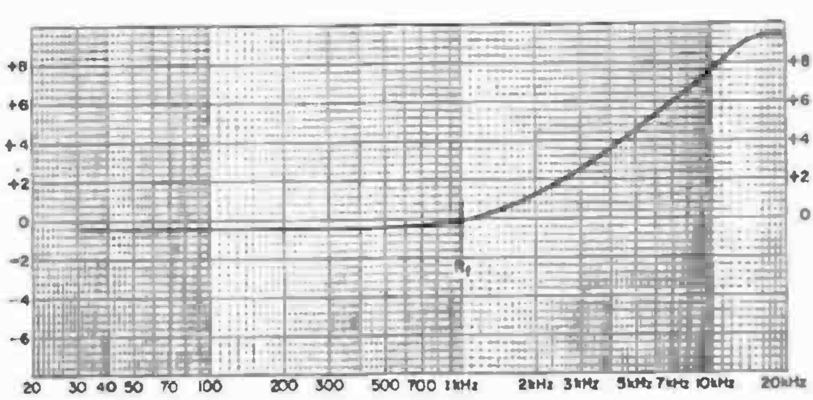


Fig. 6-91B. Frequency response of the grid-circuit equalizer in Fig. 6-91A.

tween 250,000 ohms and 1 megohm in value. The take-off point will depend on the value of C1. The amount of equalization and shape of the curve are controlled by the value of R1.

Because of the insertion loss of the equalizer, the overall gain of the amplifier is reduced at the reference frequency; therefore, the amplifier should be designed with sufficient gain to compensate for the equalizer loss and sufficient output level to prevent overloading at the maximum frequency of equalization, 15,000 Hz. The frequency response for a grid-circuit equalizer using a 65-pF capacitor and 1.25-megohm resistor is shown in Fig. 6-91B.

6.92 ***What frequency is used for the reference in equalizer design?***—Generally, 1000 Hz; however, 400 Hz is also used.

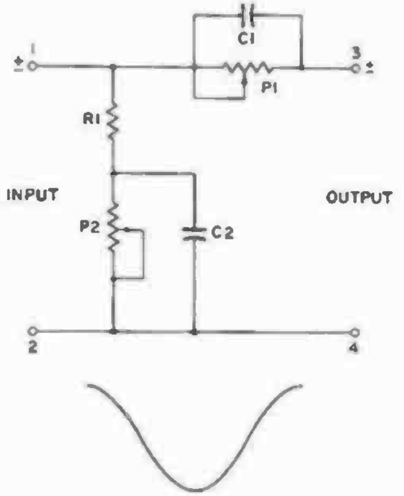


Fig. 6-93. An RC high- and low-frequency equalizer.

6.93 ***Show an RC configuration that may be used for both high- and low-frequency correction.***—Such a circuit is shown in Fig. 6-93, with its frequency characteristic given at the right. The amount of frequency correction is obtained by the position of the two variable pots, P1 and P2. In the mid-frequency range, the circuit is practically resistive, consisting principally of the two variable pots, and resistor R1. The capacitor C1 is practically an open circuit because of its high reactance, while the low reactance of C2 at these frequencies causes it to act as a short circuit across P2. Thus, only P1 and R1 are effectively in the circuit.

As the frequency increases, the reactance of C1 decreases until, finally, at some frequency, its reactance becomes so small as to effectively short circuit P1. Conversely, as the frequency becomes lower, the reactance of C1 increases until only the resistance which is constant for all frequencies remains in the circuit. However, the shunt reactance of C2 increases with the decrease in frequency, leaving only resistance, which results in a higher voltage across the output terminals 3 and 4. Therefore, the equalization depends on the setting of P1 and P2.

6.94 ***Show an amplifier with RC equalizers for increasing or decreasing the high- and low-frequency response.***—Fig. 6-94 shows a two-stage RC coupled amplifier, with variable RC equalizers for increasing or decreasing the high- and low-frequency response. To compensate for the insertion loss of the RC circuits, two pentode amplifier stages are necessary. The variable controls P1 and P2 are standard, center-tapped, audio-taper controls. This amplifier is designed to be inserted between the output and input of two triode amplifier stages. The frequency response is uniform when the pots P1 and P2 are in their midpositions. The overall gain of the amplifier will change with changes of equalization; therefore, the gain control P3 will require resetting to maintain the same gain through the amplifier. About 14 dB of equalization is available at either end of the spectrum. This, together with 14 dB of attenuation at either end, makes a total of approximately 28 dB of equalization or attenuation available.

6.95 ***What is the procedure for designing an L-type equalizer?***—L-type equalizers consist, basically, of two configurations as shown in Figs. 6-95A and B. Although the L-type equalizer presents a constant impedance only in one direction (terminals 1 and 2), it is used in many applications where the circuit is not impedance sensitive. The impedance at terminals 3 and 4 is subject to wide variation, similar to the L-type attenuators described in Questions 5.40 and 5.41.

The design of an L-type equalizer is started by first selecting a design point called F_r which is the frequency of 3 dB insertion loss. F_r is the frequency where both the reactive elements L and

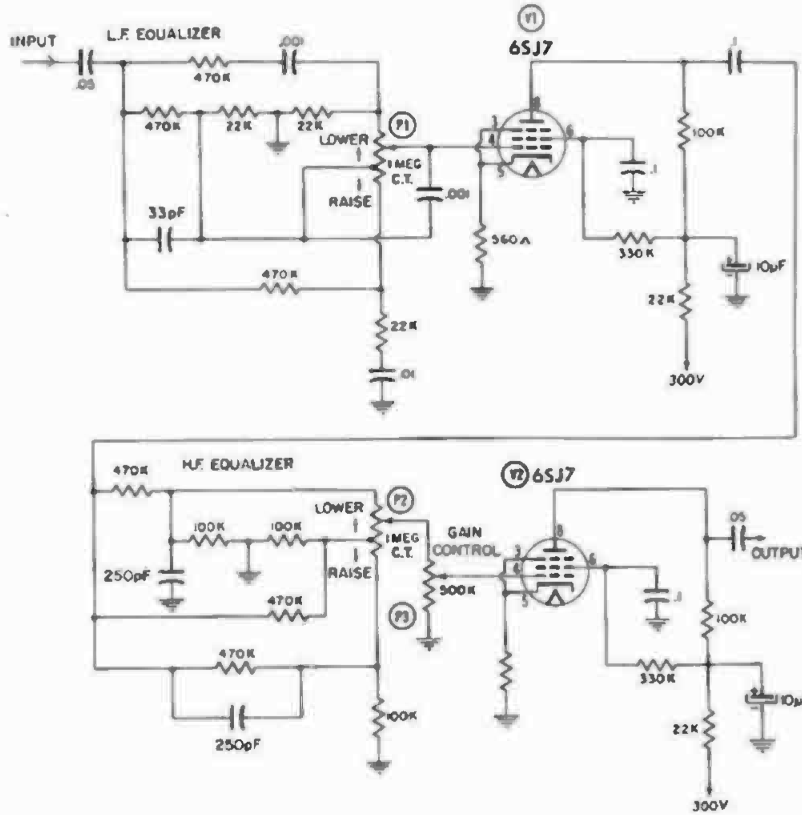


Fig. 6-94. An amplifier with high- and low-frequency variable equalization.

C have the same reactance and equal the line impedance R_o. Knowing the 3 dB insertion-loss frequency, the frequency response for a given value of F_c may be plotted by the aid of the graphs in Fig. 6-80C.

Assume, for a given frequency response, that F_c is equal to 2000 Hz. Referring to curve A of Fig. 6-80C, frequency F_c is represented by the figure 1.0 at the bottom of the graph. This point represents the 3 dB insertion loss point. Therefore, 2000 Hz will be down 3 dB. The figure 2.0 on the chart will then represent 4000 Hz and 4.0, 8000 Hz. Figure 0.5 represents 1000 Hz and 0.25, 500 Hz. The response at any other frequency may be calculated by dividing F_c by F, F being any frequency. The loss for each frequency of interest is obtained by entering the graph at the bottom and following the frequency line to where it intersects the curved line, then reading the loss from the right- or left-hand side of the graph. Several plots appear in Figs. 6-95C and D, illustrating how the insertion loss is plotted for a given value of F_c. After

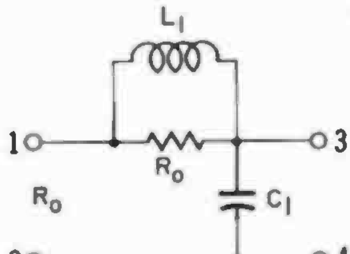


Fig. 6-95A. A low-frequency boost, high-frequency attenuated inverted L-type equalizer.

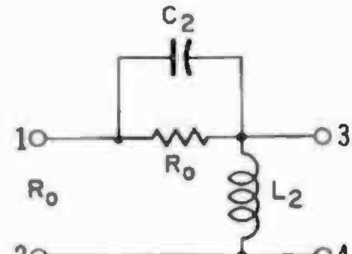


Fig. 6-95B. A high-frequency boost, low-frequency attenuator, inverted L-type equalizer.

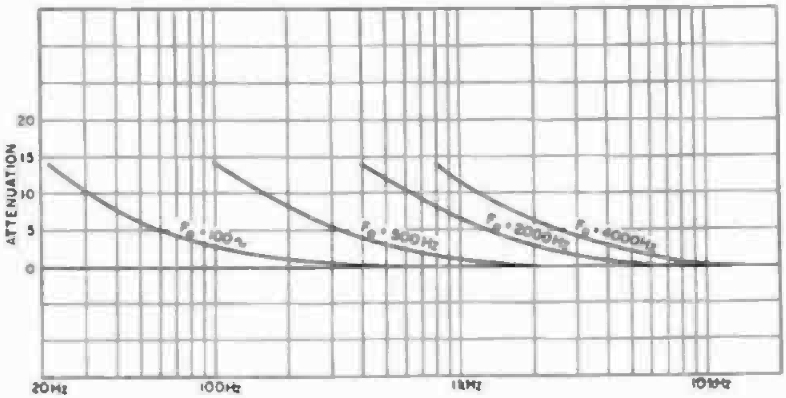


Fig. 6-95C. Typical response curves for different values of F_o for inverted L-type equalizer shown in Fig. 6-95A.

the response has been plotted, the values of inductance and capacity are calculated:

$$L_1 \text{ or } L_2 = \frac{R_o}{2\pi F_o}$$

$$C_1 \text{ or } C_2 = \frac{1}{2\pi F_o R_o}$$

where,

R_o is the circuit impedance,
F_o is the frequency of 3 dB insertion loss.

The insertion loss for any frequency may be calculated:

$$10 \operatorname{Log}_{10}\left[1 + \left(\frac{F}{F_o}\right)^2\right]$$

or

$$10 \operatorname{Log}_{10}\left[1 + \left(\frac{F_o}{F}\right)^2\right]$$

where,

F_o is the frequency of 3 dB insertion loss,
F is the frequency.

The first equation is used with curve A of Fig. 6-80C and the second equation is used with curve B. Curve A is used when designing an equalizer for increasing the low-frequency response or attenuating the high frequencies. Curve B is used when designing an equalizer for increasing the high-frequency re-

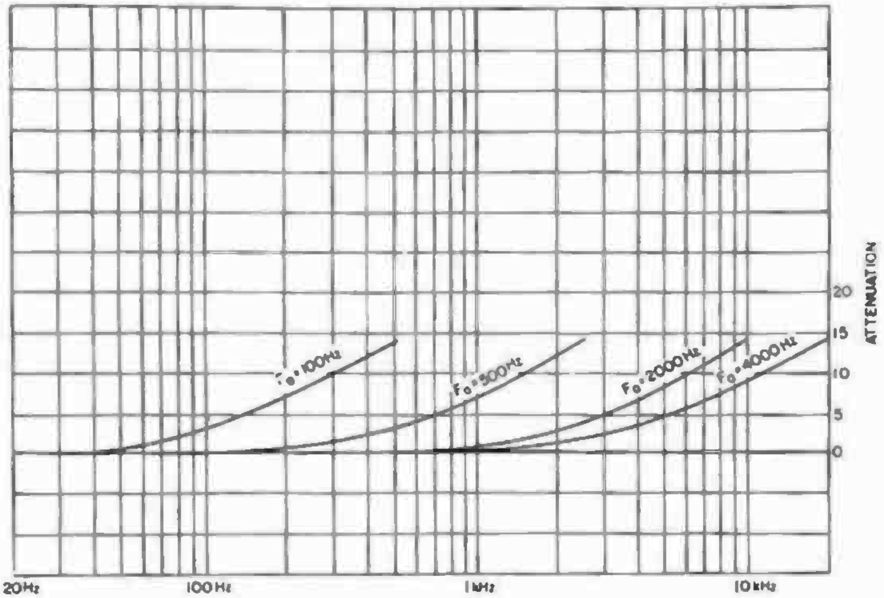


Fig. 6-95D. Typical response curves for different values of F_o for inverted L-type equalizers shown in Fig. 6-95B.

sponse and attenuating the low frequencies.

6.96 ***Show a typical amplifier circuit with the points in the circuit where the frequency characteristics may be changed.***—A schematic of a two-stage amplifier with various points where equalization may be inserted to alter the frequency characteristics is given in Fig. 6-96. Neither the circuit nor the values of the components in Fig. 6-96 are intended to represent any particular amplifier or produce a given response. It presents only a composite picture of the methods used to secure frequency compensation.

Beginning at point 1, Fig. 6-96, T1 is the conventional input transformer with a terminating resistor R1 across the primary to present a solid termination to the source impedance.

At point 2 is a capacitor C1 and re-

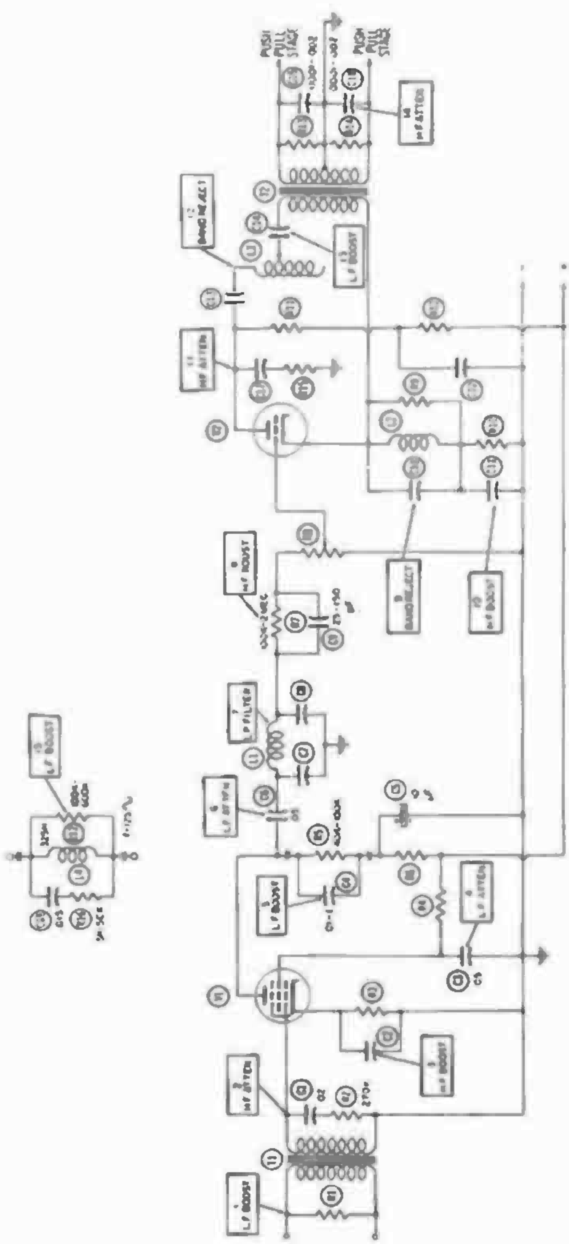


Fig. 6-96. A two-stage amplifier showing different points where equalization may be inserted. The values are for illustration only.

sistor R2 which may be used to secure a slow rolloff at the higher frequencies. The capacitor value selected determines the frequency at which the rolloff begins, while the series resistor controls the amount of attenuation.

At point 3, tube V1, varying the size of the cathode-bypass capacitor C2 will control the high-frequency boost. The smaller the value of this capacitor the greater will be the high-frequency response, because of degeneration in the cathode circuit. For high-frequency boost, the capacitor will vary from 0.10 to 0.02 μF. If the cathode circuit is not used for frequency correction, it is bypassed with a 40 μF or greater capacitor.

Low-frequency attenuation may be obtained by reducing the size of the screen-grid bypass capacitor C3, at point 4. Normally, the capacitor would be 10 μF. By reducing the size to 1 μF or less, the low-frequency response may be reduced accordingly.

Low-frequency boost may be obtained at point 5 by the use of capacitor C4 in the plate circuit of V1 which bypasses a portion of the plate-load resistor R5. Capacitor C4 is selected for a value that will attenuate the high frequencies, beginning at a predetermined frequency; thus, a low-frequency boost is obtained. Capacitor C5 and resistor R6 comprise the usual decoupling circuits common to high-gain amplifiers.

Decreasing the size of the grid-coupling capacitor C6 at point 6, results in a decrease of the low-frequency response. However, it is better to secure low-frequency attenuation in some other part of the circuit and not reduce the coupling capacitor below 0.05 μF.

Should a low-pass filter be required, it may be inserted in the plate circuit of V1 at point 7. The resistor R7 and capacitor C9 shown at point 8 are used for high-frequency equalization and form the familiar grid-circuit equalizer so frequently found in many general-purpose amplifiers. The value of C9 is, as a rule, on the order of 25 to 150 pF. The value of R7 will vary from a few hundred ohms to several megohms.

A dip filter consisting of inductance L2, capacitor C10, and resistor R9 is included in the cathode circuit of V2 at point 9. This circuit is used to remove specific bands of frequencies between 150 and 350 Hz. It is this range of frequencies which causes tubby reproduction in large auditoriums. The inductance L2 is resonated by C10. Resistor R9 is used to broaden the resonance curve. Capacitor C11 across the cathode of V2 at point 10 may also be used for high-frequency boost and is similar in action to that of C2 at point 3.

The series resistor R15 and capacitor C17 at point 11 may be connected from the plate of V2 to ground for the reduction of the higher frequencies. Its action is similar to that of the familiar tone control. Band rejection may also be secured by means of capacitor C13 and coil L3 connected in series with the primary winding of the push-pull interstage transformer T2.

At point 13 the primary of the push-pull transformer is parallel-coupled through capacitor C14 to the plate of V2. If a value of C14 is selected to resonate the primary of T2 at some frequency, a low-frequency boost is obtained. However, it is not good practice to nse this method, as distortion may be induced.

High-frequency attenuation may also be secured by shunting capacitors C15 and C16 across the secondary of the interstage transformer T2, indicated at point 14. Resistive terminations R13 and R14 across the secondary may also be used. Resistor R12 and capacitor C12 are connected in the plate circuit of V2 for decoupling purposes.

In the upper portion of the diagram at point 15 is shown a low-frequency booster circuit which may be used in place of C4 and R5 (point 5) in the plate circuit of V1. The circuit consists of a large inductance, L4, of approximately 325 henries resonated by capacitor C18 in series with R16. The parallel resistor, R17, controls the amplitude, and resistor R16 controls the width of the resonance curve.

Numerous other methods will suggest themselves for securing frequency correction; however, it should be remembered that circuits such as those illustrated, while effective, must be very carefully applied to prevent overload and distortion, since these latter conditions rise rapidly when circuits as described are employed. It is good engineering practice to divide the equalization between stages whenever possible. Many of the circuits shown in Fig. 6-96 are used in projection amplifiers in motion picture theaters to obtain characteristics to suit individual installations.

6.97 ***Show a three-channel equalizer amplifier.***—Fig. 6-97A depicts a three-channel equalizer amplifier which will permit control of high, low, and intermediate frequencies independently of each other. This device consists of three separate amplifier stages, each designed to cover a limited frequency range. The three interstage amplifiers are fed from a single input stage, while the outputs are combined and fed into one output amplifier stage. At the input of each amplifier is an RC network for limiting the amplifier response. The network consists of elements C1, C2, R1, and P1. The gain of each stage may be controlled by means of the individual gain controls P1, P2, and P3. The frequency response of a single stage is given in Fig. 6-97B.

6.98 ***How can the frequency response of an amplifier using negative feedback be altered?***—Frequency correction may be connected in any stage outside of the feedback loop. Frequency correction may also be obtained by changing the values of the elements in

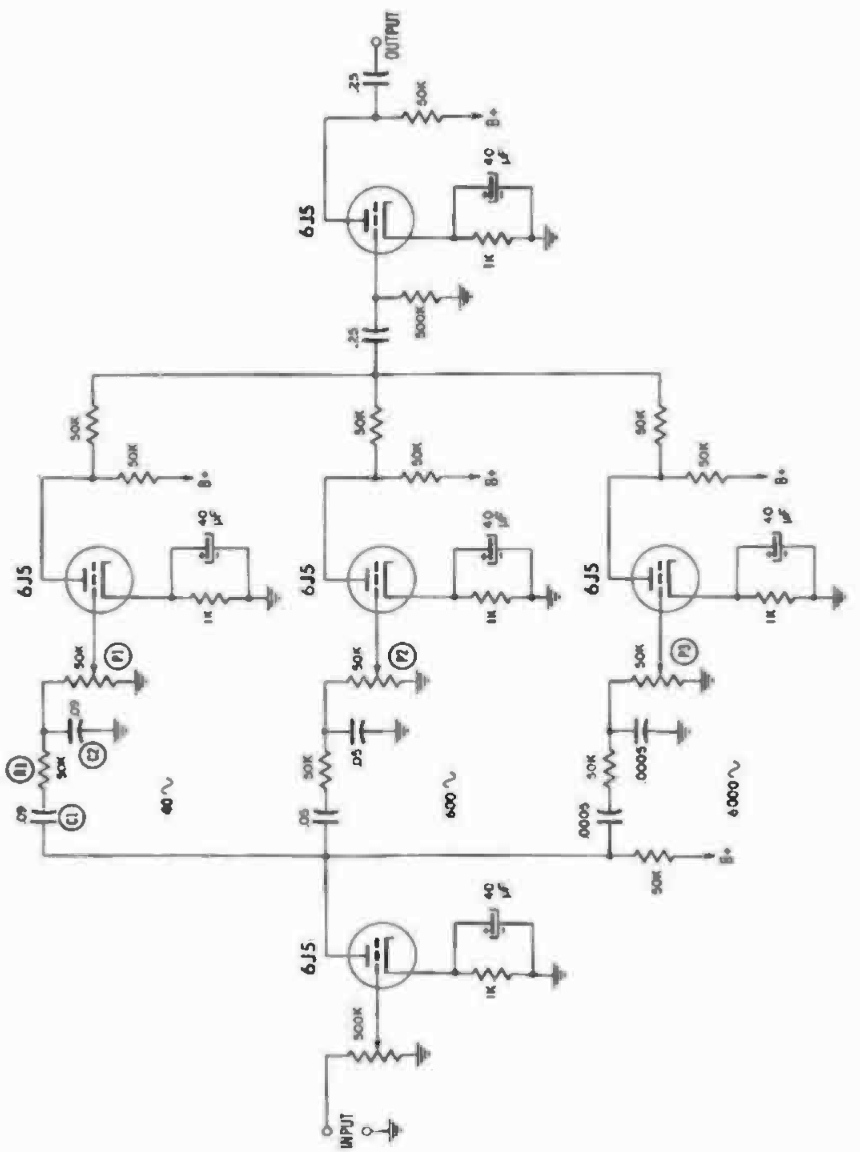


Fig. 6-97A. A three-channel equalizer amplifier.

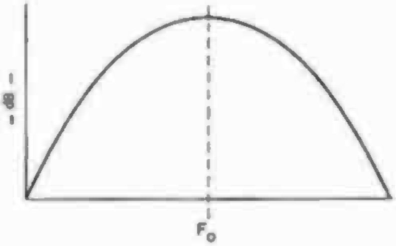


Fig. 6-97B. Frequency response of a single stage of the amplifier shown in Fig. 6-97A.

the feedback loop. Negative-feedback amplifiers are discussed in Questions 12.136 through 12.168.

6.99 *What is a tone control?*—A simple, high-frequency attenuating network used in radio sets to reduce the high-frequency response. A typical circuit is shown in Fig. 6-99.

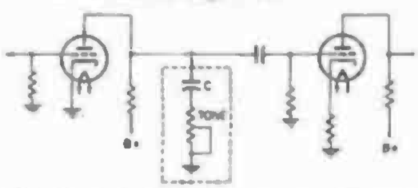


Fig. 6-99. A typical tone-control circuit used in radio receivers.

6.100 *Describe a microphone equalizer.*—Microphone equalizers are used ahead of the mixer control and are installed in the mixer panel as a part of the mixer network. They permit the recording engineer to select a given microphone which provides an optimum response and polar pattern, and match the response of the other microphones

Fig. 6-100A. Altec-Lansing microphone equalizer Model 9060A.

to the master microphone. Such equalizers are designed for correcting the variations in response, changes in the *apparent* response caused by variation in the microphone to source distance, and the acoustical characteristics of the recording stage. They may also be used to a good advantage for matching indoor and outdoor scenes or live and dead rooms, as well as for dialogue equalization as discussed in Question 6.80.

An equalizer of this design is shown in Fig. 6-100A. The device consists of a *passive* network employing inductance, capacitance, and resistance in a bridged-T configuration. Two slide-type switches provide a maximum equalization of plus 12 dB at 100 Hz and 7000 Hz, and a minimum attenuation of minus 16 dB at 100 Hz and 10,000 Hz. The insertion loss is 14 dB, and has a

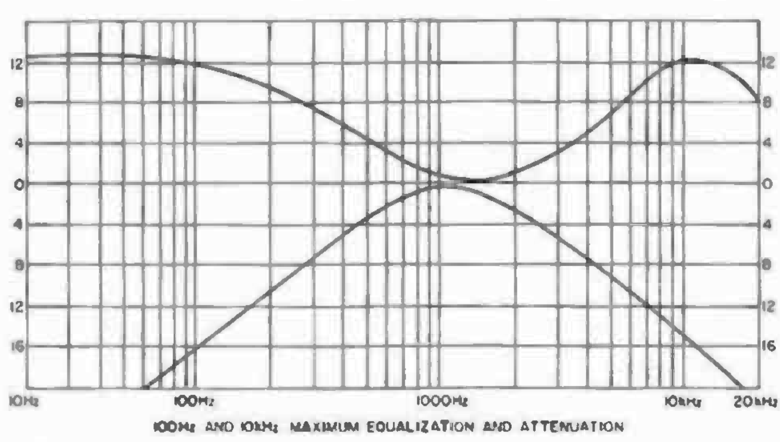


Fig. 6-100B. Frequency characteristics of Altec-Lansing microphone equalizer Model 9060A.

constant input and output impedance of 600 ohms. The maximum and minimum limits of equalization and attenuation are shown in Fig. 6-100B. The basic design is the *constant-B* equalizer discussed in Question 6.124. The internal wiring consists of printed-circuit boards.

6.101 ***Give the component values for an RC equalizer suitable for reproducing the RIAA disc record characteristic.***—A circuit, with its component values, designed to be connected between the plate and grid circuits of two vacuum tubes is shown in Fig. 6-101.

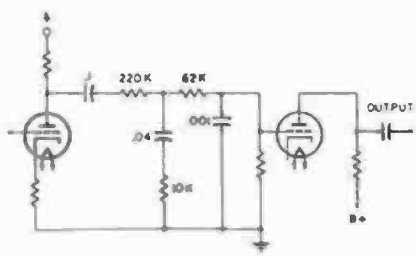


Fig. 6-101. RC equalizer network suitable for reproducing using the RIAA standard reproducing characteristic.

6.102 ***What are the component values for a 600-ohm high-frequency bridged-T equalizer suitable for reproducing the RIAA characteristic?***—Such a circuit appears in Fig. 6-102.

6.103 ***What are the component values for a 600-ohm low-frequency bridged-T equalizer suitable for reproducing the RIAA characteristic?***—Such a circuit is shown in Fig. 6-103.

6.104 ***What are the component values for a 600-ohm high-frequency bridged-T recording equalizer suitable for recording the RIAA characteristic?***—The circuit and component values are given in Fig. 6-104.

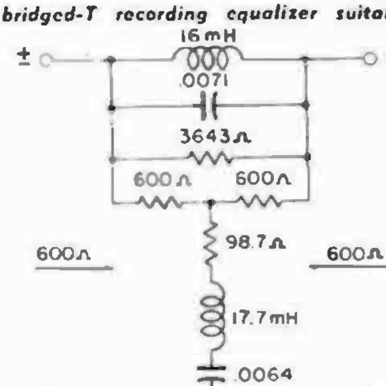


Fig. 6-102. A high-frequency, bridged-T, post-equalizer for reproducing recordings using the RIAA characteristic.

Equalizers of this design and characteristic were used in disc recording circuits before the adoption of the RIAA Standard Reproducing Characteristic. (See Question 13.95.)

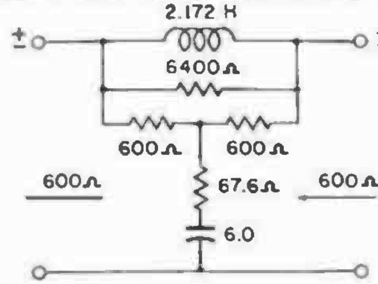


Fig. 6-103. A low-frequency, bridged-T, post-equalizer for reproducing the RIAA characteristic.

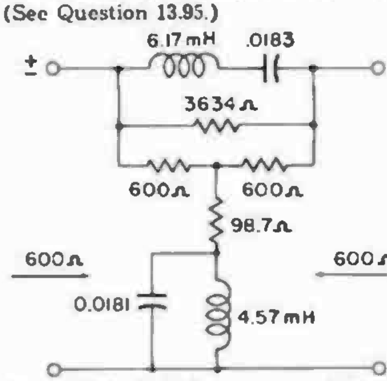


Fig. 6-104. A high-frequency, bridge-T, pre-equalizer for recording using the RIAA recording characteristic.

6.105 ***What are the RIAA recording and reproducing characteristics?***—The recording and playback characteristics as standardized by the RIAA are plotted in Fig. 13-95. These standards are employed by the majority of the recording companies in the United States and also by some foreign recording companies. It will be noted the reproducing characteristic is an inverse curve to the recording characteristic. The curves shown are used for recording 16-inch transcriptions, 78 rpm, and microgroove records. A different recording and reproducing characteristic is used for vertical recording and reproduction. This characteristic is shown in Fig. 13-94.

6.106 ***What are the component values for a bridged-T variable equalizer***

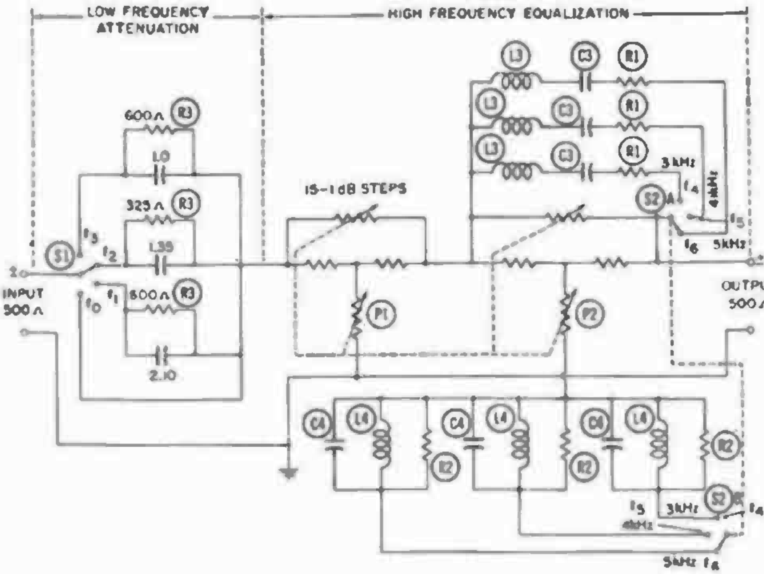


Fig. 6-106A. A constant-loss variable equalizer suitable for motion picture rerecording.

suitable for rerecording purposes?—Fig. 6-106A depicts a constant-loss bridged-T equalizer with three high-frequency characteristics (Fig. 6-106B). In addition to the above, three low-frequency attenuation positions are provided. Resistors R1 and R2 are connected in series and in shunt with the inductances to permit broadening the resonance curves, if desired. Resistors R3 are connected in shunt with the capacitors in the low-frequency attenuation circuits to provide a shelving characteristic at the extreme low-frequency end. Circuit element values may be calculated from the information contained in Question 6.53.

6.107 *Show a schematic diagram for a variable dialogue equalizer suitable for motion picture rerecording.*—Fig. 6-107A is the schematic diagram of a six-position, variable dialogue equalizer with suitable characteristics for use during rerecording operations. Basically, the device consists of three high-frequency equalizers, Sections 1, 3, and 5, resonated at 1.5, 3, and 6 kHz; and three dip equalizers resonated at the same frequencies. Each of the six equalizer circuits may be varied individually in steps of 1 dB, for a total of 12 dB. The six networks are connected in tandem and may be varied during recording, if necessary.

Sections 1, 3, and 5 are of constant-loss design as described in Question 6.51. Sections 2, 4, and 6, the dip equalizers, require only one attenuator because the network is designed to induce a loss at the resonant frequency. If these sections were to be made constant-loss, no effect would be noted when the attenuator was rotated.

Two amplifiers are used, one ahead of the networks and the other following. Their purpose is to compensate for the insertion loss of the attenuator networks and permit the device to be operated in a recording channel as a no-loss, no-gain device. The required

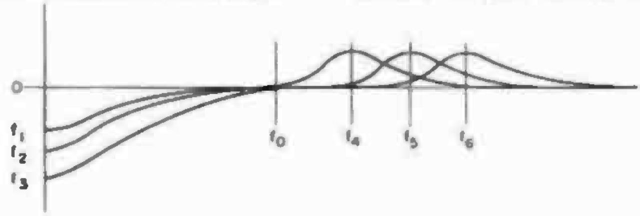


Fig. 6-106B. Frequency characteristics of the variable equalizer shown in Fig. 6-106A.

gain is divided between the two amplifiers. The gain is so set that, when a given signal level is applied to the input (zero equalization), the same signal level is obtained at the output. Operating in this manner permits the complete equalizer to be keyed into and out of the circuit without disturbing the gain of the recording channel.

The coils for the equalizer and dip sections are toroidal wound and encapsulated in plastic. (See Fig. 8-94.) Because these coils are toroidal wound, they may be mounted one above another without fear of intercoupling. The Q of the coils ranges from 60 to 125 at the resonant frequency, which is necessary if they are to meet the frequency response shown in Fig. 6-107B. Although the attenuators are shown having a 12-dB loss, the circuit elements have been computed using a maximum

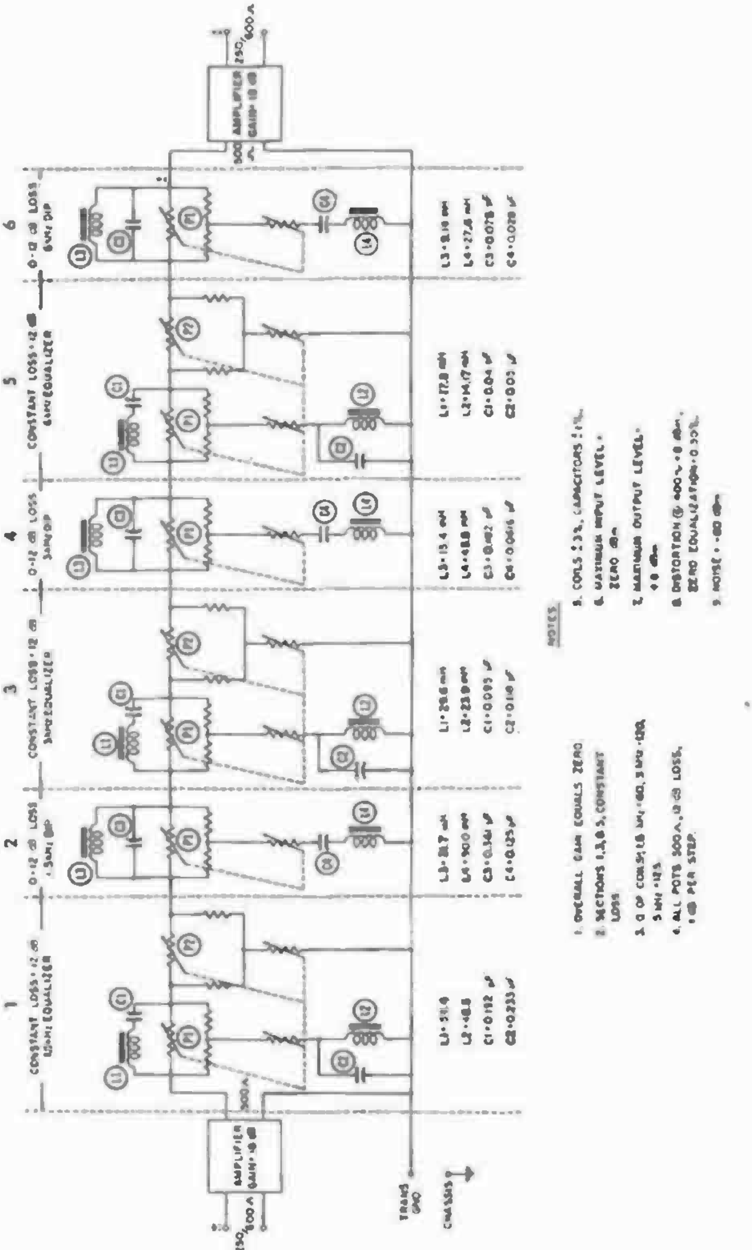


Fig. 6-107A. Schematic diagram for three-frequency variable dialogue and sound effects equalizer.

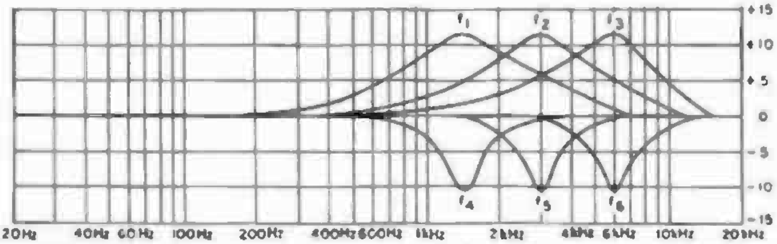


Fig. 6-107B. Frequency response of equalizer shown in Fig. 6-107A.

loss of 7 dB. This was done to obtain a broader response curve for the first 7 steps of the attenuator. After the 7th step, the curves become somewhat steeper.

6.108 Describe a graphic equalizer? —It is a variable equalizer used for the rerecording of motion pictures and is similar to that shown in Fig. 6-108A. The equalizer consists of two units: an amplifier and a group of resonant circuits; and a control panel containing a group of attenuator controls, with straightline characteristics, operating in conjunction with a panel designed to show the equalization graphically by the position of the equalizer controls. Each control permits a portion of the audio frequency spectrum to be increased or decreased 8 dB in steps of 1 dB. Frequencies of 63, 160, 400, 1000, 2500, and 6300 Hz have been selected to provide a balanced energy response in the intermediate frequencies. Because of the control panel design the mixer sees the amount of equalization graphically at all times; hence its name. As a rule, the device is operated as a no-gain, no-loss device as explained in Question 6.107.

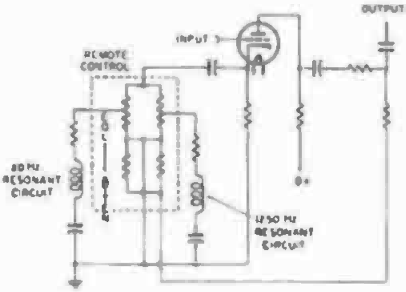


Fig. 6-108B. Simplified diagram of graphic equalizer showing the circuitry of the first amplifier stage.

A simplified diagram of one of the three first amplifier stages is shown in Fig. 6-108B. The same principle of operation and design is employed in the second and third stages.

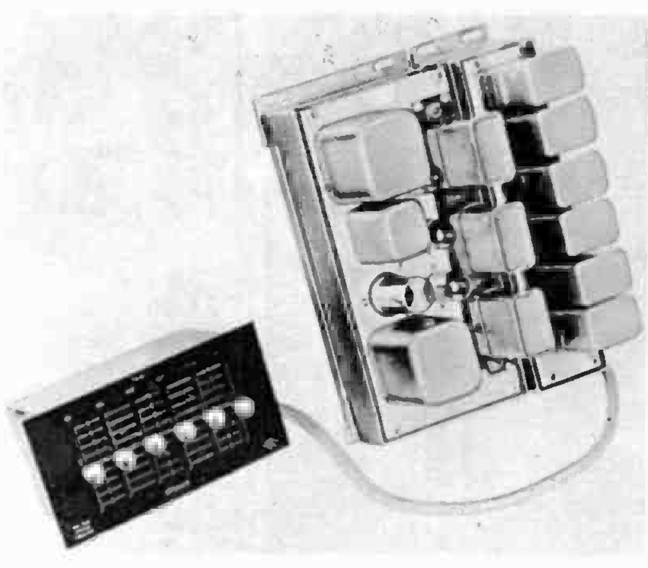


Fig. 6-108A. Cinema Engineering Co., type 7080, graphic equalizer amplifier.

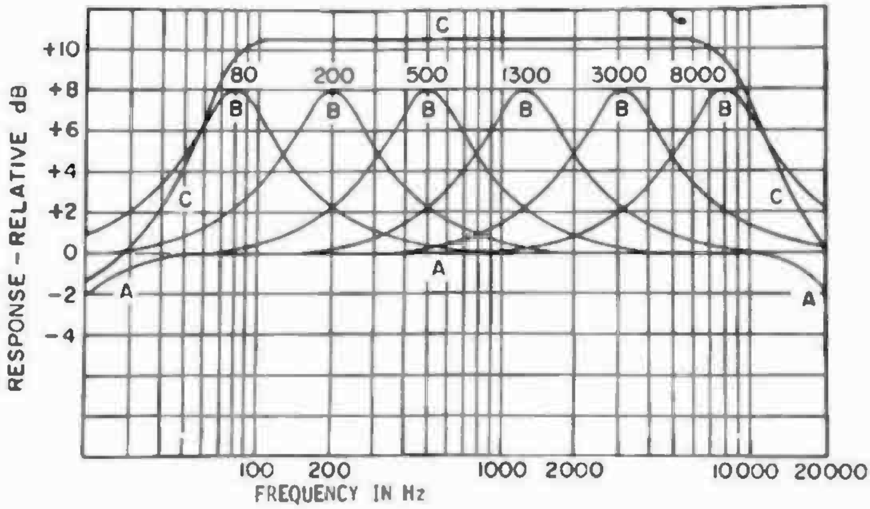


Fig. 6-108C. Frequency characteristics of a typical graphic equalizer using 80, 200, 500, 1300, 3000, and 8000 Hz.

The control potentiometers between the top and the midpoint (which connects to ground) connect the tuned circuits across the cathode resistor, providing a rise of 8 dB in 1 dB steps, at two different frequencies.

The lower portion of the control pots connect the tuned circuits in series with a network in the plate circuit, thus providing attenuation of the frequency bands covered by the tuned circuits.

The "Q" of the tuned circuits is such that when the controls are set in line mechanically the circuit is electrically flat with respect to frequency.

In Fig. 6-108C are shown the frequency characteristics of the device with all the controls set to zero (curve A), the *individual* frequency response of each control at plus 8 dB *equalize* (curve B), and the frequency response with *all* controls set to plus 8 dB equalize (curve C).

In the attenuate positions the frequency characteristics of the individual controls are inverse to that of when the controls are set to the equalize (rise) positions. For certain types of recording it might be desirable to use frequencies of 82, 205, 500, 1250, 3200, and 8000 Hz.

The schematic diagram of a typical graphic equalizer is shown in Fig. 6-108D with only *two* of the controls and their tuned circuits shown to simplify the diagram. The total resistance of the control is approximately 10,000 ohms, and is calibrated in steps of 1 dB, 8 dB, plus, and 8 dB minus from center.

Graphic equalizers may also be designed using *passive* networks requiring no amplifiers. The gain of the system is adjusted to compensate for the *fixed* insertion loss of the equalizer network (about 14 to 16 dB). Such devices are discussed in Question 6.126.

6.109 Show the configuration for a high-frequency post-equalizer for reproducing vertical cut records.—A high frequency post-equalizer suitable for reproducing vertical cut records is shown in Fig. 6-109. The low-frequency post-equalizer shown in Fig. 6-103 is connected in tandem with the high frequency equalizer to compensate for the constant-amplitude characteristics of the recording system.

6.110 Give a schematic diagram for an equalizer suitable for transferring 35-mm photographic sound track to 16-mm photographic sound track.—A bridged-T equalizer and low-pass filter, suitable for transferring 35-mm optical sound tracks to 16-mm optical film is shown in Fig. 6-110A.

The equalizer consists of three sections: a low-frequency attenuator, a low-pass filter to provide a sharp cutoff at the higher frequencies, and a high-frequency adjustable equalizer to compensate for film losses. The bridged-T equalizer employs a 10-dB pot adjustable in steps of 1 dB. In addition to the equalizer shown, an 80-Hz high-pass filter should be patched into the channel to limit the low-frequency response. The frequency characteristic of the equalizer and low-pass filter are shown in Fig. 6-110B.

This equalizer combination will be found useful also when transferring 35-mm magnetic sound track to 16-mm photographic sound-track negative; in fact, it is quite useful in many operations involving 16-mm photographic sound track.

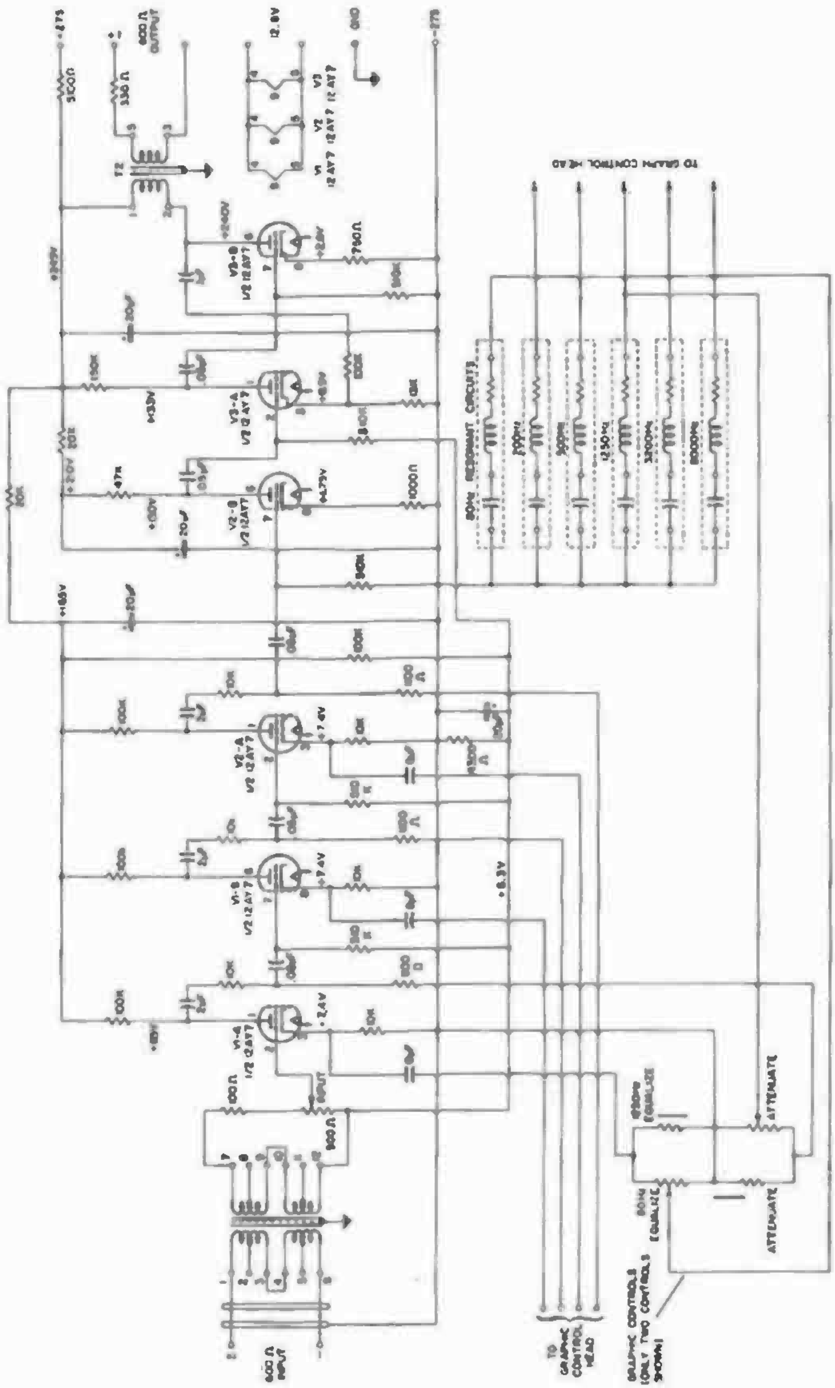


Fig. 6-108D. Schematic diagram for graphic equalizer.

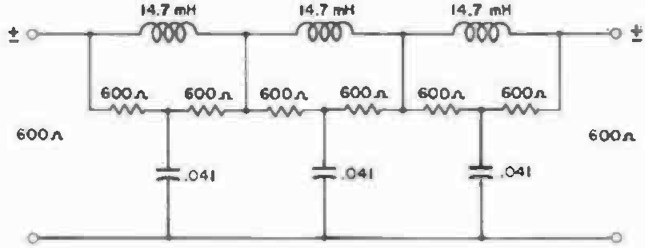


Fig. 6-109. A high-frequency post-equalizer for reproducing 16-inch 33⅓ rpm transcriptions.

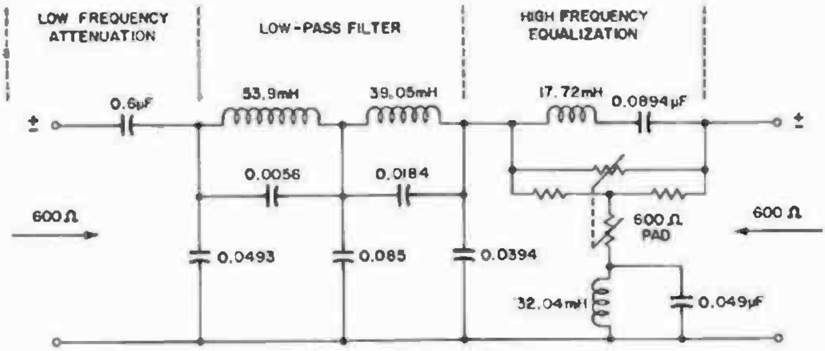


Fig. 6-110A. Variable equalizer suitable for transferring 35-mm photographic sound track to 16-mm photographic sound track.

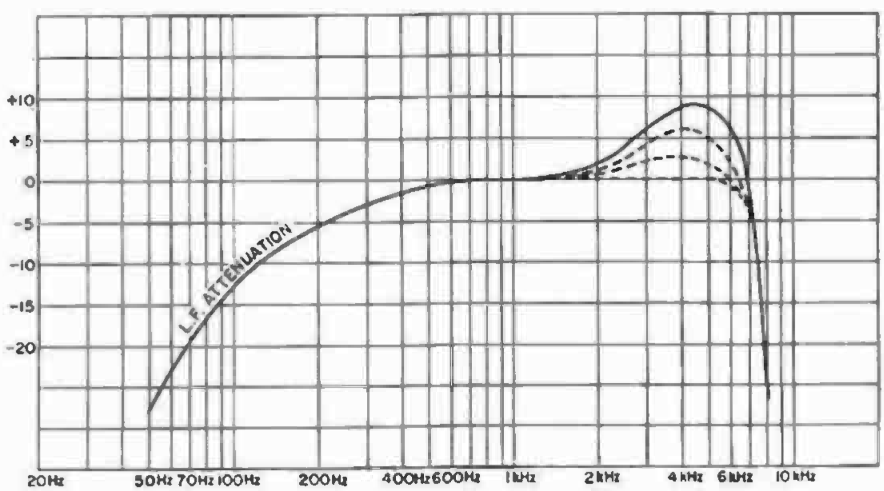


Fig. 6-110B. Frequency characteristics of 16-mm film recording equalizer shown in Fig. 6-110A.

6.111 What are the pre-emphasis and post-emphasis characteristics used in fm radio transmitters and receivers? —Because most of the energy of human speech is in the low and middle frequencies, it is necessary to increase the signal-to-noise ratio of radio-transmitted speech by pre-emphasizing the higher frequencies as illustrated in Fig. 6-111A. The FCC specifies that pre-emphasis shall be employed in accordance with the impedance-frequency characteristic of a series inductance-resistance circuit having a time constant of 75 microseconds.

To return the frequeucy response to one of uniform characteristics in the radio receiver, a post-emphasis or de-emphasis circuit must be connected at the output of the discriminator in the

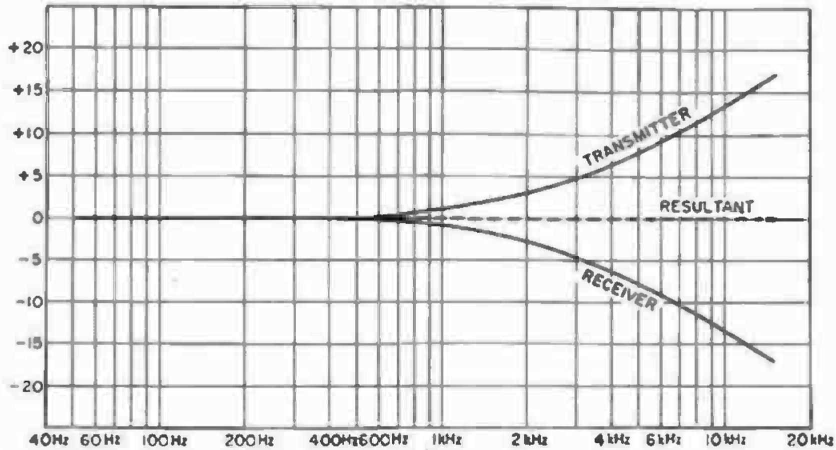


Fig. 6-111A. Pre-emphasis frequency characteristics used in fm and television transmitters and de-emphasis characteristic used in fm and television receivers.

radio receiver. The above circuit may consist of a simple RC network as shown in Fig. 6-111B. The time constant may be calculated:

$$T = R \times C$$

where,

R is the resistance in ohms,
C is the capacitance in microfarads.

Although the circuit values shown in Fig. 6-111B are 100,000 ohms with a capacitance of 0.00075 μF, any values of resistance and capacitance may be used if the time constant is 75 microseconds.

If the high-frequency response is lacking because of the sharpness of the intermediate amplifier in the rf section of the receiver, the high-frequency response may be increased by reducing the time constant to 40 or 50 microseconds. This is often done in the less expensive fm receivers to increase the high-frequency response.

6.112 How is pre-emphasis obtained in an fm radio transmitter?—By connecting an inductance in the plate circuit of a pentode amplifier stage in the speech amplifier, as shown in Fig. 6-112A. At the lower frequencies the reactance of the inductance is negligible; therefore, a plate-load resistor R_L must be connected in the plate circuit. As the frequency rises, the reactance of the coil will also rise, increasing the amplitude of the high frequencies and resulting in a frequency characteristic as shown in Fig. 6-111A.

To obtain a 75-microsecond rise, the ratio of the inductance to the plate resistance of the tube must be in the order of 13.3 ohms of resistance to one millihenry of inductance. A 75-microsecond rise time may also be obtained

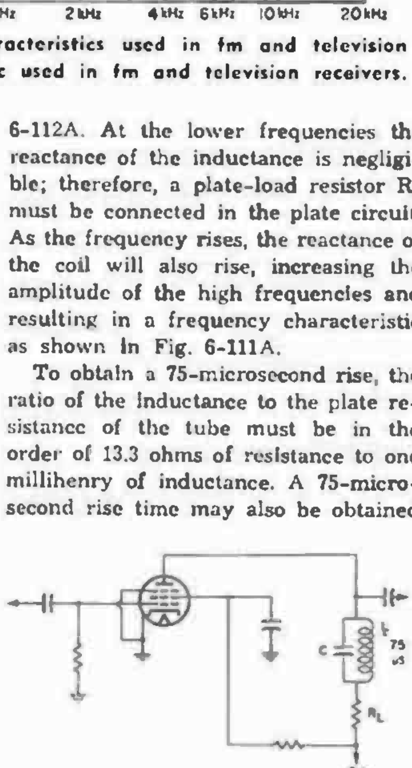


Fig. 6-112A. A 75-microsecond pre-emphasis circuit, connected in the plate circuit of a pentode, in the speech amplifier of an fm radio transmitter.

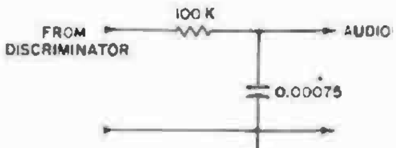


Fig. 6-111B. De-emphasis circuit for fm receiver. This network is connected between the output of the discriminator and the input of the audio stages.

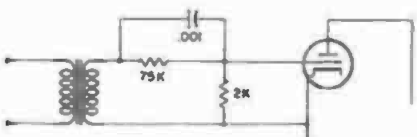


Fig. 6-112B. A 75-microsecond, RC pre-emphasis circuit connected in a low-level stage of the speech amplifier in an fm transmitter.

by the use of an RC circuit as shown in Fig. 6-112B. This circuit is also installed in one of the low-level stages of the speech amplifier. The attenuation characteristics of a de-emphasis circuit expressed in microseconds may be calculated:

$$dB = 10 \log_{10} (1 + w^2T^2)$$

where,

w equals $2\pi f$,
T is the time constant in seconds ($R \times C$).

For a 75-microsecond de-emphasis circuit, this may be written:

$$dB = 10 \log_{10} (1 + 0.222f^2)$$

where,

f is the frequency in hertz,
0.222 is a constant.

6.113 What is split-termination equalization?—Equalization obtained by altering the frequency characteristics of an input transformer by the use of terminating resistors across the primary and secondary as shown in Fig. 6-113.

Assume an input transformer of uniform frequency characteristics with an impedance ratio of 500 to 100,000 ohms is to be used in a microphone preamplifier and that a rising characteristic at the higher frequencies is desired. Further, assume the transformer is normally designed to operate with a 100,000-ohm resistive termination across the secondary. Now, if the secondary is terminated with a 200,000-ohm resistor and the primary terminated with a 1000-ohm resistor, the high-frequency response above 1000 Hz will be increased. By increasing the secondary termination to double that normally used, the reflected primary impedance is also doubled and becomes 1000 ohms, rather than 500 ohms. Connecting a 1000-ohm resistor across the primary returns the impedance, looking into the primary, to 500 ohms again, by

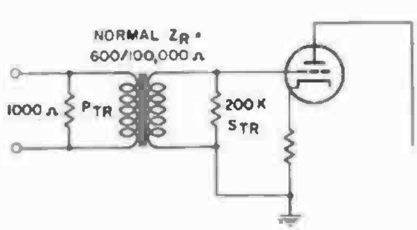


Fig. 6-113. A transformer with split termination to obtain a rising frequency response at the high frequencies.

virtue of the primary impedance being in parallel with the terminating resistor. This combination causes the secondary impedance to look like 66,000 ohms when the primary winding is operating from a 500-ohm source.

The rise in the high-frequency response may be attributed to the mutual capacity existing between the windings of the primary and secondary. The rise at the high frequencies will continue to increase (within limits) as the secondary termination is increased. Double-terminating a transformer increases its insertion loss; however, as a rule, this is not important in an input transformer and may be disregarded.

Split terminations are often used in preamplifiers designed for use with both microphones and photocells. However, they are not recommended for microphone preamplifiers because of the loss of voltage at the secondary. This subject is also discussed in Questions 8.43 to 8.48.

6.114 What is the approximate length of transmission line that may be used without equalization?—The maximum lengths are given:

No. 16 wire	5 miles
No. 19 wire	2.5 miles
No. 22 wire	1.5 miles

6.115 What does the term LC ratio mean?—It is the ratio of inductance to capacitance in a resonant circuit. The frequency of resonance may be calculated for either a series or parallel circuit by the formula:

$$F_r = \frac{159}{\sqrt{LC}}$$

where,

F_r is the frequency of resonance,
L is the inductance in henries,
C is the capacitance in microfarads,
159 is a constant.

As it will be seen, only one value of L times C will resonate to a given frequency. The intended use of the circuit will determine the LC ratio. For a series resonant circuit, the LC ratio is made high. Reducing the LC ratio reduces the steepness of the resonant curve, and the selectivity.

The LC ratio affects a parallel resonant circuit in a manner opposite to the series resonant circuit. As the LC ratio is increased, the selectivity of a parallel resonant circuit is decreased.

6.116 *What is the impedance characteristic of a parallel resonant circuit?*—In a parallel resonant circuit, the impedance reaches a maximum at the resonant frequency. The impedance of a parallel resonant circuit is determined by the formula:

$$Z = Q \times L$$

where,

Z is the impedance,
Q equals X_L/R,
L is the inductance,
R is the dc resistance of the coil.
(See Question 8.72.)

6.117 *What is the impedance characteristic of a series resonant circuit?*—The impedance is at a minimum at the resonant frequency.

$$Z = R$$

where,

Z is the impedance,
R is the ac resistance of the coil.

6.118 *Does the Q of a coil affect the bandwidth?*—Yes. The bandwidth may be expressed as:

$$F_2 - F_1 = \frac{F_r}{Q}$$

where,

F_1 and F_2 are the freqnencies at the edge of the bandwidth, at 0.707 times the current at resonance,
F_r is the frequency of resonance,
Q equals $X_L R$.

6.119 *Where is the bandwidth taken for a series resonant circuit?*—It is designated as a point on the resonance curve that is 0.707 times the current at resonance. Bandwidth is expressed as $F_2 - F_1$. These frequencies are the highest and lowest frequencies at the edges of the bandwidth.

6.120 *What are the essential differences between a series and a parallel resonant circuit?*—The series resonant circuit passes a given band of freqnencies and excludes all others. A parallel resonant circuit excludes a given band of frequencies and passes all others.

6.121 *How may the Q of a circuit be lowered?*—By connecting a resistor in series or parallel with the resonant circuit.

6.122 *What factors influence the equalization characteristics of a dialogue recording channel?*—Several characteristics must be considered. Among them are:

(a) Probable accentuation of the dialogue low frequencies due to the reverberation of the set and recording stage.
(b) The accentuation of the low frequencies because the reproducing level in a theater is generally 6 dB higher than the recording level.
(c) Low-frequency accentuation due to the lowering of the voice approximately 5 dB while recording.

The foregoing factors are plotted as shown in Fig. 6-122 and evaluated as follows: Curve A is the reverberation characteristics of a typical recording stage and set; Curve B is the accentuation of low frequencies, because the recording level is approximately 6 dB lower than the reproduction level in an average theater; Curve C is the accent-

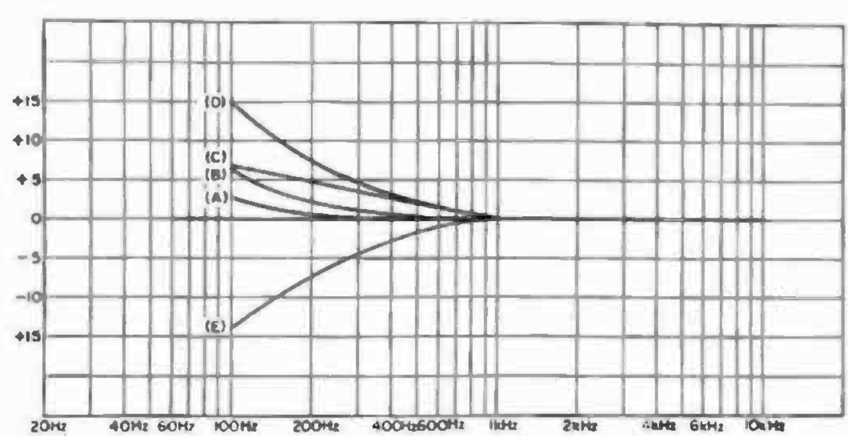


Fig. 6-122. Factors affecting the choice of equalization for dialogue recording.

uation of the low frequencies when the voice is lowered 5 dB during recording. Curve D is the algebraic sum of Curves A, B, and C. Curve E is the inverse of Curve D minus the low-frequency dialogue equalization required for naturalness and intelligibility when reproducing the human voice.

The resulting characteristic is that shown by Curve E, indicating the dialogue should be rolled-off at the lower frequencies, starting at 800 Hz and continuing downward until 100 Hz is attenuated approximately 14 dB with respect to 1000 Hz, when using a microphone of uniform frequency characteristics below 1000 Hz.

In actual practice, 100 Hz is attenuated either 8 or 12 dB with reference to 1000 Hz. This subject is further discussed in Question 18.81 and in Questions 4.114, and 4.27.

6.123 ***What is a delay equalizer?***—A network which adds a time delay in a given frequency band to correct phase-shift distortion. It is used in long line transmissions. The configuration is generally balanced and of the lattice type.

6.124 ***What is a constant-B type equalizer?***—A special type bridged-T equalizer which incorporates two additional resistance arms, R5A and R6A (Fig. 6-124A). It will be noted that one arm is in series with the series-reactive element X_1 and the other is in shunt with the shunt-reactive element X_2.

The conventional bridged-T equalizer requires that the values of the reactive elements be changed to maintain a constant slope for different values of attenuation.

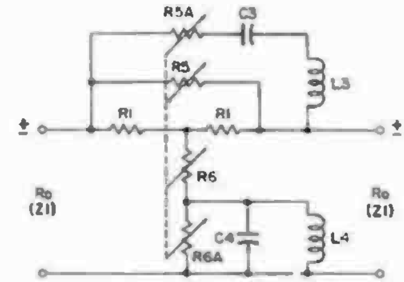


Fig. 6-124A. Constant-B equalizer configuration. The four arms of the attenuator must be varied simultaneously.

The constant-B equalizer controls the slope of equalization without changing the values of the reactive elements. The values of the reactive elements are changed only when changing the resonant frequency. The name constant-B is acquired from the fact that the design frequency for this type equalizer is at one-half the equalizer loss and is designated F_b.

The mechanical construction of a constant-B attenuator is more complicated than the conventional bridged-T attenuator, inasmuch as it requires *four* variable arms that must be operated simultaneously. In the conventional bridged-T equalizer, the shape of the equalization envelope changes with each change of attenuation or equalization, and if plotted for several steps,

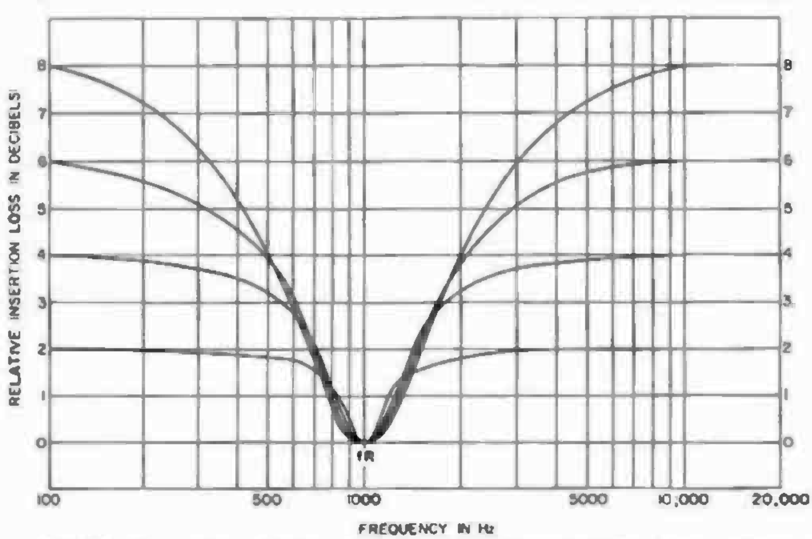


Fig. 6-124B. Conventional bridged-T equalizer characteristics (after Miller and Kimball).

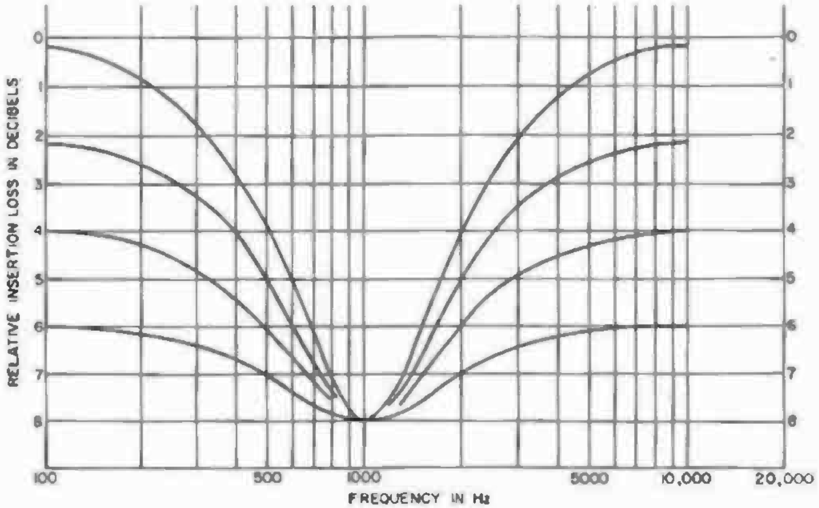


Fig. 6-124C. Constant-B equalizer characteristics (after Miller and Kimball).

will appear as shown in Fig. 6-124B. It will be observed that for an attenuator loss of 2 dB, the half-loss (1 dB) frequency equals 850 Hz, and for 4-dB loss it is 700 Hz and for an 8-dB loss it is 500 Hz. To maintain the half-power frequency at its original point would require that the values of the reactive components be altered each time the attenuation was changed. This can be confirmed by referring to the tables in Questions 6.51 to 6.55. To overcome the characteristics of the bridged-T equalizer, the constant-B equalizer was devised. While it should be understood that the conventional bridged-T equalizer is quite satisfactory for the majority of work (as evidenced by its extensive use in the industry), for equalizers such as described in Questions 6.12, 6.100 and 6.126, constant-B design is a necessity to obtain the proper operating characteristics.

Referring to Fig. 6-124C, the frequency characteristics for a constant-B equalizer, it will be noted as the equalization is increased the half-loss frequency remains the same for any setting between 1-dB and 8-dB attenuation. The frequency of the half-power point is held constant by the additional arms, R5A and R6A. The addition of

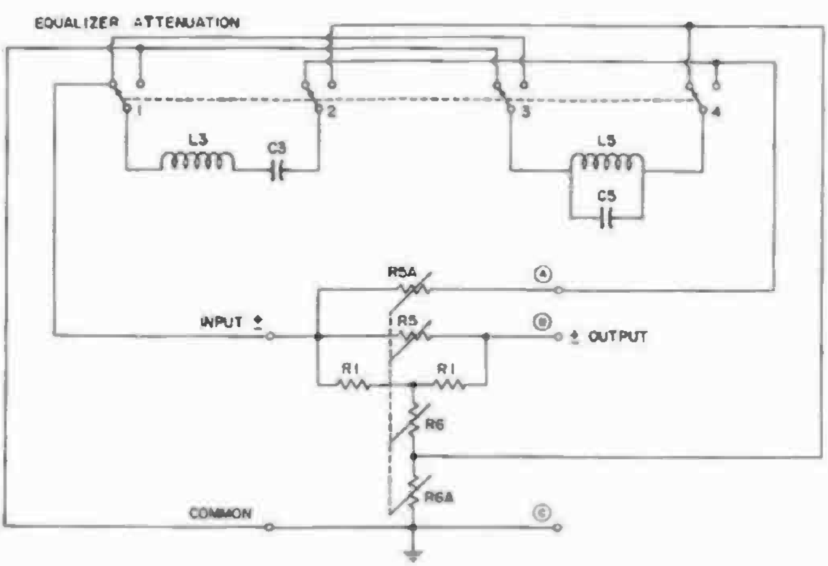


Fig. 6-124D. Constant-B equalizer designed for equalization or attenuation.

these arms modifies the characteristic curve by controlling the amount of maximum and minimum insertion loss while maintaining the difference between the two constants. At a frequency where the reactance X_1 is infinite and X_2 is zero, the configuration becomes the well-known bridged-T pad. When the conditions are reversed and X_1 is zero and X_2 infinite, the circuit becomes an attenuator in which R5 and R5A are in parallel, and R6 and R6A are in series, and is a condition of minimum loss.

To design a constant-B equalizer, the first step is to select the curve of maximum equalization, which means R5A is infinite and R6A is zero, reducing the circuit to a conventional bridged-T equalizer. The circuit is designed as discussed in Question 6.51, and the values of X_1 and X_2 determined. The condition where F_0 will remain constant for all steps of equalization may be approximated:

$$\left[\begin{array}{l}\text{Maximum dB loss}\\ \text{any step}\end{array}\right]^2 = \left[\begin{array}{l}\text{Equalization on}\\ \text{same step}\end{array}\right] \times \left[\begin{array}{l}\text{Equalization on}\\ \text{top step}\end{array}\right]$$

where, equalization is defined as the difference between the *maximum* and *minimum* values of insertion loss, or equalization is equal to the *maximum loss minus the minimum loss*. The top loss is zero but this is not true for the other steps.

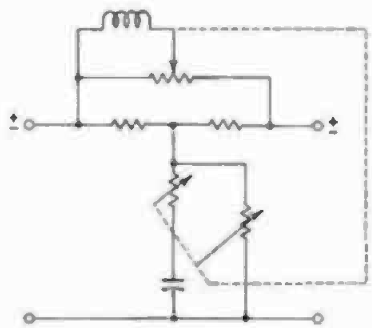
Since the values on the right of the equation are known, only the left side of the equation is to be calculated. A bridged-T attenuator with this loss is now designed to obtain the values of R5 and R6, using the design data given in Question 5.37. To calculate the values for R5A and R6A:

$$\text{Equalization} = \text{Maximum loss} - \text{Minimum loss}$$

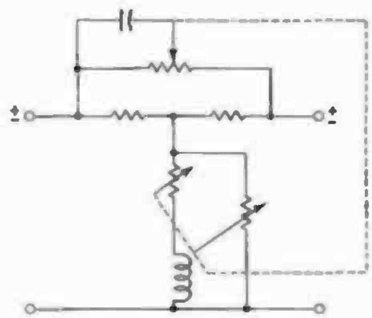
A bridged-T pad is now designed having the loss as determined by the above equation. Resistors R5 and R5A now in parallel form the bridging arm, and as the value of R5 is known, R5A may be calculated. Resistors R6 and R6A form the shunt-arm, and in series must equal the shunt-arm value; R6A may be calculated since R6 is known. In this manner, each step of the network configuration is determined.

Another advantage of the constant-B equalizer network is that it may be used to attenuate or equalize the same band of frequencies by *reversing* the reactive elements (Fig. 6-124D). Symmetrical equalizer and attenuate characteristics will be attained when the maximum loss of the constant-B attenuator is 8.36 dB.

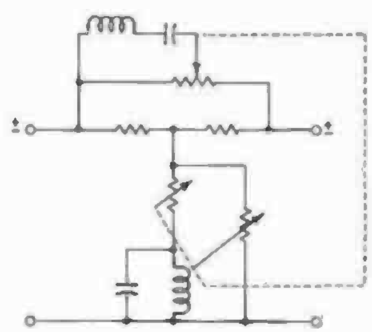
6.125 Describe a constant-S type equalizer.—It has been stated for the bridged-T equalizer that the insertion loss for the unequalized frequencies is changed from step-to-step, and the maximum equalization held constant. In the constant-B equalizer, the slope



(a) *Low frequency.*



(b) *High frequency.*



(c) *Midfrequency.*

Fig. 6-125. Basic configurations for constant-S type equalizer networks (after Solomon and Broneer).

Fig. 6-126A. Graph equalizer Model 9062A manufactured by Altec-Lansing.

of the equalization envelope is controlled without having to change the values of the reactive elements. In the constant-S type equalizer, the insertion loss for the unequalized frequencies is held constant and the point of maximum equalization is changed from step to step. The term "S" is the symbol for the insertion loss.

The constant-S configuration is used in small compact equalizer networks, since the number of contact arms may be reduced to three. Also, the insertion loss is less than for other configurations for a given value of equalization. Three possible configurations are given in Fig. 6-125. The reader is referred to the reference.

6.126 Describe a graphic equalizer of the passive network type.—Graphic equalizers may be designed in two different manners, the *active* type, using a degenerative amplifier as discussed in Question 6.108, or one using a *passive* network requiring no amplification. The statement that a passive network requires no amplification means that it has no internal amplifiers and requires no source of power for its operation. However, because it is a resistive network, it does have an insertion loss; therefore, the system gain following the equalizer must be increased to compensate for the network insertion loss.

The basic design for the equalizer to be described is the constant-B attenuator network, discussed in Question 6.124. Because of its design, the insertion loss is constant for any given frequency, with the equalization variable over its complete operating range, without changing the shape of the equalization characteristic. The constant-B network for this particular equalizer is designed around the half-loss point, for a total loss of 16 dB, resulting in a capability of plus 8-dB equalization and minus 8-dB attenuation, resulting in a total operating range of 16 dB. A front panel view of such an equalizer is shown in Fig. 6-126A.

Two groups of operating frequencies

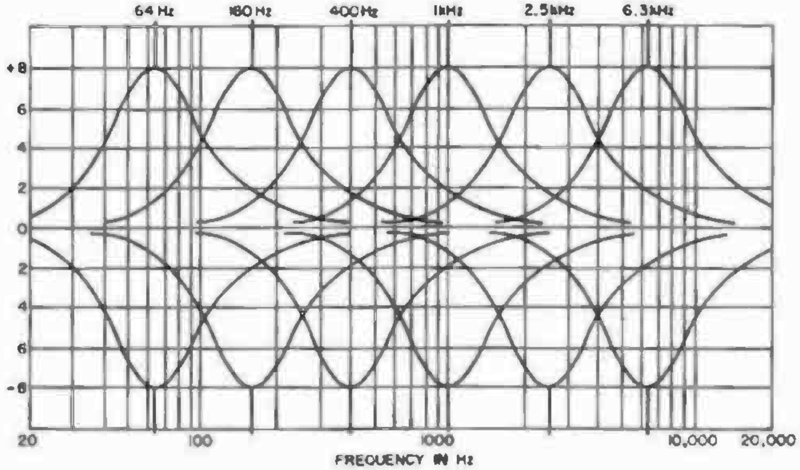


Fig. 6-126B. Frequency response curve for Altec-Lansing Model 9073A motion picture rerecording and dialogue.

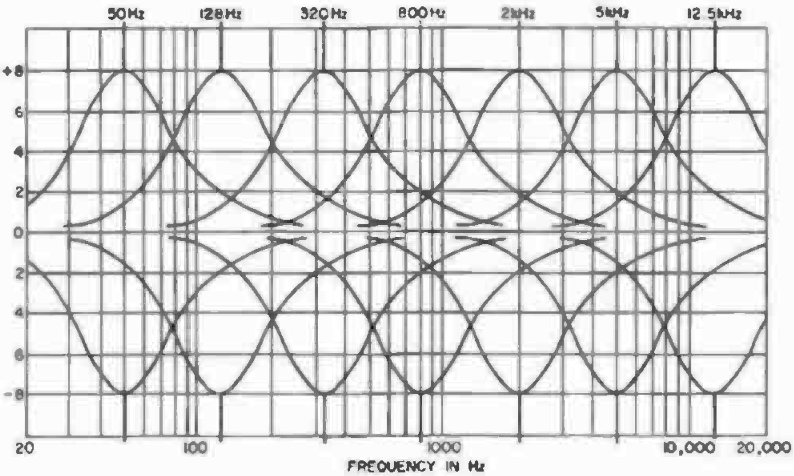


Fig. 6-126C. Frequency-response curve of graphic equalizer for music recording.

are available. The one is for motion picture rerecording, using 64, 160, 400, 1000, 2500, and 6300 Hz. The second group is for wide-range music recording, using frequencies of 50, 128, 320, 800, 2000, 5000, and 12,000 Hz. In the center or zero position the frequency is flat, within plus or minus 0.5 dB. The frequency response for the two type equalizers is given in Figs. 6-126B and 6-126C. The internal circuitry consists of printed-circuit boards. The physical dimensions are 3½-inches high, 10-inches wide and 5¼-inches deep. The usual method of connection is to insert the equalizer between the output of a microphone preamplifier, or reproducing machine, and the input to the mixer console, as in Fig. 6-126D.

6.127 ***Describe a film-loss equalizer and its purpose.***—Film-loss equalizers are used to compensate for the inherent high-frequency losses of photographic film, recorder optics, laboratory processing, and the linear speed of the film. Film loss is measured as set forth in Question 18.170.

After the loss has been plotted, the equalizer is connected in the recording circuit and adjusted for a flat response from the film, within plus or minus 1 dB,

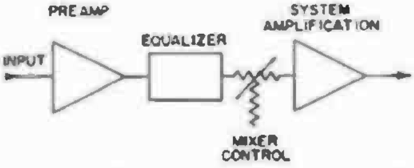
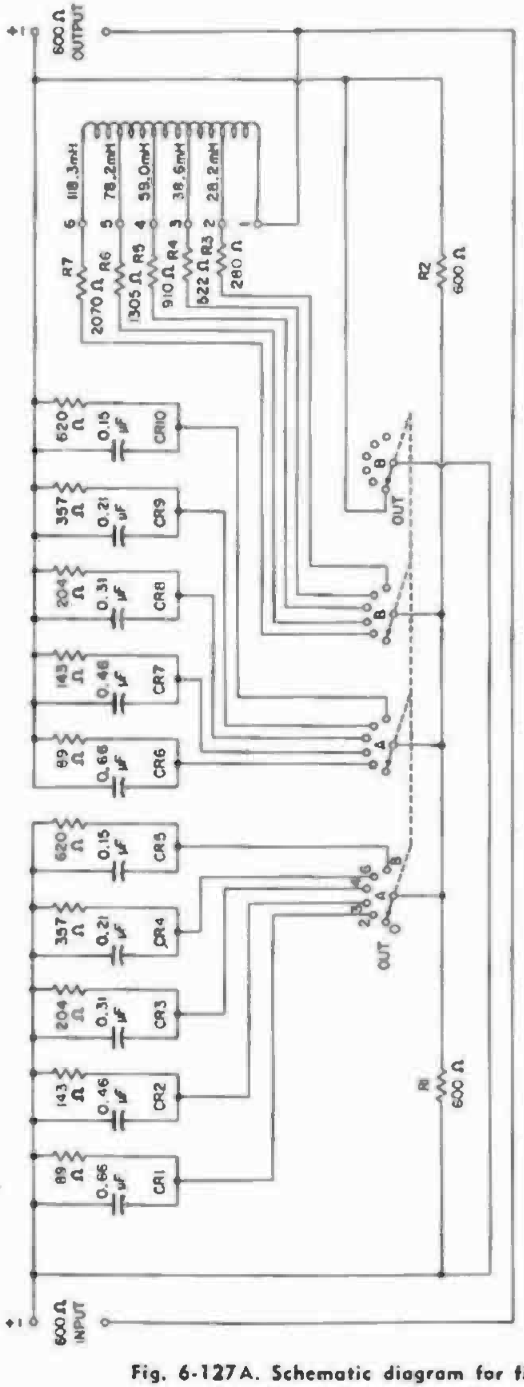


Fig. 6-126D. Connections for a graphic equalizer and mixer network.

by making further measurements. In some instances, the equalization is increased 1 or 2 dB over that required to compensate for high-frequency losses due to variation in recording stock and processing. These are determined from day to day, when the cross-modulation or intermodulation tests are read. Once adjusted the equalizer will require no further attention, unless a radical change is made in the recording stock or laboratory processing.

The schematic diagram for a film-loss equalizer is shown in Fig. 6-127A, with a family of response curves in Fig. 6-127B. The equalizer shown is suitable for either 6000 Hz used for 16-mm film recording, or 9000 Hz used with 35-mm film. The equalization is obtained by means of capacitors and resistors in parallel, with capacitors and building-out resistors in series with the coil taps. The purpose of the resistors in parallel with the capacitors is to shape the curve. If a different slope is required, they may be altered to suit the situation. However, the resistors must be matched in pairs. Since the configuration is unbalanced, the low potential side must be grounded to prevent leakage at the high frequencies. The position of the equalizer in a recording channel is discussed in Questions 18.330 and 18.331.

6.128 ***What is a loudspeaker low-frequency boost equalizer?***—It is a shelf-type equalizer, which is shown in Fig. 6-49B, for increasing the low-frequency response of small loudspeaker systems. It is installed between the output of the driving amplifier and the input to the loudspeaker. Basically the device is designed for 16 ohms, but it may be used with either 8-ohm or 16-ohm systems. However, if used in an 8-ohm circuit the increase in the low-frequency response is less, as indicated by the frequency-response curve of Fig. 6-128. Because the equalizer induces a 4 to 6-dB loss at 100 Hz, the amplifier must be capable of developing sufficient power to compensate for the insertion loss. At least 40 watts of power should be available; if not, severe overloads may be noticed on the heavy low-frequency passages.



RESISTORS ± 1%
CAPACITORS ± 2%
COIL ± 3%

POS	dB EQUALIZATION AT 6000 HERTZ	dB EQUALIZATION AT 9000 HERTZ	dB INSERTION LOSS AT 400 HERTZ
1	OUT	OUT	0
2	1.8	2.0	1.9
3	2.6	2.9	3.1
4	3.4	3.8	4.1
5	4.8	5.5	6.4
6	7.2	8.3	9.2

Fig. 6-127A. Schematic diagram for film-loss equalizer.

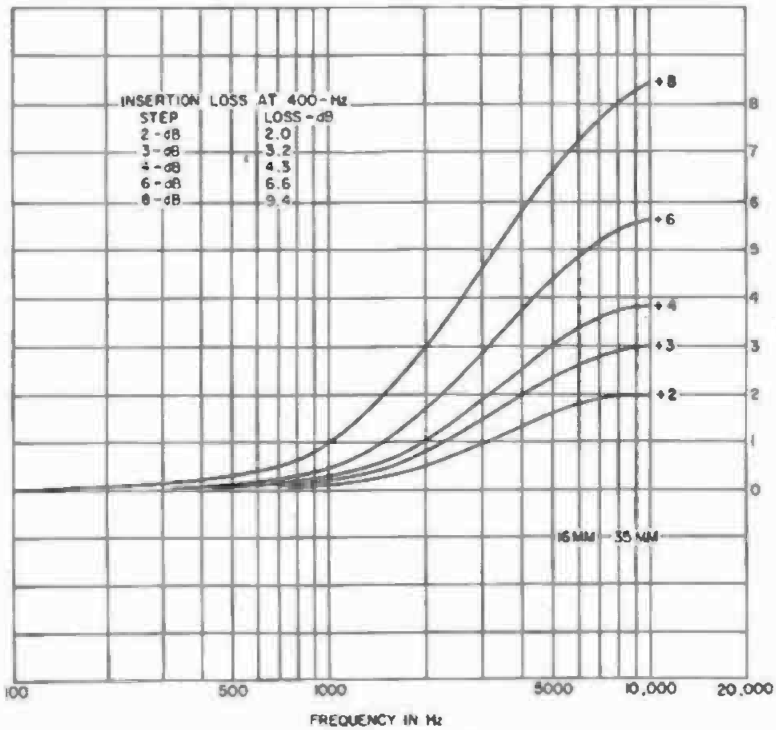


Fig. 6-127B. Film loss equalizer frequency response.

It should be remembered, inducing only a 3-dB loss at 100 Hz lowers the power output from the amplifier to the loudspeaker by 50 percent. Therefore, for a given loudness from the loudspeaker, the amplifier must be driven harder, which in some instances may drive the amplifier into overload. The circuit elements may be designed as discussed in Question 6.49.

6.129 ***Describe a stereophonic pickup equalizer suitable for low-impedance lines.***—A dual equalizer developed by Shure Bros. Inc., for use with stereophonic pickup cartridges is shown in Fig. 6-129A. Although the circuitry (Fig. 6-129B) was developed for use with the manufacturer's pickup, it may be used equally well with pickups of other manufacture. Three different frequency-response characteristics are available: flat response for the repro-

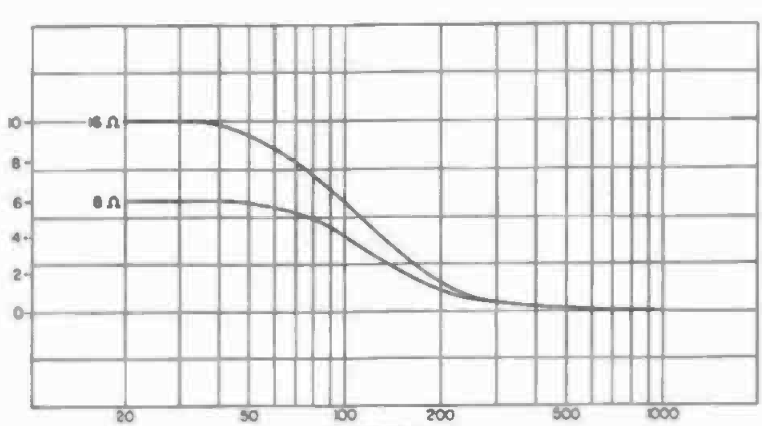


Fig. 6-128. Frequency response of a loudspeaker low-frequency equalizer.

duction of recordings without high-frequency pre-emphasis, standard RIAA response, and a high frequency rolloff for reducing surface noise (Fig.

Fig. 6-129A. Shure Bros. Inc. Model M66 broadcast stereophonic equalizer.

6-129C). For use with high-impedance cartridges having an inductance of 365 to 500 millihenries, the response is plus or minus 1 dB from 30 to 20,000 Hz. For low-impedance cartridges having an inductance of 175 millihenries, the response is within plus or minus 2 dB from 50 to 20,000 Hz. For monophonic reproduction, the inputs of the two sides are connected in parallel. The channel separation is better than 30 dB over the complete spectrum, and may be used with line impedances of 150 to 600 ohms.

6.130 *What is a universal equalizer?*—It is an equalizer having multiple frequency characteristics. The universal equalizer to be discussed has incorporated in its design many features that make it ideal for both large and small studios, where many different types of equalization or filtering are a

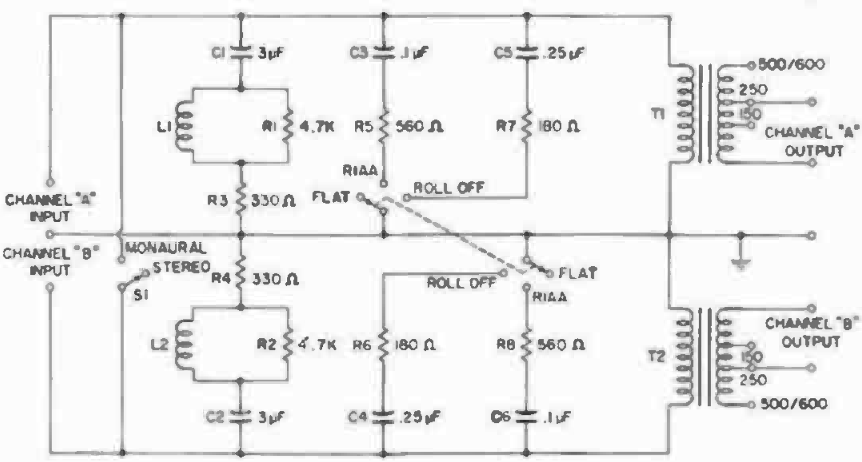


Fig. 6-129B. Schematic diagram for Shure Bros. Inc. Model M66 broadcast stereophonic equalizer.

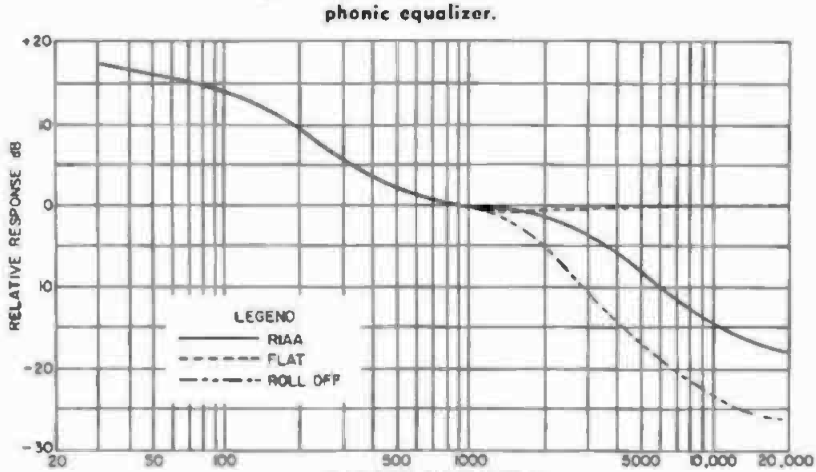


Fig. 6-129C. Frequency response for Shure Bros. Inc. Model M66 broadcast stereophonic equalizer.

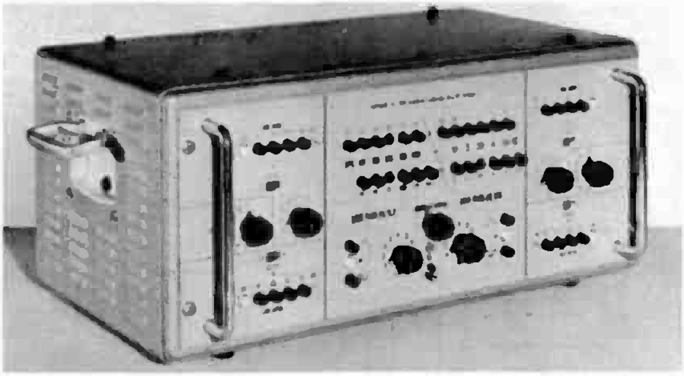


Fig. 6-130. Gotham Audio Corp. Model EQ 1000 universal equalizer.

part of the daily routine. A front view of a universal equalizer, Model EQ 1000, Gotham Audio Corp., and manufactured by Klein-Hummel of West Germany, is pictured in Fig. 6-130.

At the left are several push-buttons for selecting low-frequency equalization or attenuation at a rate of 6, 12, or 24 dB per octave. In the center is a second group of push-buttons for selection of high-pass, low-pass or band-rejection filters. At the right are several push-button controls for high-frequency equalization or attenuation, at rates of 6, 12, and 24 dB per octave.

The circuitry of this instrument consists of only resistance and capacitance; no inductance is used in any of the equalizer or filter circuits to achieve its characteristics. The elimination of inductance helps to reduce harmonics and intermodulation distortion, eliminate pickup from stray magnetic fields, and permit a good square-wave response, with smooth transition from one set of conditions to another. Each set of push-buttons has in modular form its own set of amplifiers, resistors, and capacitors, divided into seven subassemblies. The device is designed to be connected directly into a microphone line, or used as a bridging input. It can also be used to reproduce a given frequency response, compensate for unwanted response, or be adjusted by listening for the desired characteristic. In stereophonic reproduction its design permits operation in either one or both channels simultaneously, without a phase shift. Due to its design, previous settings may be duplicated accurately.

The frequency response in its linear condition is plus or minus 0.25 dB, from 20 to 20,000 Hz. The input impedance is 5000 ohms balanced; output is 600 ohms. The total rms distortion from 50 to 10,000 Hz is less than 0.4 percent. Maximum output is plus 22 dBm. Intermodulation distortion is less than 1 percent, using 50 and 7000 Hz, mixed in a ratio of 4:1, at plus 18 dBm out. Noise level unweighted is minus 78 dBm, or using the CCIR weighted curve, minus 93 dBm. It may be operated as a zero-gain device, or with a gain of 5 dB. It requires 14 vacuum tubes for its operation.

6.131 Describe a constant-impedance low-frequency attenuator equalizer. —As was pointed out in Question 6.80, capacitor equalizers used to achieve low-frequency attenuation for dialogue recording are not constant impedance. In mixer networks, where a constant impedance is essential, the equalizer circuit in Figs 6-131A and B is employed. This particular design uses a 600-ohm T-type attenuator network; however, the usual shunt-resistor connected in series with the inductance is omitted to achieve the proper frequency characteristic. A key switch is employed to change the value of inductance and capacitance for 6-dB or 8-dB rolloff. The actual configuration for the two key positions is shown in Fig. 6-131B. It will be noted the greatest amount of low-frequency attenuation is attained with the smallest value of capacitance. This may appear to be confusing; however, it should be remembered that the *smaller* the value of a capacitance, the *greater its reactance* at a given frequency. Therefore, the greatest rolloff is attained with the smallest value of capacitance and vice versa. The coil has a total inductance of 0.412 henry, with a tap at 0.25 henry. To prevent pickup

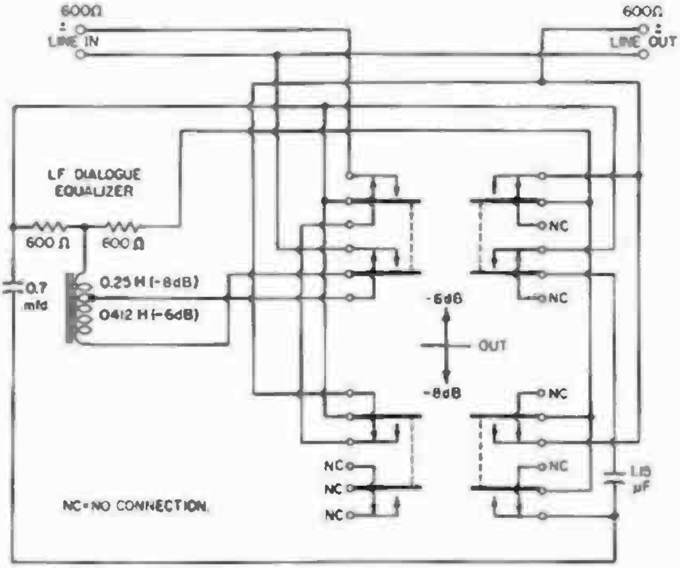


Fig. 6-131A. Low frequency attenuator-type equalizer and its switching circuit.

from stray magnetic fields, a toroidal coil housed in a Mumetal case is necessary.

6.132 *Describe an integrated circuit equalizer-amplifier module.*—It has become common practice in the designing of mixing consoles to provide plug-in units whenever possible to facilitate construction and maintenance. This includes amplifiers, mixer controls, transformers, equalizers, and other devices. An integrated circuit equalizer-amplifier unit Model 709-L manufactured by Electrodyne, is pictured in Fig. 6-132. In this unit a microphone and line amplifier is provided, with both low- and high-frequency equalization. Each equalizer has six positions of equalization and six positions of attenuation, thus providing a total range of 24 dB for each equalizer. Frequencies of equalization are 40 and 100 Hz, 1.5, 3, 5, and 10 kHz, in steps of 2 dB each. In addition, a cuing, echo send, and a level input key to adjust the gain for different type microphones are provided.

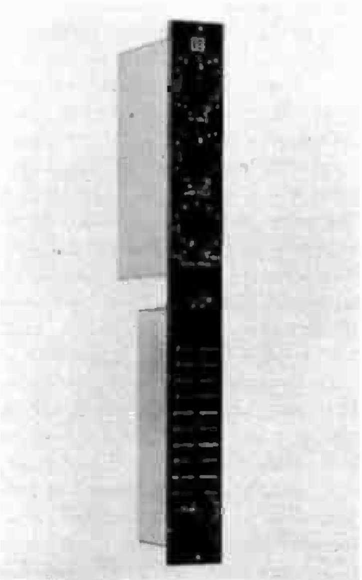
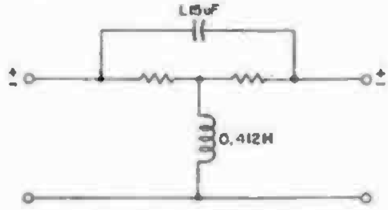
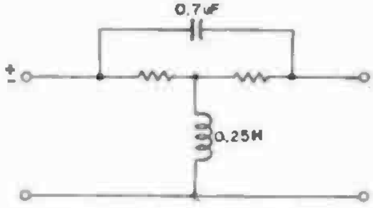


Fig. 6-132. Electrodyne Model 709-L integrated circuit equalizer-amplifier module for mixing consoles.



(*a*) *—6 dB attenuation.*



(*b*) *—8 dB attenuation.*

Fig. 6-131B. Configurations for the two switch positions in Fig. 6-131A.

The microphone and line amplifier are of the differential type. Both inputs of the microphone amplifier are utilized and are driven by a special push-pull input transformer. One input of the line amplifier is used for the signal input and the second is used for equalization. The input impedance is designed for operation with a 50-, 200-, or 600-ohm source impedance. The frequency response is plus or minus 1 dB from 20 to 20,000 Hz, with a THD of 0.5 percent at plus 18-dBm output. The unweighted noise level is equivalent to minus 127 dBm, 20 to 20,000 Hz. Isolation between the cue and program output is greater than 77 dB. The mixer control may be either straight-line or rotary design.

6.133 What type equalizers are used for auditorium tuning?—In reality, these equalizers are tuned filters, consisting of a coil and capacitor connected in parallel, then inserted in a transformer-coupled line, between the voltage and power amplifier of a sound system. The coils must be of fairly high "Q" to permit connection of a resistance in parallel with the coil to broaden the response curve. This subject is discussed in Question 2.117.

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Section 7

Wave Filters

A wave filter is an electrical network of reactive elements designed to transmit or suppress a given band of frequencies. Wave filters are used extensively for recording and reproducing sound. Much of the present day knowledge of these devices came from the research of W. H. Bode, T. E. Shay, and Otto J. Zobel, all of the Bell Telephone Laboratories.

This Section will deal with the design of constant-k, m-, mm′-, mm″-derived high-pass, low-pass, bandpass, band-elimination, and composite filters. Also included are loudspeaker crossover networks, recording and reproducing filters, pink-noise filters, phase-shift filters, parallel-T filters, Wien-bridge filters, noise-measurement filters, and dip filters. The combining of various types of wave filter sections is explained using ladder configurations. Simplified equations, graphs, and tables of constants with many practical circuits are given.

7.1 ***What is a wave filter?***—An electrical network composed of reactive elements used for attenuating or removing a given band of either audio or radio frequencies.

7.2 ***What is an ideal filter?***—A perfect filter designed without regard to losses, physical size, cost, and other factors.

7.3 ***What is a practical filter?***—A physical filter designed from the ideal filter, in which the losses, physical size, cost, and shielding efficiency have been taken into consideration.

7.4 ***What is a passive network?***—One not acting itself, but being acted upon by an external source.

7.5 ***What is an active network?***—One which supplies power. It may be composed of batteries, a generator, or an amplifier.

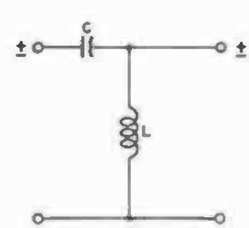
7.6 ***What is a high-pass filter?***—A network of reactive elements which attenuate all frequencies below a predetermined frequency selected by the designer. Frequencies above cutoff are passed without discrimination, as shown in Fig. 7-6.

7.7 ***What is a low-pass filter?***—A network of reactive elements which attenuate all frequencies above a predetermined frequency selected by the designer. Frequencies below cutoff are passed without discrimination. (See Fig. 7-7.)

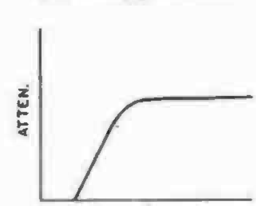
7.8 ***What is a bandpass filter?***—One in which only a predetermined band of frequencies is passed. (See Fig. 7-46.)

7.9 ***What is a band-rejection filter?***—One which rejects or removes a predetermined band of frequencies. (See Fig. 7-47.)

7.10 ***What is a band-elimination filter?***—Another name for a band-rejection filter.

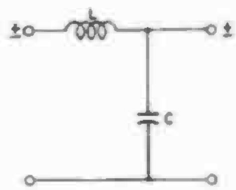


(a) Configuration.

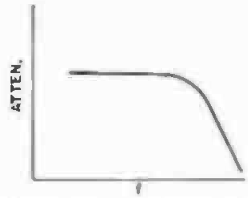


(b) Transmission characteristics.

Fig. 7-6. A high-pass filter.



(a) Configuration.



(b) Transmission characteristics.

Fig. 7-7. A low-pass filter.

7.11 What is an all-pass filter?—A filter which will pass all frequencies without attenuation from zero hertz to infinity. Such devices are used for phase-shifting and time-delay networks. (See Questions 7.23 and 7.24.)

7.12 Describe a notch or dip filter.—It is a filter for removing a given frequency or a very narrow band of frequencies. The device generally consists of an RC network using a parallel-T configuration. (See Question 7.72.) Notch filters are also called dip filters and may use LC or LCR configurations. A dual notch filter manufactured by Altec-Lansing is shown in Fig. 7-12A, with notch frequencies at 150 and 820 Hz. The configuration and frequency response for a similar filter, using 150 and 520 Hz, is shown in Fig. 7-12B.

Fig. 7-12A. Model 64A double-notch or dip filter using 150 and 820-Hz, manufactured by Altec-Lansing.

Such filters can be used for tuning an enclosure, using the Boner system as described in Question 2.117.

7.13 What is a brute-force filter?—One in which the reactive elements are so large that the filtering is forced to take place. This term is generally applied to low-pass filters used in power supplies.

7.14 What is a composite filter?—One which employs two or more filter sections. They may be of constant-k or m-derived design.

7.15 What is the transmission band of a filter?—The portion of its frequency characteristic which passes a band of frequencies without attenuation. (See Fig. 7-15.)

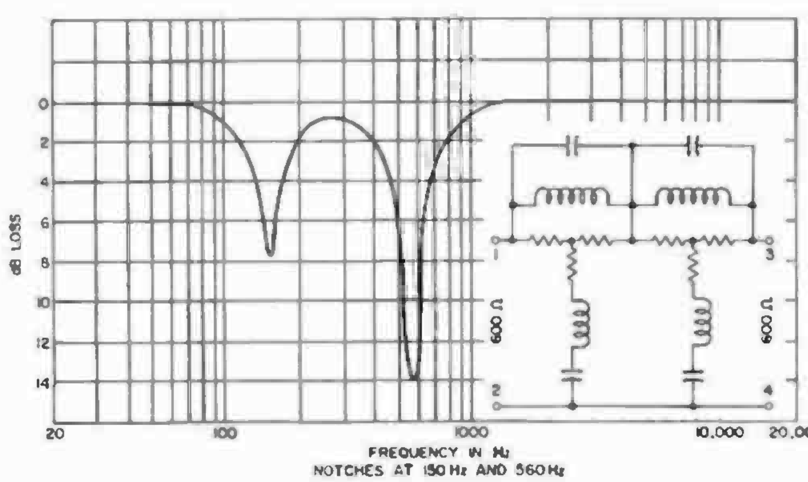
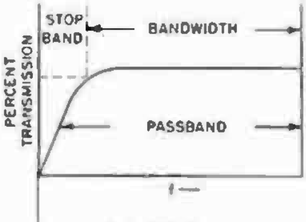
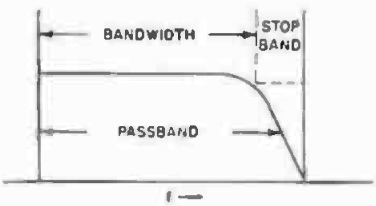


Fig. 7-12B. Frequency response of double-notch filter using 150 and 520-Hz.



(a) *High-pass.*



(b) *Low-pass.*

Fig. 7-15. Transmission characteristics of a high- and low-pass filter.

7.16 What is the passband of a filter?—It is another name for the transmission band. (See Question 7.15.)

7.17 What is the bandwidth of a filter?—The number of hertz expressing the difference between the lower and upper limiting frequencies. It is also called the passband. (See Fig. 1-15.)

7.18 What is a stop band?—The frequency of a band-rejection or band-elimination filter.

7.19 What is the cutoff frequency of a filter?—The frequency above or below which a filter fails to respond. The terminology used for expressing the characteristics of filters is discussed in Question 7.102.

7.20 What is a two-terminal network?—A configuration such as shown in Fig. 7-20.

7.21 What is a three-terminal network?—An unbalanced configuration with one terminal common to both the input and output. (See Fig. 7-21.)

7.22 What is a four-terminal network?—A balanced configuration as shown in Fig. 7-22.

7.23 What is a phase-shift or phase-correction network?—An electrical network composed of reactive elements as shown in Fig. 7-23. The purpose of such networks is to induce a time delay in a given frequency band to correct for phase distortion in long-line telephone transmission. They are also used in recording circuits to compensate for phase shift of certain type equalizers.

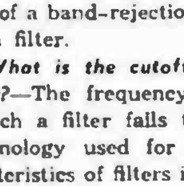


Fig. 7-20. A two-terminal network.

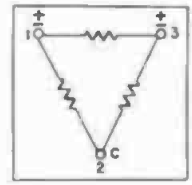
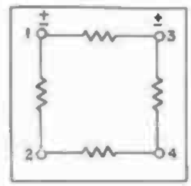
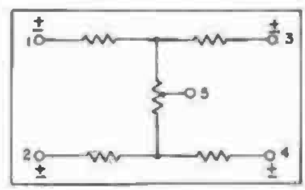


Fig. 7-21. A three-terminal network.

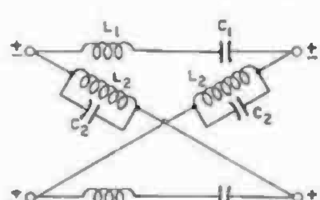


(a) *Without ground terminal.*

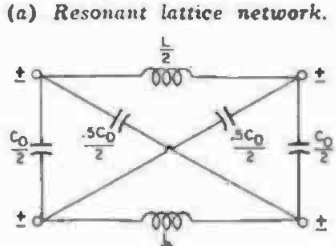


(b) *With ground terminal.*

Fig. 7-22. Four-terminal networks.



(a) *Resonant lattice network.*



(b) *Single nonresonant lattice network.*

Fig. 7-23. Configurations of balanced phase-correction networks.

A lattice (balanced) configuration phase-correction network employing both series and parallel resonant circuits is shown at (a) of Fig. 7-23. Each complementary pair is resonated to a selected frequency band to induce a given time-delay. The circuit shown at (b) of Fig. 7-23 is a similar type network using nonresonant circuits for high-frequency phase correction. (See Question 7.112.)

7.24 ***How are wave filters classified?***—Wave filters are classified according to their function and the procedure used for determining their circuit element values. The four design procedures are: the constant-k (prototype); the m-derived; the mm' (double-m, prime derived); and the mm" (double-m, double-prime derived). The latter two are used only in advanced filter design where the characteristics must be closely controlled. For audio work, usually a constant-k or an m-derived type is used, or a combination of both types. Such a filter is termed a composite filter.

7.25 ***What is the characteristic impedance of a wave filter?***—Its design impedance, as designated by the symbol R_o. A typical impedance curve for a constant-k type filter is shown in Fig. 7-25. The actual impedance, Z_1, presented by the filter to the line is equal to the line impedance R_o only at one frequency and is one of the characteristics of a constant-k filter.

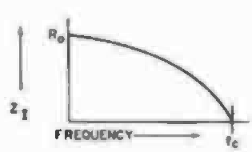


Fig. 7-25. Impedance characteristic of a constant-k filter.

7.26 ***What is surge impedance?***—The terminating impedance required at the end of a transmission line to prevent reflection losses.

7.27 ***What is the terminal impedance of a filter?***—The impedance seen when looking into the input or output terminals when measured with an impedance bridge.

7.28 ***What does the term constant impedance mean?***—Any device which maintains a fixed terminal impedance under normal operating conditions.

7.29 ***What is a pi-type filter?***—A filter which has a configuration resembling the Greek letter pi. (See Fig. 7-39.)

7.30 ***What is a constant-k filter?***—A wave filter in which a constant termed k is used. This term is sometimes misconstrued as meaning that the impedance of the filter is constant; however, this is not the case. The impedance of a constant-k filter is equal to the line impedance at only one frequency and presents a mismatch at all other frequencies, as described in Question 7.25.

7.31 ***What is an m-derived filter?***—A wave filter in which the impedance and attenuation characteristics may be controlled by the designer. The term m is a constant lying between zero and one. An m-derived filter can be designed to match approximately 85 percent of the frequency band covered by the filter. For audio frequencies, the constant m is generally made to equal 0.60. An m-derived filter may also be combined with a constant-k section to secure a given frequency characteristic. When a constant-k and an m-derived filter section are combined, the filter is called a composite filter.

7.32 ***What is the insertion loss of a filter?***—It is the loss in level measured at a given frequency in the passband, with the filter in and out of the circuit. The insertion loss of a filter (in decibels) may be found by determining the amount of the current reduction at the load side of the network. As a general practice, the insertion loss of any network is measured at a frequency within the *flat* portion of the passband; the exact frequency depends on the individual characteristics of the network. The expression for calculating the insertion loss is:

$$IL = 20 \mathrm{Log}_{10}\left(\frac{E_1}{E_2}\right) - 20 \mathrm{Log}_{10}\left(\frac{R_{in} + R_{out}}{R_{out}}\right)$$

where,

E_1 and E_2 are the voltages at the inpnt and output terminals,
R_{in} and R_{out} are the impedances of the network at its terminals.

This formula takes into consideration the *mismatch* of impedances at either end of the network and the effect of *series and shunt reactances* in the network.

7.33 What causes insertion loss?—The dc resistance of the coils and the series and shunt reactance of the circuit elements.

7.34 How can the impedance of an unknown filter be measured?—By measuring the geometric mean of the filter impedance in the passband at a given frequency. The impedance is first measured with the far end shorted, then open. The characteristic impedance R_o is then equal to:

$$R_o = \sqrt{Z_1 \times Z_2}$$

where,

Z_1 is the first measurement,
Z_2 is the second measurement.

7.35 What is the transmission characteristic of a wave filter?—It is the frequency characteristic of the filter plotted as frequency versus attenuation and is the characteristic seen by the circuit in which it operates (Fig. 7-15).

7.36 What are the attenuation characteristics of a wave filter?—The loss of a filter plotted as frequency versus insertion loss in decibels.

7.37 Define the term, "image impedance."—The image impedance of a wave filter is presented by a group of curves, each curve having a different impedance characteristic, depending on the character of the configuration. (See Fig. 7-48A.) The image impedance consists of two impedances, not necessarily alike, that will simultaneously terminate the network at its two ends, thereby avoiding internal reflection losses. This occurs when the image impedance is equal to the impedance looking into the filter network. When the image impedances are equal in value (input and output), the filter is symmetrical and is equal to Z_1, the characteristic impedance of the network.

The term impedance can best be explained by means of Fig. 7-37A. The image impedances are equal to Z_1 and Z_2 if the termination impedances are of such value that, with Z_1 disconnected, the impedance looking into the input of the filter equals Z_1; with Z_1 connected and Z_2 disconnected, the impedance

Fig. 7-37A. Filter network terminated in its image impedances.

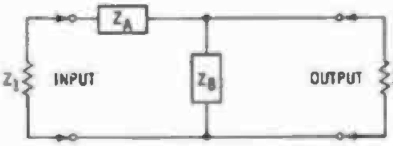


Fig. 7-37B. Elementary filter network terminated in its image impedances.

looking into the output terminals equals Z_1. Thus, the term image impedance may be explained as follows; looking into the input terminals Z_1 sees its image, and looking back into the network Z_2 sees its image. This is the condition when a filter network is said to be terminated in its image impedance. Mathematically the image impedance, with Z_1 disconnected and Z_2 connected is:

$$Z_{in} = Z_A + \frac{(Z_B Z_2)}{(Z_B + Z_2)}$$

Disconnecting Z_2 and connecting Z_1, the image impedance is:

$$Z_{out} = \frac{(Z_A + Z_1)\, Z_B}{(Z_A + Z_1 + Z_B)}$$

Therefore,

$$Z_1 = \sqrt{(Z_A + Z_B)\, Z_A}$$

$$Z_2 = Z_B \sqrt{\frac{Z_A}{(Z_A + Z_B)}}$$

where,

Z_1, Z_2, Z_A and Z_B are as shown in Fig. 7-37B.

Mathematically, the foregoing gives the image impedance for any given set of circuit constants. However, in design of practical filters, the image impedance is known in advance, and the circuit constants are then computed to result in the desired image impedance. The image or characteristic impedance of an existing filter may be measured as given in Question 7.34.

7.38 What is meant when it is said two impedances are conjugates of each other?—When their resistive components and their reactive components are equal in magnitude but opposite in sign.

7.39 What are the basic configurations and equations for constant-k low pass filters?—The basic designs for constant-k low-pass filter sections are shown in Fig. 7-39. It will be noted that a T-type network has an inductance (L_1) in series with the line and a capacitance (C_2) in shunt with the line. Operation of the network may be visu-

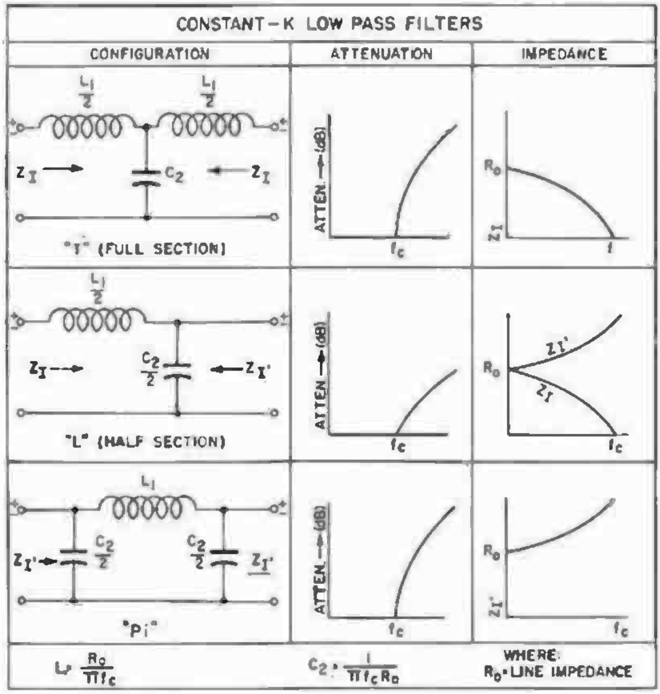


Fig. 7-39. Configuration and frequency characteristics of a constant-k low-pass filter.

alized by consideration of the properties of inductive and capacitive reactance.

Inductive reactance varies directly with frequency. Since an inductive reactance is connected in series with the line, it offers an increasing opposition to transmission as the frequency is increased. The capacity in shunt with the line presents a capacitive reactance which decreases with frequency. As the frequency is increased, this capacitive reactance becomes more effective in shunting the audio currents, thereby reducing the transmission through the filter. The combination of the inductive and capacitive reactance produces a net attenuation characteristic beginning at the cutoff frequency and continuing beyond the cutoff frequency. The impedance presented by a T network to the transmission line is designated Z_I. This impedance is equal to the line impedance at zero frequency only and decreases progressively through the passband. This may be seen by referring to the impedance curve for a T section (Fig. 7-39).

An L-type network may be thought of as one-half of a T section or one-half of a pi section. An L section is also called a half-section and has an attenuation characteristic which is half as much as the attenuation of a full section (T or pi) at the same frequency. The impedance of an L filter is designated Z_I and Z_I'. These two designations are used because the impedance of the filter at one set of terminals differs from the impedance at the other. It will be seen that at one end the impedance is the same as for the T network and, at the other, the impedance characteristic is the same as that of the pi network.

The attenuation characteristics of a pi section can be made identical to those of a T section, as indicated by the attenuation curves. The attenuation of either a pi or a T section is twice as much as for the L or half-section network. The impedance of the pi section is designated Z_I' for both ends. A constant-k pi section presents an impedance to the line which equals the line impedance at zero frequency and increases as the frequency increases within the passband.

7.40 How are the different filter sections selected?—By their attenuation and impedance characteristics. The attenuation of a constant-k full section is approximately 23 dB, one octave removed from the cutoff frequency. This is not considered to be a sharp rate of cutoff. If the attenuation is not sufficient, combinations of filter sections may be connected in tandem.

7.41 How are different type filter sections combined?—To better understand how filter sections are combined, in Fig. 7-41A is shown the familiar ladder-type network with its series reactive element Z_1 and shunt reactive element Z_2. These reactive elements (impedance) may be either capacitive or inductive. If two lines, A and B (Fig. 7-41B) are drawn through the midpoint of the two series elements, the familiar T-section is developed. In this configuration the series arms are denoted $Z_1/2$, since they are a half of Z_1 in Fig. 7-41A, and they are referred to as *midshunt* terminals. Referring again to Fig. 7-41B, drawing a line at C, the structure between B and C becomes an L-type configuration. In this instance the series arm is also $Z_1/2$, while the shunt arm is $2Z_2$, and the terminals are termed midshunt. Cutting the network with a line D, the configuration between C and D becomes a pi-type configuration. The series element is Z_1, while the shunt element is $2Z_2$. For the element (Z_2) to equal Z_2 in Fig. 7-41A, its value must be *doubled*. In practice, when two elements appear in series or parallel, they are combined into one element.

The input terminals of the L-type configuration, Fig. 7-41B are termed *midseries*, while the output terminals are called midshunt. An L section is also referred to as a *half-section*, since two L sections connected in tandem form a T type or *pi section*, depending on whether they are joined at their midshunt or midseries terminals. The method used for combining the elements of two T-type configurations is shown in Fig. 7-41C. Combining two pi sections is shown in Fig. 7-41D.

Any number of filter sections may be combined in tandem to secure the desired attenuation. A filter of this design is referred to as a composite filter. For

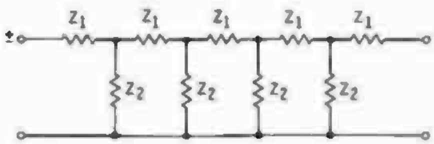


Fig. 7-41A. Ladder network used in basic design.

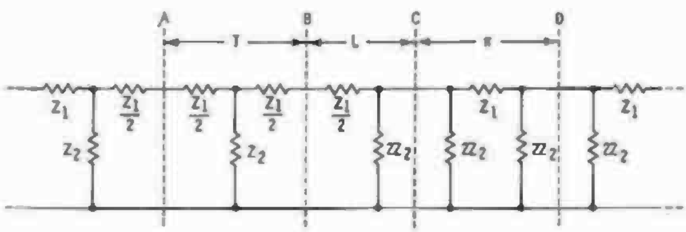


Fig. 7-41B. Ladder network broken down into filter configurations.

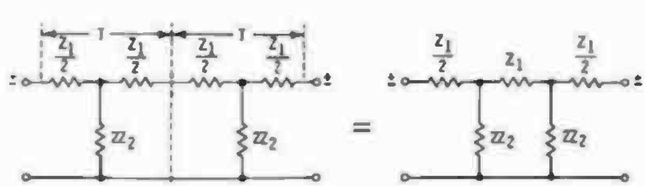


Fig. 7-41C. Combining two T filters.

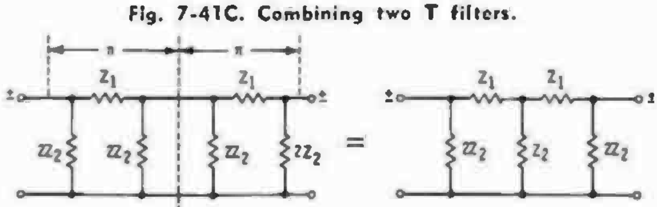


Fig. 7-41D. Combining two pi filters.

audio frequency work, generally, two sections are sufficient. In assembling a composite filter, sections with similar impedance characteristics are connected in tandem. This is important in order that one section may properly terminate the other. It is apparent that none of the sections present a constant impedance; however, if two sections whose impedances vary in exactly the same manner are connected in tandem, then each will properly match the other at all frequencies. Thus, within a composite filter there are no mismatches of impedance and, therefore, no reflection losses. This is accomplished by observing the terminology Z_1 to Z_1, Z_1' to Z_1', etc.

From the characteristics given in Fig. 7-39 it is evident that a T section cannot be connected to a pi section. However, an L section may be connected between them to match the impedances in both directions. Although an internal impedance match can be obtained in a composite filter, none of the configurations matches the line impedance through the passband. This is the disadvantage of the constant-k filter.

7.42 How are the circuit elements combined in a composite filter?—A T section and an L section, connected in tandem, are shown in Fig. 7-42. In the actual physical filter, the right-hand inductance of the T section and the inductance of the L section are combined into one inductance. Whenever two capacitors appear in parallel adjacent to one another, they are also lumped into one unit.

7.43 What are the basic considerations when designing a filter?—To design a constant-k filter, the following information is required:

Line impedance,

Filter type (low-pass, high-pass, etc.),

Tolerable mismatch for R_o.

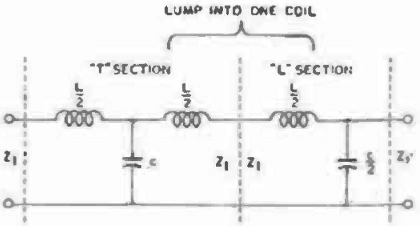


Fig. 7-42. Method of combining two filter sections.

The line impedance (R_o) is the impedance of the circuit in which the filter is to operate. It will be noted that, if the value of the inductance has been calculated for a pi section, the value is divided by two when used in the L or T sections. However, in the pi section, the value of the inductance remains unchanged. It will also be noted that the calculated value of the capacity for a T section is divided by two for the L and pi sections.

7.44 Show the design procedure for a 500-ohm constant-k 10,000 Hz low-pass filter which is required to have at least 40 dB of attenuation at 20,000 Hz.—Two full sections will have approximately 46 dB of attenuation, at twice the cutoff frequency; therefore, two such sections using the T configuration will be used. The first step is to calculate the values of L and C:

$$L = \frac{R_o}{\pi f_c}$$

$$= \frac{500}{3.141 \times 10{,}000}$$

$$= 0.0159 \text{ henries}$$

$$C = \frac{1}{\pi f_c R_o}$$

$$= \frac{1}{3.141 \times 10{,}000 \times 500}$$

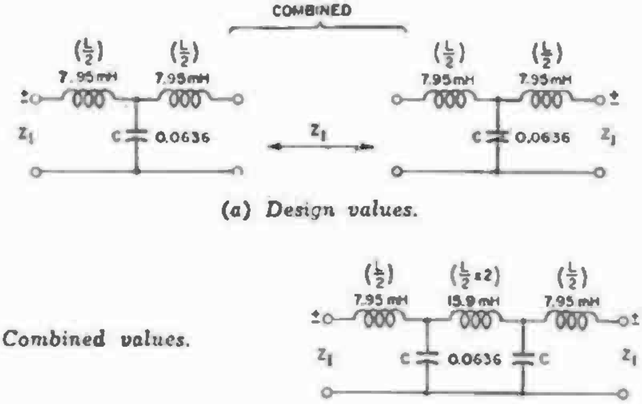
$$= 0.0636\ \mu F$$

where,

R_o is the line impedance,
f_c is the cutoff frequency.

The two sections of the filter are shown at (a) in Fig. 7-44. At (b) in Fig. 7-44 the circuit element values have been combined and one coil is used in place of the two nsed at the output and the input of the separate filters shown at (a).

7.45 What are the basic configurations and equations for constant-k high-pass filters?—The basic designs for constant-k high-pass filter sections are shown in Fig. 7-45. It will be noted the positions of the inductance and capacity are inverse to those of the low-pass filter. The design equations appear at the bottom of the configurations. A full section will provide approximately 23 dB of attenuation one octave removed from the cutoff frequency. A half section or L configuration will provide one half the attenuation of a full section at any frequency. The various sections may be connected in tandem to



(a) *Design values.*

(b) *Combined values.*

Fig. 7-44. Method of combining filter element values.

form a composite filter. As for the low-pass filter, the impedance match must be made so that each section is properly terminated. This is accomplished by observing the impedance characteristics for each type section. The procedure for designing and combining high-pass, constant-k filters is the same as that described for the low-pass filter in Questions 7.39 and 7.40.

7.46 What are the configuration and equations for designing a constant-k bandpass filter?—The configuration for a bandpass filter and its equations are shown in Fig. 7-46. It will be observed that these filters utilize resonant circuit arms, both in series and in shunt with the line. The frequencies f_1 and f_2 in the equations are frequencies at the edges of the passband. The frequency f_m is the frequency in the center of the passband. The frequencies of $f_1\infty$ and $f_2\infty$ are the frequencies at the edge of the widest part of the passband.

7.47 What are the configuration and equations for designing a constant-k band-rejection filter?—The configuration for a band-rejection or band-elim-

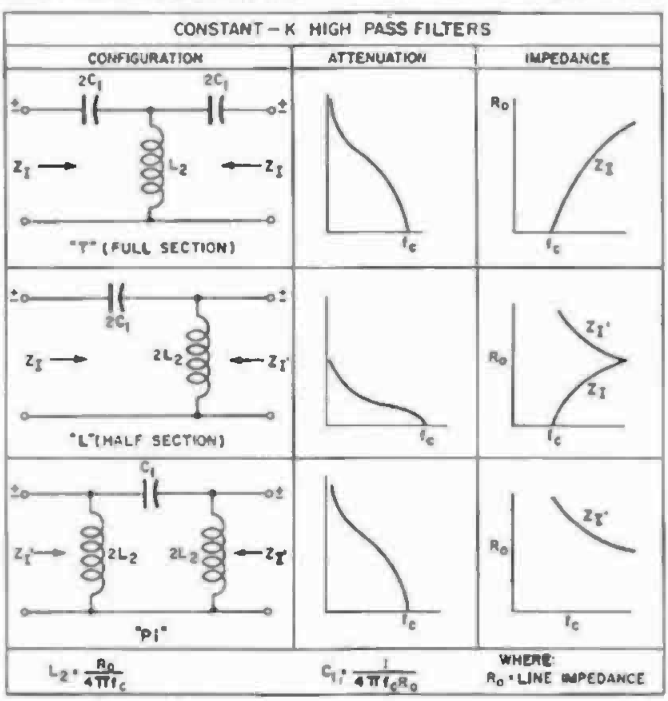
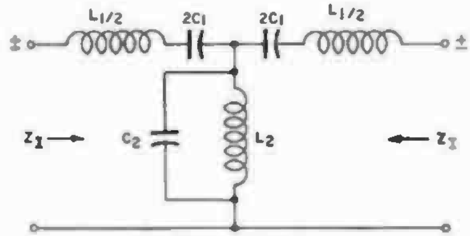
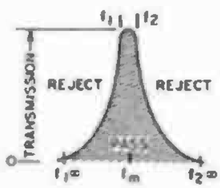


Fig. 7-45. Configurations and frequency characteristics of a constant-k high-pass filter.



(a) Configuration.



(b) Transmission characteristics.

$$L_1 = \frac{R_0}{\pi(f_2-f_1)}$$

$$L_2 = \frac{(f_2-f_1)R_0}{4\pi f_1 f_2}$$

$$C_1 = \frac{(f_2-f_1)}{4\pi f_1 f_2 R_0}$$

$$C_2 = \frac{1}{\pi(f_2-f_1)R_0}$$

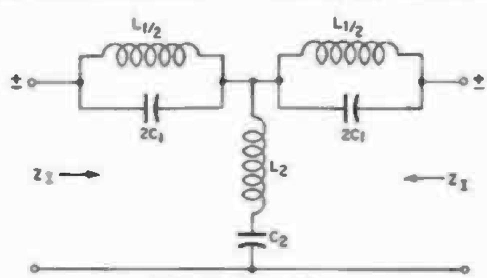
(c) Equations.

Fig. 7-46. Constant-k bandpass filter.

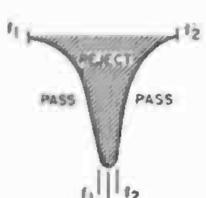
ination filter is shown in Fig. 7-47. It will be noted that the configuration is reversed from that of the bandpass filter. For this filter, the frequencies $f_1\infty$ and $f_2\infty$ are frequencies at the edge of the reject band. Frequencies f_1 and f_2 are at the edge of the widest part of the reject band, and f_m is the center frequency of the reject band.

7.48 What are the basic principles of an m-derived filter?—An m-derived filter is so designed that either the impedance or the attenuation characteristic, but not both, may be controlled by the designer. This overcomes some of the objection to the constant-k type filter. An m-derived filter is designed by first calculating the values of capacitance and inductance for the constant-k type filter and then modifying these values by an algebraic expression containing the term m. The term m is a positive number lying between one and zero. The value of m governs the characteristic of the m-derived filter.

In certain type sections m governs the impedance. By the proper selection of m it is possible to match the line impedance over approximately 85 percent of the transmission band, which is



(a) Configuration.



(b) Transmission characteristics.

$$L_1 = \frac{(f_2-f_1)R_0}{\pi f_1 f_2} \qquad L_2 = \frac{R_0}{4\pi(f_2-f_1)}$$

$$C_1 = \frac{1}{4\pi(f_2-f_1)R_0} \qquad C_2 = \frac{(f_2-f_1)}{\pi f_1 f_2 R_0}$$

$$f_m = \sqrt{f_1 f_2} \qquad R_0 = \sqrt{\frac{L_1}{C_2}} = \sqrt{\frac{L_2}{C_1}}$$

(c) Equations.

Fig. 7-47. Constant-k band-rejection filter.

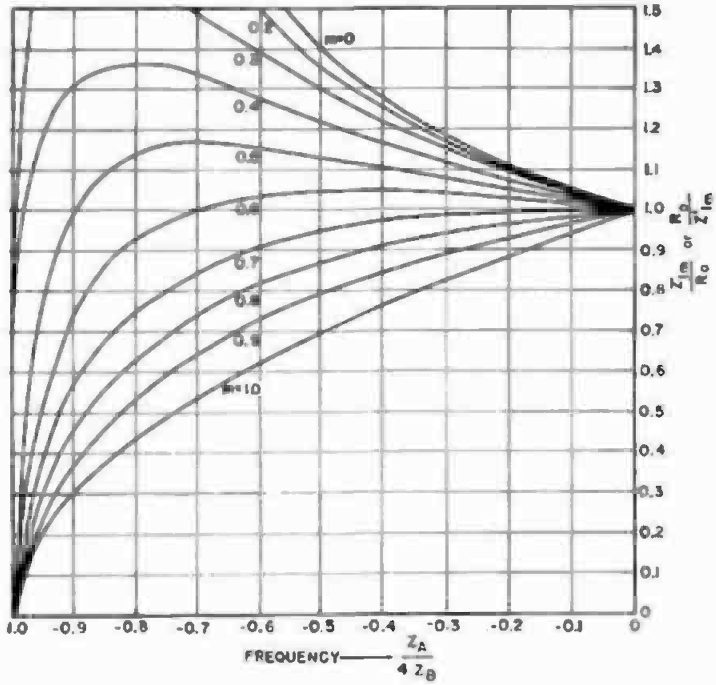


Fig. 7-48A. Impedance characteristics of m-derived filters, for different values of m.

considerably better than the constant-k type filter. An m-derived filter employs resonant circuits in the series and shunt arms. Theoretically, the filter presents an infinite attenuation termed f∞ (infinity).

In the design of an m-derived filter two frequencies are involved, the cutoff frequency and the frequency of infinite attenuation. The term m may be used to control the spacing between the frequency of cutoff and that of attenuation. Fig. 7-48A shows the effect on the impedance for different values of m between zero and one. It will be noted that the best impedance match is obtained when m equals 0.60. This is the value generally used for audio-frequency filters. It should be understood these data are for an ideal filter and

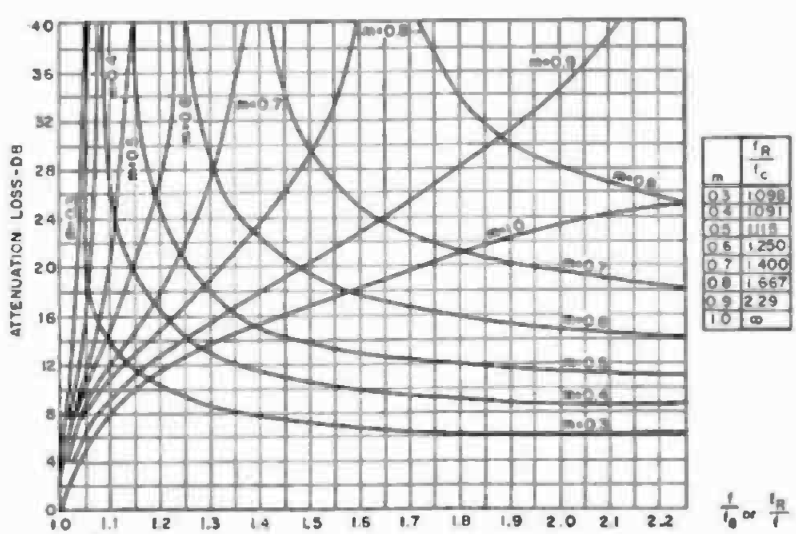


Fig. 7-48B. Attenuation characteristics of m- derived filter sections.

will vary in the practical filter due to the components, particularly the inductances. Attenuation characteristics for different values of m are shown in Fig. 7-48B. It will be noted from this family of curves that the attenuation rises to a maximum and then decreases. This is due to the resonant circuit incorporated in the m-derived filter. The graph in Fig. 7-48B is plotted to show the attenuation losses. This ratio may be inverted, if necessary, to make the number greater than one. The graph may be used for either low- or high-pass filters.

7.49 ***What are the advantages of combining a constant-k filter with an m-derived filter?***—The attenuation of a constant-k network rises progressively as the frequency departs from cutoff. The attenuation of an m-derived type rises to a maximum and then falls off. In a composite filter, the m-derived section governs the initial rate of attenuation and the constant-k section maintains attenuation at frequencies where the m-derived section becomes less effective.

It may be desirable to incorporate a full or half section at each end of the filter having an impedance characteristic controlled by m and for which m is 0.60. This will provide a match to the line at either end of the filter. The block diagram of a composite filter of this type is shown in Fig. 7-49. This filter has four sections. The attenuation characteristic for each section is shown below the block. The total attenuation is the sum of the individual attenuation of each section. Because the end sections have been designed using a value of 0.60 for m, they present to the line a very nearly constant impedance, R_o, over most of the passband. Configurations which match each other have been chosen so that at every junction of the sections an impedance match exists.

7.50 ***What is the effect of an m-derived filter if m equals one?***—The filter degenerates into a constant-k type. Therefore, it may be said that a constant-k filter is a special case of the m-derived filter with m equal to one.

7.51 ***What sections of an m-derived filter have the same impedance characteristics as a constant-k section?***—The T section, series m-derived. The series m-derived family consists of the T, L, and the pi sections.

7.52 ***Show the configurations and equations for a series m-derived low-pass filter.***—The configurations and equations appear in Fig. 7-52A. It will be noted the configurations are similar to those of the constant-k type filter, except series resonant circuits are used in the shunt arms. The terms $L_{1(K)}$ and $C_{2(K)}$ in the equations below the configurations are the values obtained for a constant-k filter. These values are then used with the m-derived equations given for L_1, L_2, and C_2 to calculate their values. A table of the most commonly used factors employed in the design of m-derived filters is given in Fig. 7-52B.

7.53 ***How is the constant "m" determined?***—As previously stated, m is a number between zero and one and controls the point of maximum attenuation with reference to the cutoff frequency of a filter. By the proper selection of m, the rate of attenuation for a given section can be determined; that is, it can be designed to attenuate gradually beyond the cutoff point or it can be designed to cut off sharply. To determine the value of m, the cutoff frequency and frequency of infinite at-

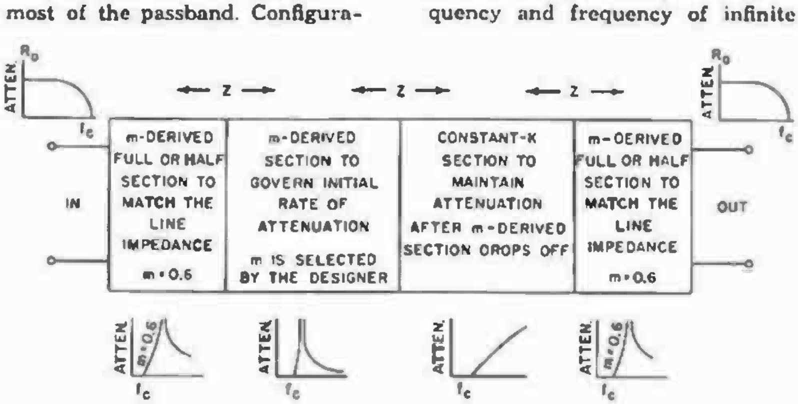
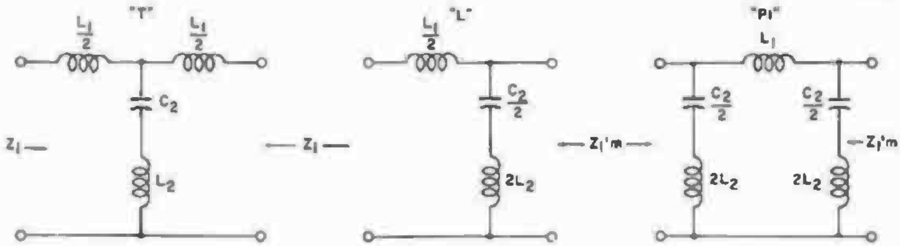


Fig. 7-49. A composite filter consisting of constant-k and m-derived sections.



$$L_{1(K)} = \frac{R_0}{2\pi f_c} \qquad C_{2(K)} = \frac{1}{\pi f_c R_0}$$

$$L_1 = mL_{1(K)} \qquad L_2 = \frac{1-m^2}{4m} L_{1(K)} \qquad C_2 = mC_{2(K)}$$

Fig. 7-52A. Configurations and design equations for series *m*-derived low-pass filters.

tenuation (maximum) must be known. The values may then be substituted in the equation:

$$m = \sqrt{1 - \left(\frac{f_c}{f\infty}\right)^2}$$

$$m = \sqrt{1 - \left(\frac{f\infty}{f_c}\right)^2}$$

where,

f_c is the cutoff frequency in hertz,
$f\infty$ is the frequency of maximum attenuation.

The first equation is for low-pass filters and the second is for high-pass filters. The constant m is always less than one.

7.54 What are the configurations and equations for series m-derived high-pass filters?—The configurations with their equations are given in Fig. 7-54.

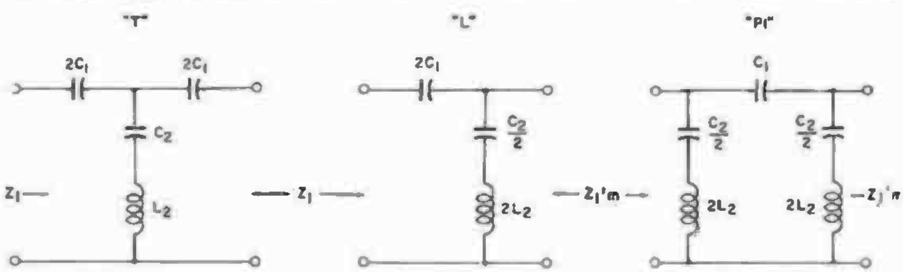
7.55 What are the configurations and equations for shunt m-derived low-pass filters?—The configurations with their equations are given in Fig. 7-55.

7.56 What are the configurations and equations for shunt m-derived high-pass filters?—The configurations with their equations are given in Fig. 7-56.

7.57 What are the advantages of the shunt-type m-derived filter over the series m-derived type?—Electrically there are none. However, the two types of design permit a wide selection of circuits. In building a filter, it is more economical to use the series-derived

m	$1-m^2$	$\frac{4m}{1-m^2}$	$\frac{1-m^2}{4m}$	$4\pi m$
0.10	0.990	0.404	2.475	1.256
0.15	0.987	0.613	1.630	1.884
0.20	0.960	0.833	1.200	2.512
0.25	0.938	1.066	0.938	3.140
0.30	0.910	1.318	0.758	3.768
0.35	0.878	1.593	0.627	4.396
0.40	0.840	1.904	0.525	5.024
0.45	0.798	2.255	0.443	5.652
0.50	0.750	2.666	0.375	6.280
0.55	0.698	3.151	0.317	6.908
0.60	0.640	3.758	0.266	7.536
0.65	0.578	4.498	0.222	8.164
0.70	0.510	5.490	0.184	8.792
0.75	0.438	6.849	0.146	9.420
0.80	0.360	8.888	0.112	10.048
0.85	0.278	12.230	0.081	10.676
0.90	0.190	18.940	0.052	11.304

Fig. 7-52B. Commonly used factors for the design of m-derived filters.



$$L_{2(K)} = \frac{R_0}{4\pi f_c} \qquad C_{1(K)} = \frac{1}{4\pi f_c R_0}$$

$$C_1 = \frac{C_{1(K)}}{m} \qquad L_2 = \frac{L_{2(K)}}{m}$$

Fig. 7-54. Configurations and design equations for series m-derived high-pass filters.

type because of the cost of an inductance compared to a capacitor. In certain phases of telephone work, it is desirable that the filter have a capacitor input rather than an inductive input. The two designs will permit either type input to be used, with the same frequency characteristics.

7.58 Show a typical design problem for a 600-ohm, 1400 Hz shunt m-derived low-pass filter.—Basically, the filter will consist of a constant-k pi section with two shunt m-derived end sections as shown in Fig. 7-58A. The calculations have been tabulated to demonstrate the design procedure in Fig. 7-58B. The derived values are given in Fig. 7-58C and the final values used are given in Fig. 7-58D.

7.59 What is a balanced filter?—Filters, like attenuators and equalizers, may be designed for balanced and unbalanced operation. In Fig. 7-59 the filter of Fig. 7-44(b) is shown converted to a balanced configuration. It will be noted the values of the inductances have been divided by two, and one half placed in each side of the line. The capacitor values remain unchanged.

7.60 How is an unbalanced high-pass filter converted to a balanced configuration?—To illustrate the procedure, Fig. 7-60 shows a 45-Hz high-pass filter both as an unbalanced and a balanced

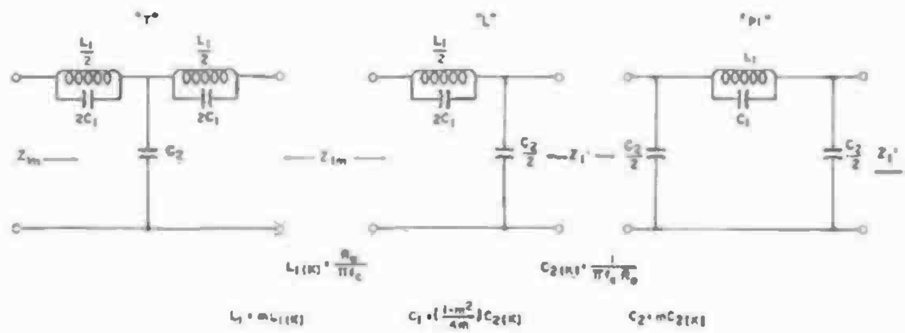


Fig. 7-55. Configurations and design equations for shunt m-derived low-pass filters.

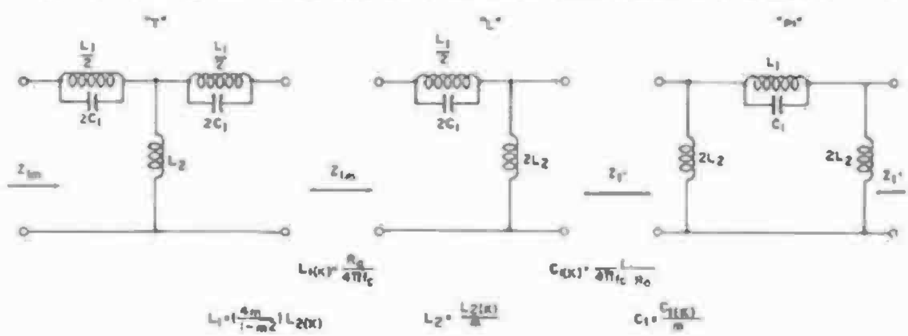


Fig. 7-56. Configurations and design equations for shunt m-derived high-pass filters.

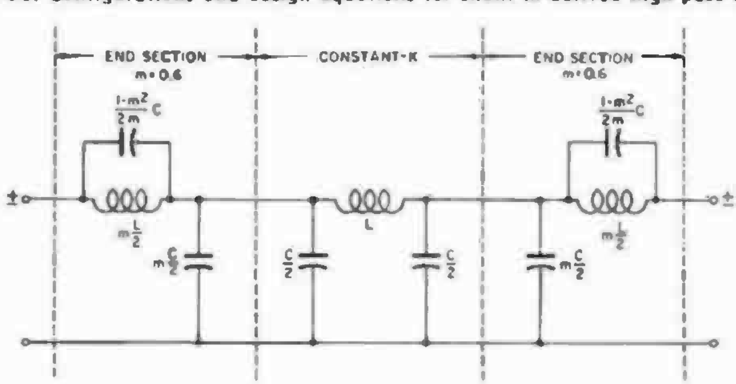


Fig. 7-58A. Basic design of composite shunt m-derived low-pass filter.

L	$= \frac{R_o}{\pi f_c} = \frac{600}{\pi(1400)} =$	0.1336 H
C	$= \frac{1}{\pi f_c R_o} = \frac{1}{\pi(1400)(600)} =$	0.379 μF
$\frac{L}{2}$	$= \frac{0.1366}{2} =$	68.3 mH
$\frac{C}{2}$	$= \frac{0.379}{2} =$	0.189 μF
m	$=$	0.600
$\frac{1-m^2}{2m}$	$= \frac{1-0.36}{1.2} = \frac{0.64}{1.2} =$	0.533
$\frac{1-m^2}{2m}C$	$= (0.533)(0.379) =$	0.202 μF
$m\frac{L}{2}$	$= (0.600)(68.3) =$	41.0 mH
$m\frac{C}{2}$	$= (0.600)(0.189) =$	0.114 μF

Fig. 7-58B. Design data for 600-ohm, 1400 Hz, shunt m-derived low-pass filters.

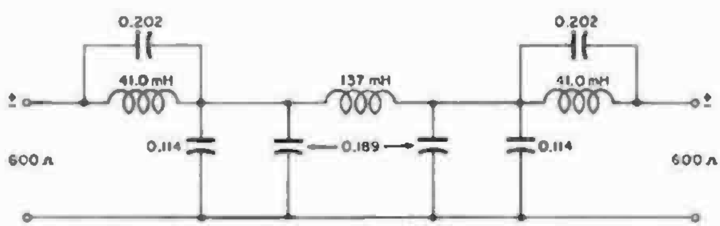


Fig. 7-58C. Design values for the low-pass filter shown in Fig. 7-58A.

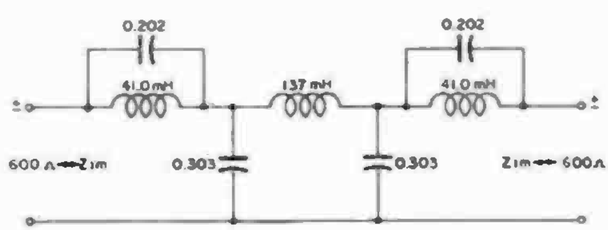


Fig. 7-58D. Composite filter or final values for the filter shown in Fig. 7-58A.

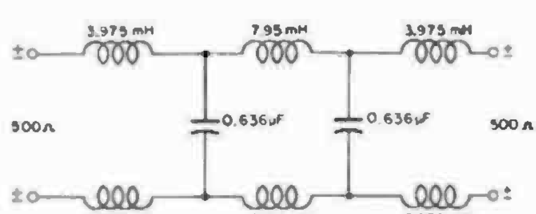


Fig. 7-59. The low-pass filter shown in Fig. 7-44(b) converted to a balanced configuration.

configuration. In the balanced configuration it will be noted that the inductances remain unchanged in value but the capacitor values have been doubled. The reason for this is that the capacitors of the balanced configuration may be considered to be in series in each side of the line. Therefore, to maintain the same impedance, the capacitance must be doubled because capacitors in series divide themselves.

Filters, like equalizers, are designed to present a given impedance to certain frequency bands and to accomplish this a definite value of capacity and inductance is required. If the circuit element values are maintained regardless of the type configuration employed, the impedance presented to the circuit is the same; therefore, the frequency characteristic will be the same.

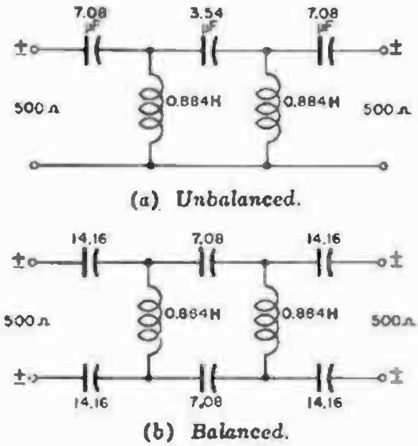


Fig. 7-60. A 45-Hz unbalanced high-pass filter converted to a balanced configuration.

7.61 How are the frequency characteristics of a filter measured?—This subject is discussed in detail in Question 23.61.

7.62 How may a filter of known characteristics be converted to an impedance other than the one for which it was originally designed?—By the use of the following equations, if the circuit constants are known.

$$L_x = \frac{Z_1}{R_o} \times L$$

$$C_x = \frac{C}{\frac{Z_1}{R_o}}$$

where,

L_x is the new value of inductance,
L is the original value of inductance,
R_o is the characteristic impedance,
Z_1 is the new impedance,
C_x is the new value of capacitance,
C is the original value of capacitance.

7.63 How may a filter of known frequency characteristics and components be converted to another frequency?—By changing the values of the inductance and capacitance inversely with frequency. For example, if a 1000-Hz low-pass filter is to be converted to 100 Hz, the new values may be calculated:

$$C_x = C \times 10$$

$$L_x = L \times 10$$

where,

C_x is the new value of capacity,
C is the original value of capacity,
L_x is the new value of inductance,
L is the original value of inductance.

Although the frequency is changed, the impedance remains the same.

7.64 What type coils are recommended for filter construction?—Toroidal coils because of their high Q and the fact that they are not affected by extraneous magnetic fields. However, conventional coils may be used, if properly shielded. (See Questions 6.68 to 6.71 and Questions 8.72 and 8.73.)

7.65 What tolerance should be specified for coils and capacitors used in filters?—For coils, plus or minus two percent; for capacitors, plus or minus three percent.

7.66 How are the coils for filters specified for manufacture?—By the value of the inductance, the signal level at which the coil is to operate, the frequency of maximum Q, the filter configuration, the impedance of the circuit, and the effective shielding of the coils in decibels.

7.67 If a low-pass and a high-pass filter are connected in tandem, what is the overall frequency characteristic?—A bandpass characteristic.

7.68 What is the overall frequency response of two low-pass filters connected in tandem?—The frequency response is the algebraic sum of the individual frequency characteristics. This is explained in more detail in Question 6.66.

7.69 What is the center frequency of the passband and how is it determined?—The center frequency of the passband of any type filter, or combination of filters, is the geometric mean of

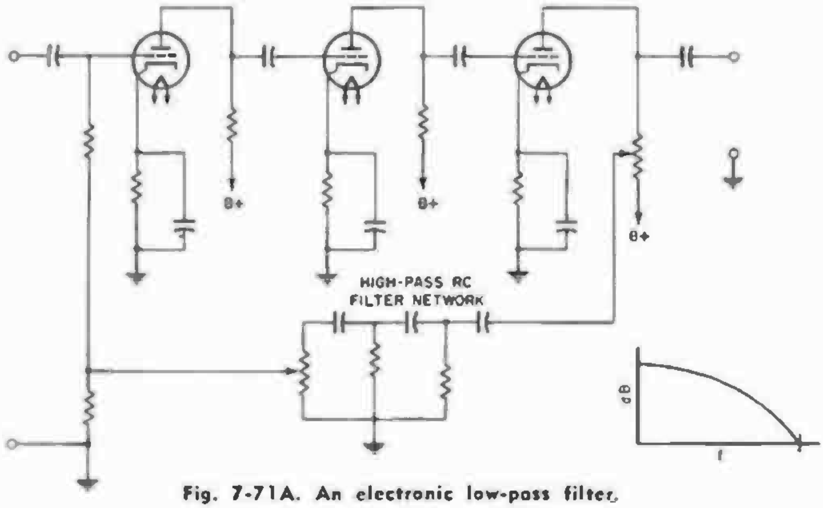


Fig. 7-71A. An electronic low-pass filter.

the lowest and highest frequency of the passband. A typical example would be a recording channel employing a 40-Hz high-pass filter and a 10,000-Hz low-pass filter. The geometric mean, or center frequency, is:

$$f_m = \sqrt{f_1 \times f_2}$$
$$= \sqrt{40 \times 10{,}000}$$
$$= 632.4 \text{ Hz}$$

where,

f_1 is the cutoff frequency of the high-pass filter,
f_2 is the cutoff frequency of the low-pass filter.

The cutoff frequency of a filter is taken at the frequency of 10 dB attenuation. Filter nomenclature is discussed in Question 7.102.

7.70 What is a crystal filter?—A filter composed of quartz crystals. They are used in wave analyzers to obtain an extremely sharp passband characteristic. (See Question 22.65.) Crystal filters are also used in communication-type radio receivers for reducing noise in the reception. Such filters are also referred to as mechanical filters.

7.71 What is an electronic filter?—An amplifier especially designed to have a selective transmission characteristic. Such filters make use of a negative-feedback amplifier with a frequency-selective network such as the parallel T which may be controlled from the panel of the instrument.

Electronic filters are also used for removing unwanted noises in sound

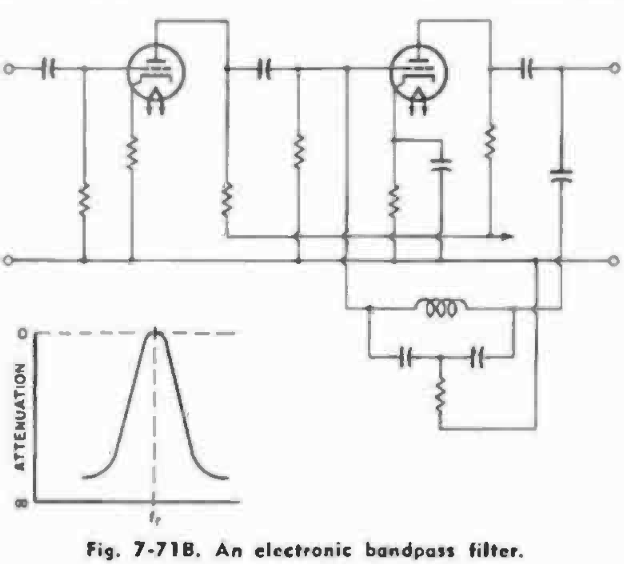
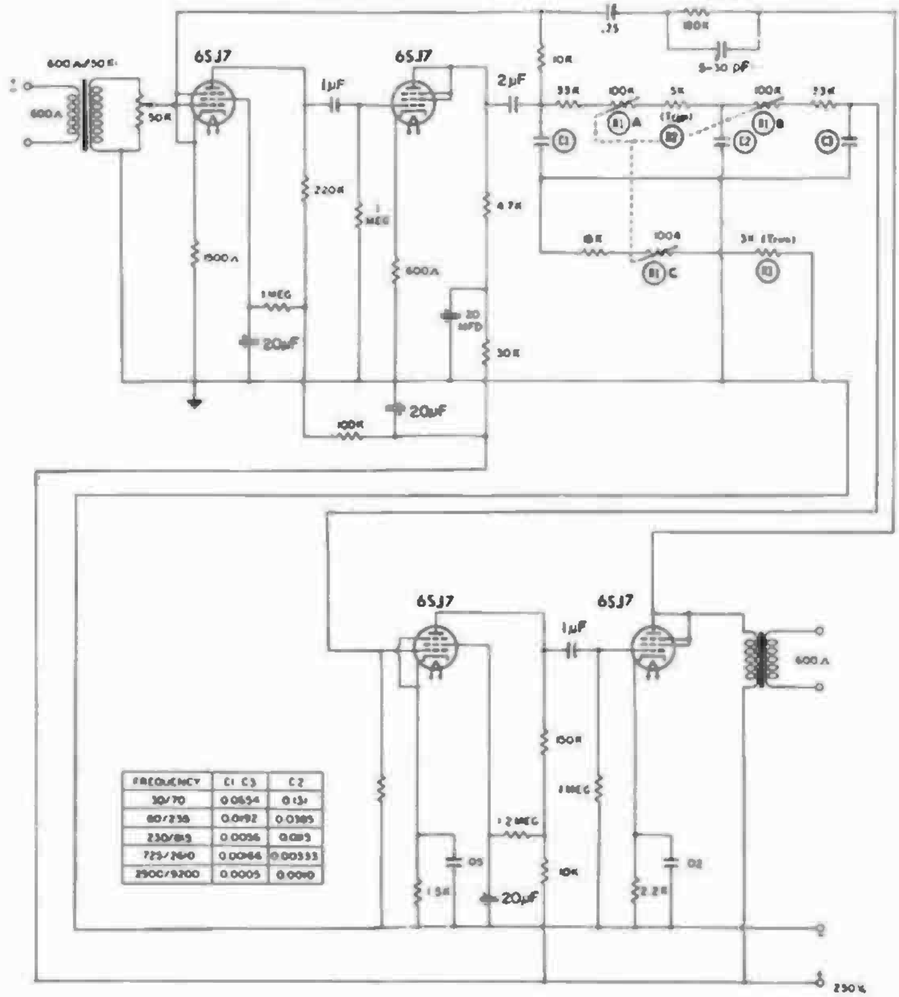


Fig. 7-71B. An electronic bandpass filter.

tracks during the rerecording of motion pictures. Typical low-pass and band-pass filters of the electronic type are shown in Figs. 7-71A and B.

7.72 Show a schematic diagram for a variable dip-filter for rerecording use.—Variable dip filters are used in rerecording to remove unwanted frequencies such as radar, beacon signals, ac hum, arc-light whistles, and many other sounds picked up acoustically or electrically. Fig. 7-72A shows the schematic



FREQUENCY	C1 C3	C2
30/70	0.0654	0.131
80/235	0.0192	0.0385
230/815	0.0056	0.0113
725/2600	0.00166	0.00333
2500/9200	0.0005	0.0010

Fig. 7-72A. Schematic diagram for a parallel-T dip filter.

(a) *Without feedback loop.* (b) *With feedback loop.*

Fig. 7-72B. Frequency response of a parallel-T network.

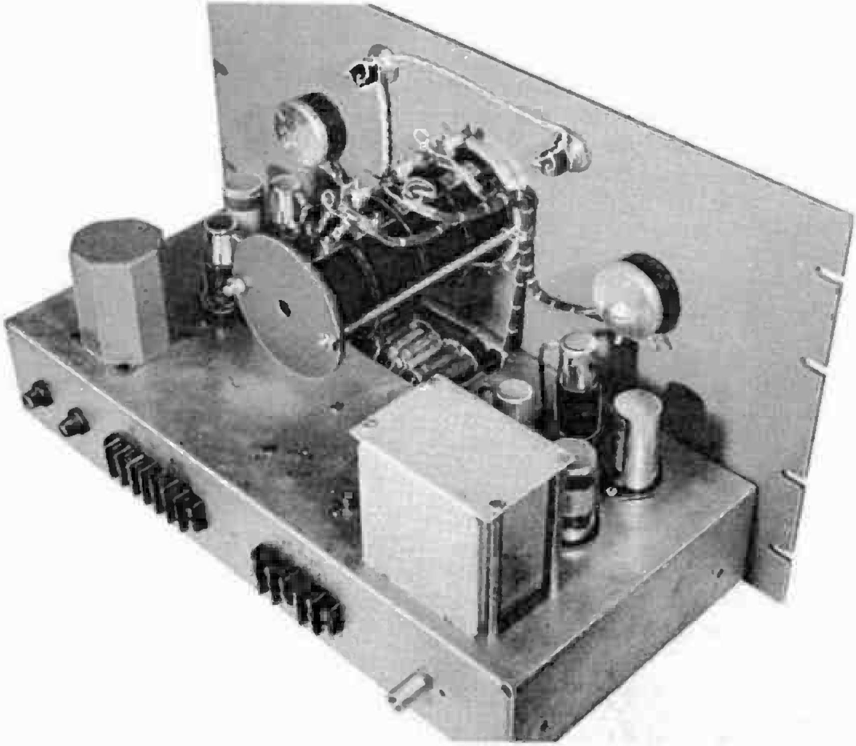


Fig. 7-72C. Rear view of variable parallel-T dip filter.

diagram for a dip filter which has a frequency range of 30 to 9000 Hz, with a band rejection, at the peak frequency, of 50 dB or better. Basically the device is a variable parallel-T network, consisting of capacitors C1, C2, and C3, and three wirewound potentiometers R1A, R1B, and R1C, ganged together. Two additional potentiometers R2 and R3 are used as trimmers to obtain a sharp null point. By the use of a large amount of negative feedback (26 dB) around the parallel-T network, the rejection characteristic is considerably sharpened, as

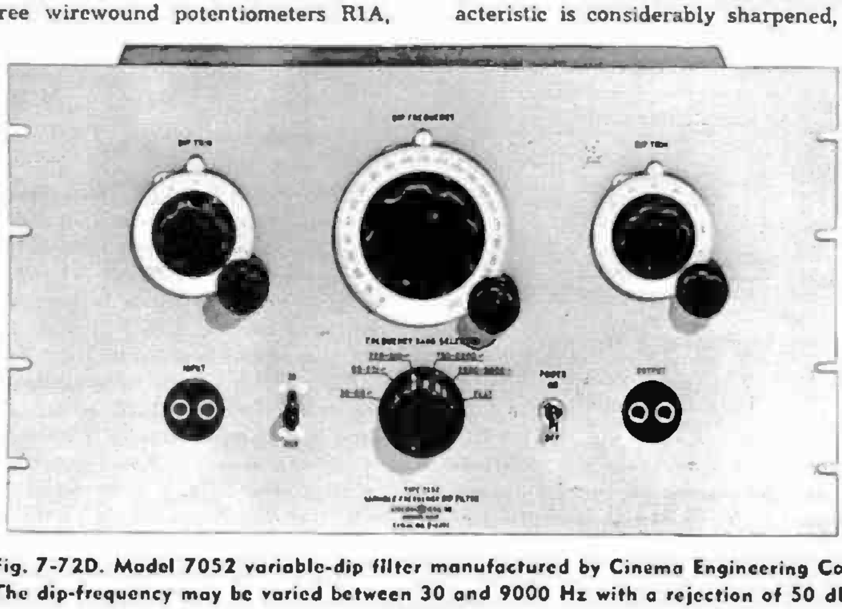
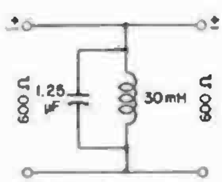


Fig. 7-72D. Model 7052 variable-dip filter manufactured by Cinema Engineering Co. The dip-frequency may be varied between 30 and 9000 Hz with a rejection of 50 dB or better.

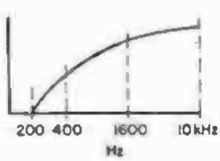
shown in Fig. 7-72B. A rear view of such an instrument built by the author is shown in Fig. 7-72C. For convenience of operation a switch is provided to cut the instrument in and out of the circuit. Therefore, the gain is generally adjusted for a no-gain condition.

After the frequency of an unwanted frequency has been determined, the filter is switched in and adjusted for a *null* at the unwanted frequency, and a test recording made and played back. If the original sound track contained several harmonics besides the fundamental frequency, the rerecorded sound track is played back and the remaining frequencies are dipped out. This procedure may be followed to a third generation provided the signal-to-noise ratio is not too seriously affected. After the final transfer is made it can then be re-equalized. If carefully done, the dip filter may be used in dialogue without it affecting the final result, because of the narrow bandwidth. A dip filter having similar characteristics to that discussed is shown in Fig. 7-72D.

7.73 Show the configuration for a 500-ohm telephone sound-effects filter. —The configuration and component values for such a filter are shown at (a) in Fig. 7-73 and the frequency characteristic is shown at (b).



(a) *Configuration.*



(b) *Frequency response.*

Fig. 7-73. A telephone sound-effects filter for a 500-ohm circuit.

7.74 What are the circuit constants for the more commonly used 500-ohm, constant-k, low- and high-pass filters?— Using the constant-k configurations of Figs. 7-39 and 7-45, filters may be constructed using the following values for a circuit impedance of 500 ohms. If an impedance other than 500 ohms is required, the filter may be converted to any other impedance by the use of the equations given in Question 7-62. For frequencies not appearing in the tabulation, the values may be converted to

Freq. Hz	*L_1 Henry*	*C_1 µF*
30	5.31	21.20
100	1.59	6.37
250	0.637	2.55
400	0.398	1.59
500	0.318	1.27
600	0.265	1.06
800	0.199	0.796
1000	0.159	0.637
3000	0.053	0.212
10,000	0.016	0.063

Low-Pass (See Fig. 7-39)

other frequencies by the equations in Question 7-63 or by multiplying or dividing the component values by the ratio of the new frequency to the known frequency.

Freq. Hz	*L_1 Henry*	*C_1 µF*
30	1.33	5.31
100	0.398	1.59
250	0.159	0.637
400	0.0995	0.398
500	0.0796	0.318
600	0.0663	0.265
800	0.0497	0.199
1000	0.0398	0.159
3000	0.0133	0.0531
10,000	0.00398	0.0159

High-Pass (See Fig. 7-45)

7.75 What are the circuit values for the more commonly used series m-derived low-pass filters with a characteristic impedance of 500 ohms?—The values given in Fig. 7-75 are to be used with the configurations shown in Fig. 7-52. The value of m is equal to 0.6.

7.76 What are the circuit values for the more commonly used series m-derived high-pass filters with a characteristic impedance of 500 ohms?—The values given in Fig. 7-76 are to be used with the configurations shown in Fig. 7-54. The value of m is equal to 0.6.

7.77 What are the circuit values for the more commonly used shunt m-derived low-pass filters?—The values given

Freq. Hz	L_1 Henry	L_2 Henry	C_1 μF
30	3.186	1.412	12.720
100	0.954	0.4229	3.822
250	0.3822	0.1694	1.530
400	0.2388	0.1058	0.954
500	0.1908	0.0846	0.762
600	0.1590	0.0705	0.636
800	0.1194	0.0529	0.477
1000	0.0954	0.0423	0.382
3000	0.0318	0.0141	0.127
10,000	0.0096	0.00425	0.038

Fig. 7-75. Values for Question 7.75.

in Fig. 7-77 are to be used with the configurations shown in Fig. 7-55. The value of m is equal to 0.6.

Freq. Hz	L_2 Henry	C_1 μF	C_2 μF
30	2.216	8.85	19.912
100	0.663	2.65	5.962
250	0.265	1.06	2.388
400	0.165	0.663	1.492
500	0.132	0.530	1.192
600	0.110	0.441	0.993
800	0.082	0.331	0.746
1000	0.063	0.265	0.596
3000	0.022	0.088	0.199
10,000	0.0066	0.0265	0.059

Fig. 7-76. Values for Question 7.76.

7.78 ***What are the circuit values for the more commonly used shunt m-derived high-pass filters?***—The values given in Fig. 7-78 are to be used with the configurations shown in Fig. 7-56. The value of m is equal to 0.6.

7.79 ***What are the configuration and circuit constants for a variable high-pass filter suitable for rerecording purposes?***—The configuration and circuit constants for a 600-ohm unbalanced variable high-pass filter designed for the express purpose of removing the low frequency end of dialogue are shown in Fig. 7-79A. The frequency range covered is from 80 to 150 Hz. For dialogue recording, it is generally set for 80 Hz cutoff, which is standard in the motion-picture industry. At times it may be desirable to use a higher cutoff frequency to remove low-frequency noise falling between 150 and 80 Hz. However, male dialogue becomes thin if the cutoff is raised above 100 Hz.

Freq. Hz	L_1 Henry	C_1 μF	C_2 μF
30	3.186	5.639	12.720
100	0.954	1.694	3.822
250	0.382	0.678	1.530
400	0.238	0.422	0.954
500	0.191	0.337	0.762
600	0.159	0.281	0.636
800	0.119	0.211	0.477
1000	0.095	0.169	0.382
3000	0.032	0.056	0.127
10,000	0.0095	0.017	0.038

Fig. 7-77. Values for Question 7.77.

Freq. Hz	L_1 Henry	L_2 Henry	C_1 μF
30	4.897	2.216	8.850
100	1.492	0.663	2.650
250	0.596	0.265	1.060
250	0.596	0.265	1.060
400	0.373	0.1658	0.663
500	0.298	0.1326	0.530
600	0.248	0.1105	0.441
800	0.186	0.0825	0.331
1000	0.1498	0.0633	0.265
3000	0.0498	0.0223	0.0885
10,000	0.01498	0.0066	0.0265

Fig. 7-78. Values for Question 7.78.

The coils should be of toroidal design with the maximum Q at the cutoff frequency. Because this filter operates in a frequency band near frequencies which can produce hum and noise, the wiring should be well shielded and the whole unit enclosed in a metal case. The use of toroidal coils will reduce hum pickup to a minimum. The configuration should be grounded to prevent leakage. The frequency characteristic of this filter is shown in Fig. 7-79B.

7.80 ***What is a crossover network?***—In modern high-fidelity sound reproducing systems, the loudspeaker is designed to cover a wide range of frequencies, generally 30 to 15,000 Hz or more. To obtain satisfactory operation over this wide range requires the use of two or more loudspeakers, each operating in a given frequency range, as it is not practical to attempt to obtain this wide range with a single loudspeaker.

When two or more loudspeakers are used in combination, the frequency spectrum is divided into bands and controlled by a crossover or frequency-dividing network composed of two or more filters. As a rule, they are a high-

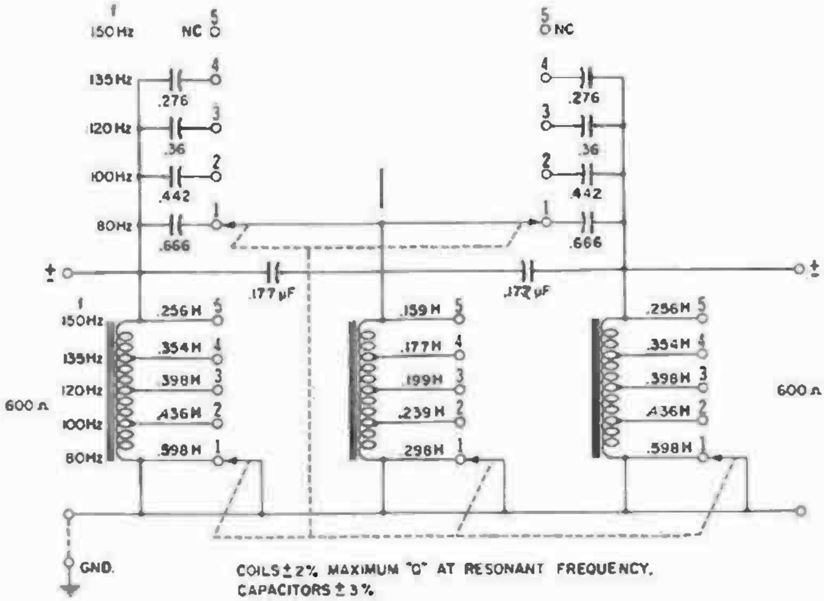


Fig. 7-79A. A variable high-pass filter for rerecording.

pass and a low-pass filter. The low-pass filter limits the high-frequency response of the low-frequency speakers and the high-pass filter limits the low-frequency response of the high-frequency speakers.

A three-way network consists of a low-pass, bandpass, and high-pass filter section. The low- and high-frequency sections function as for a two-way system. The bandpass section limits the frequency response for the midrange speaker unit, permitting it to extend only into a portion of the low- and high-frequency ranges. The result is a smooth transition over the complete range of the three speaker units.

The sole purpose of a crossover network is to limit the operating range of a given speaker unit. The network cannot be used to correct for deficiencies in the amplifier system, speaker units, or enclosure. One of the most important conditions imposed on the network is that it must not induce an appreciable amount of loss (insertion loss) between the amplifier and the speaker units; this could result in a considerable

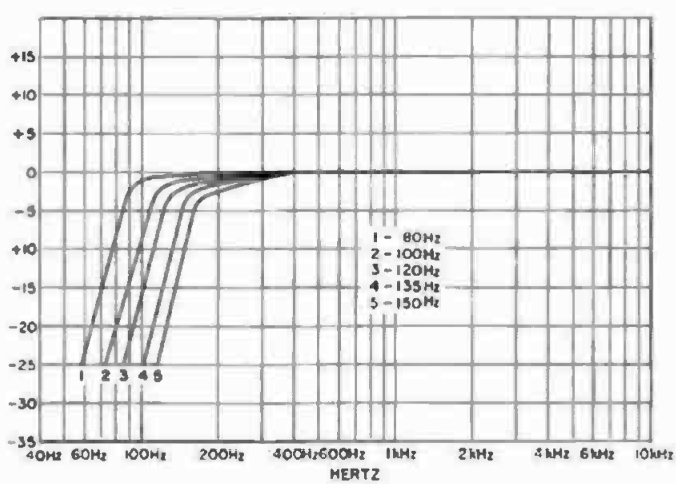


Fig. 7-79B. Frequency characteristics of the variable high-pass filter shown in Fig. 7-79A.

Fig. 7-80. A 400-Hz speaker-crossover network.

loss of power particularly with amplifiers of 50-watts power or more. (See Question 7.84.)

7.81 What is the frequency response of a typical crossover network?—A typical frequency response for a crossover network comprised of a low-pass and a high-pass filter is shown in Fig. 7-81. The curves are ideal curves for three different rates of cutoff. The point where the curves cross over is called the crossover frequency.

The characteristics above and below the crossover frequency drop off approximately 6, 12, or 18 dB per octave, depending on the design of the network. In practice, because of the dissipative losses in the circuit elements, the first octave for the 12 dB per octave filter falls off about 10 dB and 12 dB thereafter. The 6 and 18 dB per octave filters fall off correspondingly.

7.82 What are the advantages of a parallel crossover network over a series type?—The selection of the configuration for a crossover network is not critical; either the series or parallel network may be used. However, the parallel configuration does offer slightly better characteristics in the transmission and attenuation bands. Series configurations are used only with two-way speaker systems. Most commercial units employ the parallel configurations because the component values are the same for each filter; thus, manufacturing costs are reduced. Theoretically the constant-k, constant-resistance network reflects a constant resistance back to the amplifier output. Since a speaker

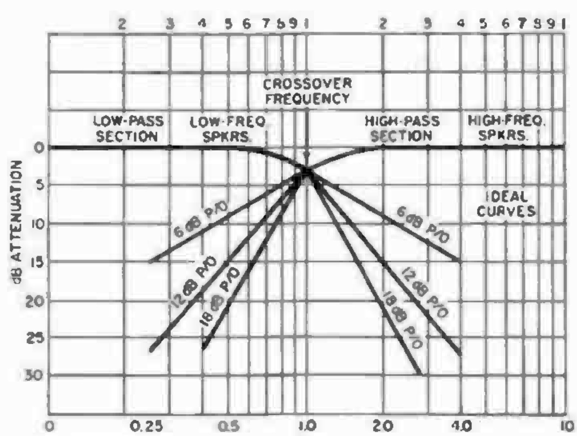


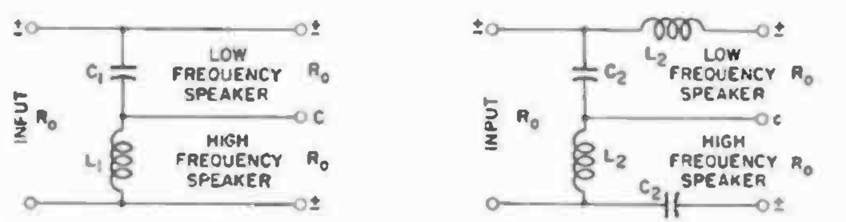
Fig. 7-81. Typical frequency characteristic for ideal 6, 12, and 18 dB per octave crossover networks.

unit does not reflect a constant resistive load, the amplifier does not see a constant resistance. This disadvantage is somewhat compensated for by the use of negative feedback in the amplifier circuitry. The only time an amplifier sees a purely resistive load is when the network is terminated, on the speaker side, in the *pure* resistance. In a two-section constant-resistance network, the frequency response is complementary, and the sum of the power delivered to the output section is constant for a constant input voltage at the input of the network. When the output sections are terminated in a resistive load, the input impedance presented to the amplifier is constant throughout the entire frequency range. The frequency response of the filter sections in the passband is uniform, with constant phase-difference at the output of the high and low-frequency sections. (See Fig. 7-82.)

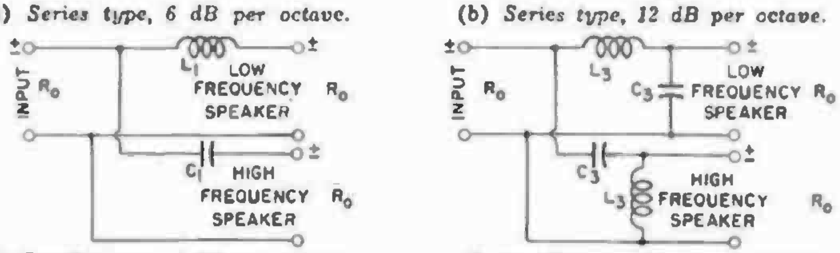
7.83 ***Are equalizers ever included in a crossover network?***—Yes, some manufacturers include a simple high-frequency equalizer in the network to increase the response of the midrange and high-frequency units. This equalizer, though included with the network, is an *added* feature and is not a part of the network proper. The equalizer generally consists of a resistor in series with the midrange or high-frequency speaker unit, and a group of capacitors which may be cut in or out of the circuit by a switch. The capacitors are connected in parallel with the resistor. The switch permits the user to select a capacitor that will increase the high-frequency response to suit his local acoustic conditions.

7.84 ***What causes insertion loss and what is the average loss for a crossover network?***—The insertion loss of any network is caused by the dc resistance of the coils, and the shunt and series reactance of the circuit elements. As a rule, the insertion loss of a crossover network due to circuit elements is approximately 0.5 dB and is exclusive of the crossover-frequency loss which is 3 dB.

Insertion losses up to 1 dB are not too important if the amplifier power output is not greater than 20 watts. However, if the amplifier system is greater than 20 watts, the insertion loss becomes important, especially in commercial installations where 100 watts or more may be used. As an example, for an insertion loss of 0.5 dB at 50 watts, the power dissipated by the network is approximately 6 watts, and for a 100-watt system the loss is more than 10 watts (measured by using a steady



(a) *Series type, 6 dB per octave.* (b) *Series type, 12 dB per octave.*



(c) *Parallel type, 6 dB per octave.* (d) *Parallel type, 12 dB per octave.*

$C_1 = \frac{1}{\omega_c R_0}$ FARAD $\quad L_1 = \frac{R_0}{\omega_c}$ HENRY $\quad \omega_c = 2\pi f_c$

$C_2 = \sqrt{2}\, C_1$ FARAD $\quad L_2 = \frac{L_1}{\sqrt{2}}$ HENRY $\quad R_0$ = SPEAKER IMPEDANCE

$C_3 = \frac{C_1}{\sqrt{2}}$ FARAD $\quad L_3 = \sqrt{2}\, L_1$ HENRY $\quad f_c$ = CROSSOVER FREQ.

Fig. 7-82. Constant-k constant-resistance, crossover networks and design equations.

input voltage in the passband). To obtain a low insertion loss, the coils must have a low dc resistance, and the capacitors must have a *low power factor*. Special nonpolarized electrolytic and paper capacitors are available for crossover network construction. Conventional nonpolarized electrolytic capacitors should not be used because of their power factor and leakage. The characteristics for capacitors manufactured especially for crossover-network use are given in Figs. 7-84A and B with standard ratings given in Fig. 7-84C. Capacity tolerances are plus or minus 20 percent of the rated value. The peak signal voltage must *never* exceed the dc working voltage ratings.

7.85 What factors dictate the crossover frequency of a network?—The choice of crossover frequency is dictated by the frequency response of the speaker units. Typical crossover frequencies for low-frequency sections are 400, 450, and 600 Hz, with 3000 and 5000 Hz for the high-frequency sections. Ideal crossover characteristics are shown in Fig. 7-85 for both two-way and three-way systems. The crossover frequency must become effective before the response of a given speaker unit falls off, and the movement of the diaphragm becomes nonlinear. Low-frequency speakers designed for multiple speaker systems seldom have much response above 1500 to 2000 Hz. The frequency response of the midfrequency speakers must be restricted to those frequencies where the wavelengths are such that the excursion of the dia-

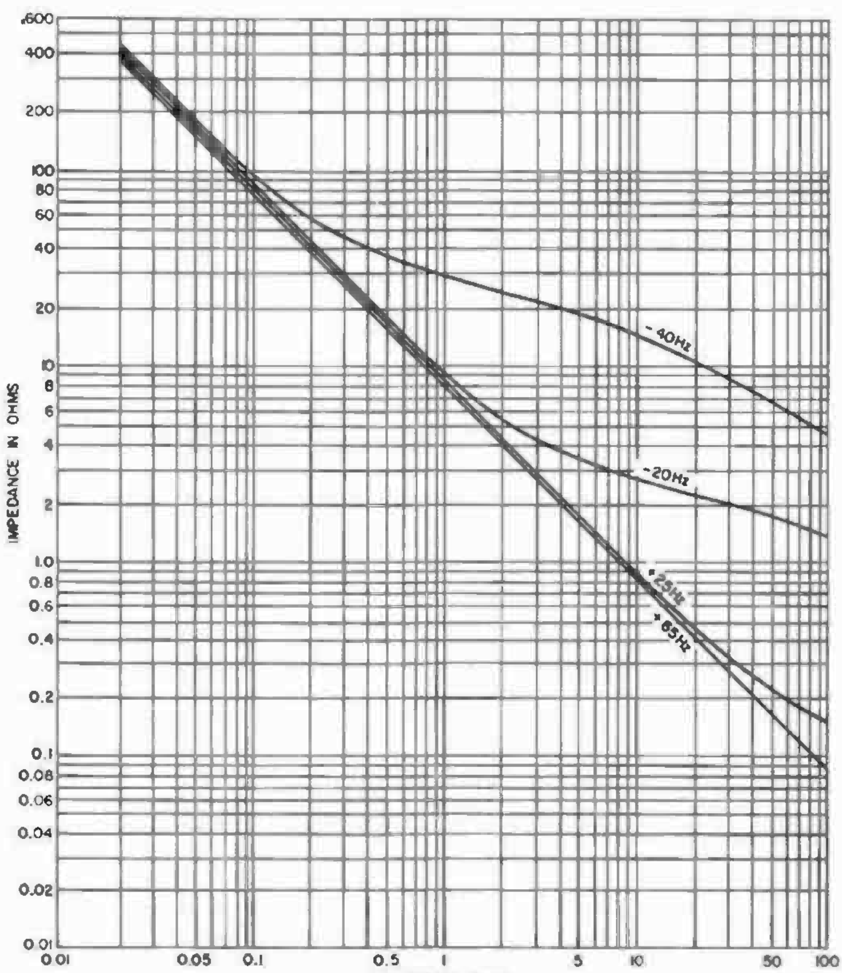


Fig. 7-84A. Impedance versus frequency characteristics of capacitors made for crossover networks. (Courtesy, Sprague Electric Co.)

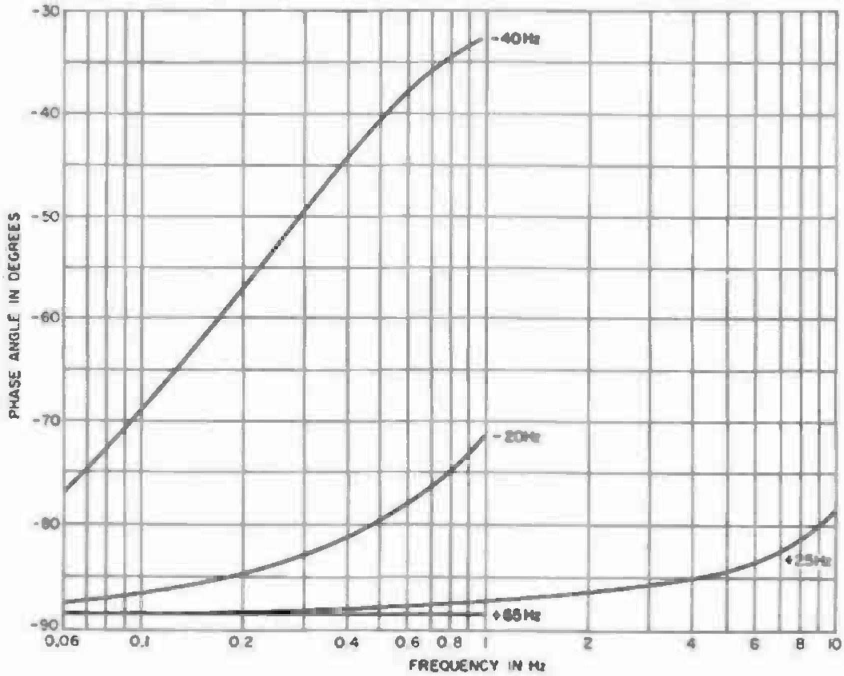


Fig. 7-84B. Phase angle as a function of frequency of crossover-network capacitors. (Courtesy, Sprague Electric Co.)

25 WVDC, 30 VOLTS SURGE				
1	31DC1019	14	7	DC
2	31DC1020	22	10	EC
3	31DC1021	30	12	EE
4	31DC1022	40	15	FE
5	31DC1023	45	17	FF
6	31DC1024	60	17	GG
8	31DC1025	70	20	HF
10	31DC1026	85	25	HG
15	31DC1027	120	30	HH
20	31DC1028	150	35	JH
25	31DC1029	170	37	JH
30	31DC1030	180	40	JH
40	31DC1031	240	47	JK
50	31DC1032	290	50	KK
65	31DC1033	340	60	KL
75	31DC1034	380	65	LK
90	31DC1035	430	70	LL

Fig. 7-84C. Standard ratings of capacitors designed for loudspeaker crossover networks. (Courtesy, Sprague Electric Co.)

phragm will not exceed the *diaphragm displacement* recommended by the manufacturer.

Networks employing a 6 dB per octave cutoff do not as a rule provide a rapid enough cutoff. The network rolloff should be at a frequency high or low enough to protect the midrange and high-frequency units from damage. The low-frequency unit must be capable of handling at least *one octave above* the crossover frequency, and the high-frequency unit *one octave below* the crossover frequency, at the full power of the driving amplifier. The use of a 12 dB per octave cutoff rate will greatly assist in overcoming this difficulty, and offers greater protection to the midrange and high-frequency units. For a two-way system, a crossover frequency above 600 Hz should be avoided, as most of the peak power is concentrated below 500 Hz. The use of a 350- to 400-Hz crossover results in equal division of

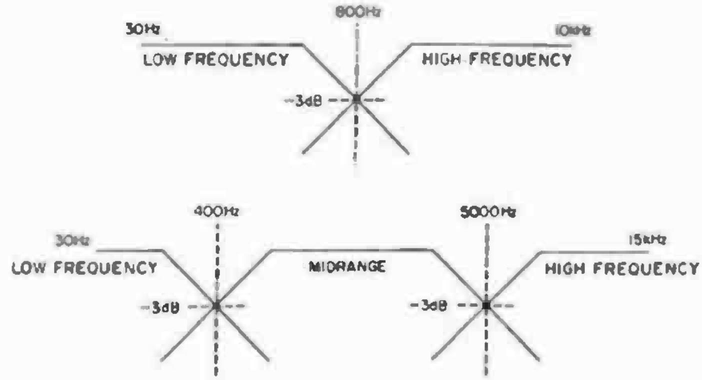


Fig. 7-85. Frequency response for two-way and three-way crossover filter networks.

the total peak power between the low- and high-frequency speaker units. For a three-way system 400 to 500 Hz for the low-frequency crossover and 5000 Hz for the high-frequency end seems a good compromise.

7.86 What is the power loss at the crossover frequency?—In a well-designed network, 3 dB, exclusive of the insertion loss. At the crossover frequency the power applied to the high- and low-frequency speakers is divided equally. Limiting the low-frequency response of the high-frequency speakers prevents damage to the diaphragm.

Above the frequency of the crossover point, the movement of the high-frequency speaker diaphragm is only a few thousandths of an inch, while the low-frequency speaker diaphragm may be moving up to ⅜ of an inch.

Although a 3-dB loss occurs at the crossover frequency, this is not particularly noticeable to the ear and is generally within the overall frequency limits of the speaker system.

7.78 What is a constant-resistance network?—A network that will reflect a constant resistance to the output circuit of the driving amplifier when terminated in a resistive load. Loudspeakers do not reflect a constant impedance; therefore, the amplifier does not see a constant resistance. This disadvantage may be somewhat compensated for by the use of negative feedback in the amplifier.

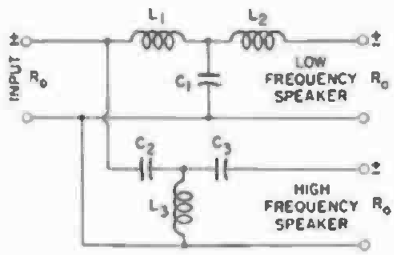
In a constant-resistance type network, the frequency response of the two filter sections is complementary. The sum of the power delivered to the output circuits of the filters is constant for a constant input voltage. When terminated in a pure resistive load, the input impedance of the filter is constant throughout the whole frequency range. The frequency response of the filters in the passband is uniform, with a constant phase difference at the outputs. Constant-resistance filters may be designed for either series or parallel operation.

7.88 Is there any advantage in increasing the cutoff-frequency rate beyond 18 dB per octave?—No, 18 dB is the maximum rate used in practice, as the improvement does not offset the additional power losses. Most crossover networks employ a 12 dB per octave cutoff rate.

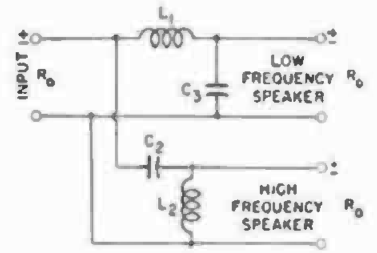
7.89 What is the advantage of an m-derived crossover network compared to a constant-k type?—Constant-k networks are limited to a maximum cutoff rate of 12 dB per octave, while the m-derived network cau provide a 12 dB greater rate of cutoff. An additional advantage of the m-derived filter is that the impedance or attenuation characteristics can be closely controlled by the designer. For most audio filter work, m equals 0.6; the same holds true for crossover networks. Configurations for both 12 and 18 dB per octave, series and parallel types are given in Fig. 7-89.

7.90 What type coil is used in crossover networks?—Air-core coils are used. Because the coils used in crossover networks are of only a few millihenries inductance, they may be wound with number 10 and number 18 gauge wire on an air core. The winding may be on a wooden core and scrambled-wound.

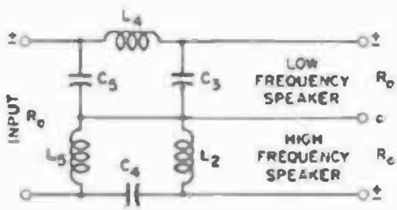
7.91 What type capacitor is used in crossover networks?—Paper dielectric or special nonpolarized electrolytic capacitors, as discussed in Question 7.84.



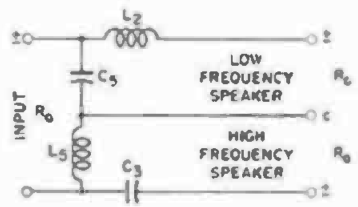
(a) *Parallel type, 18 dB per octave.*



(b) *Parallel type, 12 dB per octave.*



(c) *Series type, 18 dB per octave.*



(d) *Series type, 12 dB per octave.*

$C_1 = \frac{2}{\omega_c R_0}$ FARAD $L_1 = (1+m)\frac{R_0}{\omega_c}$ HENRY

$C_2 = \left(\frac{1}{1+m}\right)\frac{1}{\omega_c R_0}$ FARAD $L_2 = \frac{R_0}{\omega_c}$ HENRY $\omega_c = 2\pi f_c$

$C_3 = \frac{1}{\omega_c R_0}$ FARAD $L_3 = \frac{R_0}{2\omega_c}$ HENRY R_0 = SPEAKER IMPEDANCE

$C_4 = \frac{1}{2\omega_c R_0}$ FARAD $L_4 = \frac{2R_0}{\omega_c}$ HENRY f_c = CROSSOVER FREQ.

$C_5 = (1+m)\frac{1}{\omega_c R_0}$ FARAD $L_5 = \left(\frac{1}{1+m}\right)\frac{R_0}{\omega_c}$ HENRY $m = 0.6$

Fig. 7-89. Conventional m-derived crossover networks and design equations.

7.92 ***Are iron-core coils used in crossover networks?***—No. Iron-core inductances induce distortion because of saturation. Also, their cost is considerably higher than air-core types.

7.93 ***Can a crossover network be designed to operate with speakers of unequal impedance?***—Yes, by the use of the circuit in Fig. 7-93 and by using the design data given in the following equations:

$$L_1 = \frac{\sqrt{2}\,R_L}{w_c}$$

$$L_2 = \frac{\sqrt{2}\,R_H}{w_c}$$

$$C_1 = \frac{1}{\sqrt{2}\,R_L w_c}$$

$$C_2 = \frac{1}{\sqrt{2R}\,R_H w_c}$$

where,

R_L is the impedance of the low-frequency speaker,

R_H is the impedance of the high-frequency speaker,

w_c is 2π times the frequency of cutoff.

The circuit elements are connected in the output of a normal output transformer having an 8-ohm and a 16-ohm winding. If the output transformer in a particular amplifier has only one winding, a matching transformer may be used to supply the desired impedances.

7.94 ***Show the different methods of connecting loudspeakers to the output of crossover networks.***—In Fig. 7-94 part (a), an 8-ohm network is shown connected to two 16-ohm low-frequency speakers in parallel. The two 16-ohm speakers *reflect* an 8-ohm load impedance to the network. The high-frequency speaker unit has an impedance

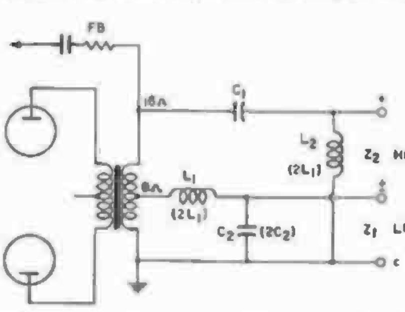
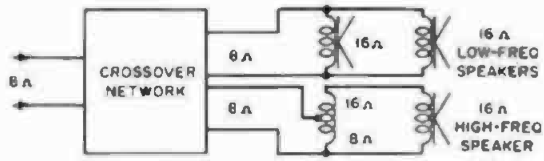
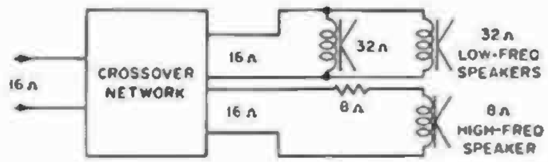


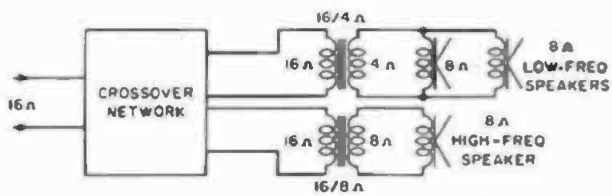
Fig. 7-93. Crossover network for loudspeakers of unequal impedance.



(a) *Two 16-ohm low-frequency speakers matched to 8-ohm network and one 16-ohm high-frequency speaker matched to 8-ohm network.*



(b) *Two 32-ohm low-frequency speakers matched to 16-ohm network and one 8-ohm high-frequency speaker matched to 16-ohm network.*



(c) *Two 8-ohm low-frequency speakers matched to 16-ohm network and one 8-ohm high-frequency speaker matched to 16-ohm network.*

Fig. 7-94. Various methods that are used for matching the speaker impedance to the network impedance.

of 16 ohms and is connected to the output of the network by means of an autotransformer to match the 8-ohm output to the 16-ohm speaker.

Two 32-ohm low-frequency speakers connected in parallel to the output of a 16-ohm network, appear in Fig. 7-49 part (b). The high-frequency unit has an impedance of 8 ohms, therefore an 8-ohm resistor is connected in series with the speaker unit to increase the load impedance to 16 ohms. It should be pointed out that the use of a resistor is *not* recommended unless the high-frequency unit has high sensitivity, since half of the power is dissipated in the resistor. Because of the loss incurred with a resistor and the higher power required to drive low-frequency speaker units, *resistors should not be used with low-frequency units.*

A third method of impedance matching is shown in Fig. 7-94 part (c). Here, two 8-ohm low-frequency units are matched to a 16-ohm network by means of a transformer having an impedance ratio of 4:16. The 8-ohm high-frequency unit is matched to the network with an 8:16-ohm transformer. Whenever possible, the network should be designed for a given impedance, and the speaker units selected for that impedance. Transformers are quite expensive and induce a *power loss*, particularly for good frequency response below 50 Hz.

7.95 ***Why is an autotransformer recommended for matching the voice-coil impedance to a crossover network?*** —Because it has less insertion loss and is less expensive.

7.96 ***What are the circuit-constants for crossover networks?***—The circuit constants for constant-k and m-derived crossover networks, both series and parallel configuration, are given in Figs. 7-96A and B. The constants for the desired network and frequency are selected from the tables, and then converted *from an impedance of 10 ohms to the required impedance.* No further calculations are required. (See Question 7.62.)

7.97 ***How are the network constants shown in Fig. 7-96A and B converted to other impedances and frequencies?***—The impedance may be changed by use of the equations given in Question 7.62,

$f_c =$	100	150	200	250	300	350	400	450	500	550	600	650	700	750	800	850	900	950	1000	3000	5000	6000
C_1	159.15	106.15	79.62	63.69	53.08	45.49	39.81	35.38	31.85	28.95	26.54	24.49	22.74	21.23	19.90	18.75	17.69	16.76	15.91	5.30	3.18	2.65
C_2	225.04	150.10	112.58	90.06	75.05	64.32	56.29	50.03	45.03	40.93	37.53	34.63	32.15	30.02	28.14	26.51	25.01	23.69	22.50	7.50	4.50	3.75
C_3	112.52	75.05	56.29	45.03	37.52	32.16	28.14	25.01	22.51	20.46	18.76	17.31	16.07	15.01	14.07	13.25	12.50	11.84	11.25	3.75	2.25	1.87
L_1	15.91	10.61	7.96	6.37	5.31	4.55	3.98	3.54	3.18	2.89	2.65	2.45	2.27	2.12	1.99	1.87	1.77	1.68	1.59	0.531	0.318	0.265
L_2	11.25	7.50	5.63	4.50	3.75	3.22	2.81	2.50	2.25	2.04	1.87	1.73	1.60	1.50	1.41	1.32	1.25	1.19	1.12	0.375	0.225	0.187
L_3	22.50	15.00	11.26	9.00	7.50	6.44	5.62	5.00	24.50	4.08	3.74	3.46	3.20	3.00	2.82	2.64	2.50	2.38	2.24	0.750	0.450	0.374

Inductance in millihenries. Capacitance in microfarads. $R_o = 10$ ohms.

Fig. 7-96A. Component values for constant-k loudspeaker crossover networks shown in Fig. 7-82.

$f_c =$	100	150	200	250	300	350	400	450	500	550	600	650	700	750	800	850	900	950	1000	3000	5000	6000
C_1	318.5	212.30	159.24	127.38	106.16	90.98	79.62	70.76	63.70	57.90	53.08	48.98	45.48	42.46	39.80	37.50	36.38	33.52	31.85	10.61	6.37	5.30
C_2	99.47	66.34	49.76	39.81	33.17	28.43	24.88	22.11	19.91	18.09	16.59	15.31	14.21	13.27	12.44	11.72	11.06	10.47	9.95	3.32	1.99	1.65
C_3	159.15	106.15	79.62	63.69	53.08	45.49	39.81	35.38	31.85	28.95	26.54	24.49	22.74	21.23	19.90	18.75	17.69	16.76	15.91	5.30	3.18	2.65
C_4	79.57	53.07	39.81	31.84	26.54	22.74	19.90	17.69	15.92	14.47	13.27	12.24	11.37	10.61	9.95	9.37	8.84	8.38	7.96	2.65	1.59	1.32
C_5	254.64	169.84	127.39	101.90	84.93	72.78	63.69	56.61	50.96	46.32	42.46	39.18	36.38	33.97	31.84	30.00	28.30	26.82	25.46	8.49	5.09	4.24
L_1	25.46	16.98	12.74	10.19	8.49	7.28	6.37	5.66	5.10	4.63	4.25	3.92	3.64	3.40	3.18	3.00	2.83	2.68	2.55	0.849	0.510	0.425
L_2	15.91	10.61	7.69	6.37	5.31	4.55	3.98	3.54	3.18	2.89	2.65	2.45	2.27	2.12	1.99	1.87	1.77	1.68	1.59	0.531	0.318	0.265
L_3	7.96	5.31	3.98	3.18	2.65	2.27	1.99	1.77	1.59	1.45	1.33	1.22	1.14	1.06	0.995	0.937	0.884	0.838	0.796	0.265	0.159	0.133
L_4	31.85	21.23	15.92	12.74	10.62	9.09	7.96	7.08	6.37	5.79	5.31	4.90	4.55	4.25	3.98	3.75	3.54	3.35	3.18	1.06	0.637	0.531
L_5	9.95	6.63	4.98	3.98	3.32	2.84	2.49	2.21	1.99	1.81	1.66	1.53	1.42	1.33	1.24	1.17	1.11	1.05	0.995	0.332	0.199	0.166

Inductance in millihenries. Capacitance in microfarads. $m = 0.6$.
$R_o = 10$ ohms.

Fig. 7-96B. Component values for m-derived loudspeaker crossover networks shown in Fig. 7-89.

while the frequency may be changed by use of those given in Question 7.63.

7.98 ***Show the configurations for three- and five-way crossover networks.*** —The three-way crossover network differs from the two-way network, only in the addition of the midrange (bandpass filter) circuit elements. The configuration for the design of a three-way constant-k network, using 6 dB per octave rolloff is given in Fig. 7-98A, and is basically the same as Fig. 7-82C, except for the addition of C_{1A} and L_{1A}. The circuit element values are calculated by using the equations below the diagram. Analyzing the frequency-response curve above the filter sections reveals that

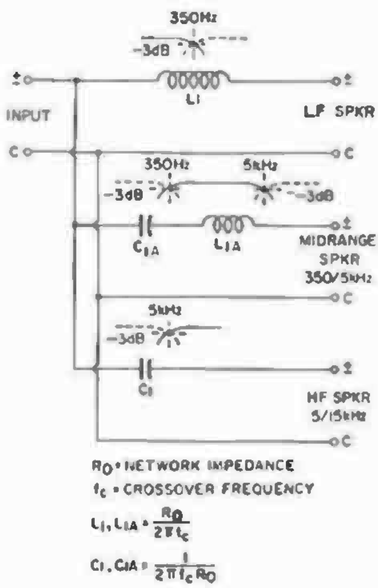


Fig. 7-98A. A constant-k, three-way 6 dB per octave crossover network.

two crossover frequencies are established at a point 3 dB down from the flat portion of the characteristic. A three-way constant-k, 16-ohm, 12 dB per octave network is shown in Fig. 7-98B, with its frequency response shown in Fig. 7-98C. Components C_{3A} and L_{3A} constitute a *low-pass* filter with a cutoff frequency of 450 Hz.

The midrange section is a *bandpass* filter, consisting of a simple 450 Hz high-pass section, and a 5000-Hz low-pass filter. Circuit element C_{3A} and L_{3A} form a high-pass filter, and L_{3B} and C_{3B} form a low-pass filter. These two sections together form a bandpass section. Comparing the low-pass section and the high-pass section of the bandpass filter, it will be observed that both curves are down 3 dB from the flat portion of their response curve, and the crossover frequency is 450 Hz. The output portion of the midrange filter L_{3C} and C_{3C} at the lower portion of the configuration form a 5000-Hz high-pass filter. Comparing the frequency response of the two sections, a crossover frequency of 5000 Hz has been established.

The design procedure is to *first* calculate the value of the low-pass section L_3 and C_3. Circuit elements L_{3A} and C_{3A} must also cross over at 450 Hz, but in an *inverse* manner; therefore, their values are equal to L_3 and C_3. The output portion of the midrange forms a low-pass filter with a cutoff of 5000 Hz; these values are calculated for that frequency. Circuit elements C_{3C} and L_{3C} form a high-pass filter (5000 Hz) with the same values as C_{3B} and L_{3B}, but inversely connected. When the frequency

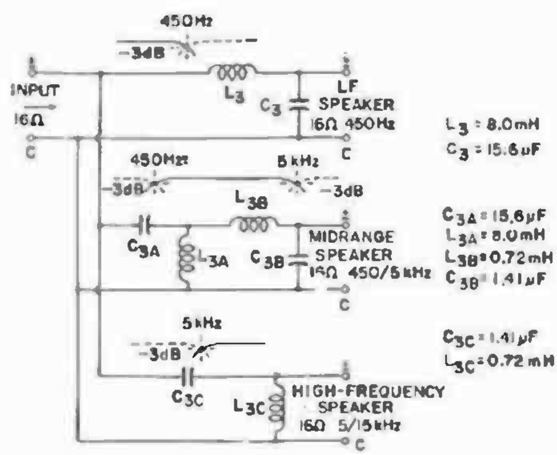


Fig. 7-98B. Three-way, 12 dB per octave, 16-ohm crossover network.

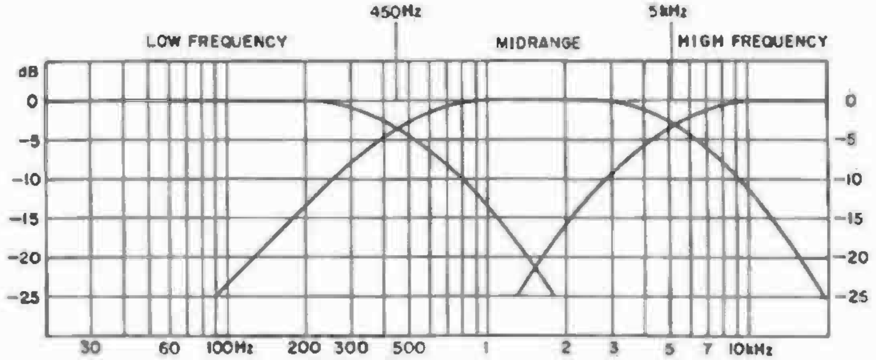


Fig. 7-98C. Frequency response for network of Fig. 7-98B.

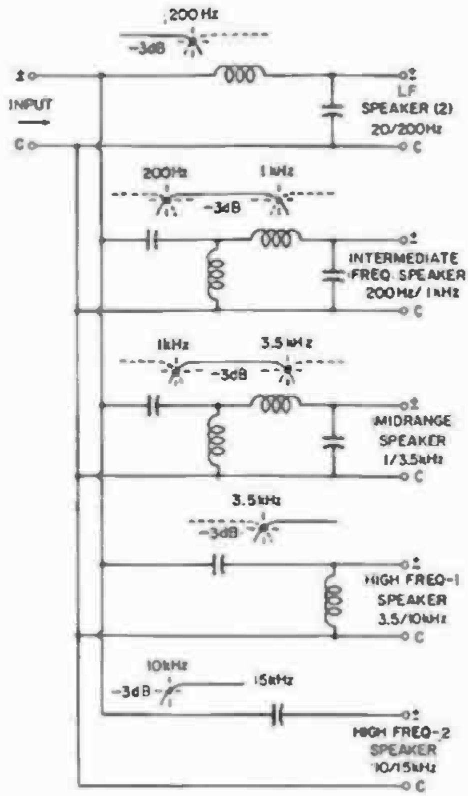


Fig. 7-98D. Five-way, constant-k crossover network.

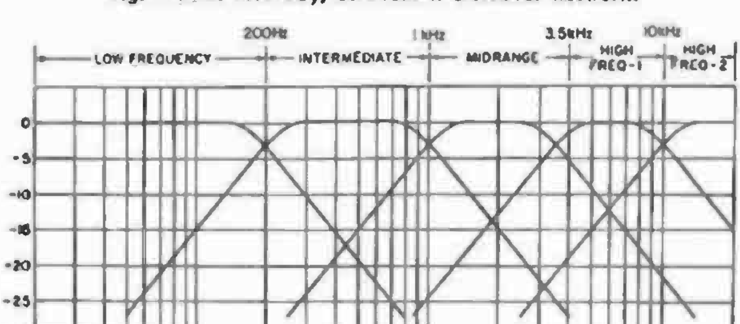


Fig. 7-98E. Frequency response for network of Fig. 7-98D.

response of the two are compared, a crossover of 5000 Hz is created. The circuit element values of Fig. 7-98B may be converted to an m-derived network by the use of the equations:

$$L_a, L_{an} = (1+m)\left(\frac{R_o}{\omega_c}\right) \text{ henry}$$

$$L_{aA}, L_{ac} = \left(\frac{R_o}{\omega_c}\right) \text{ henry}$$

$$C_{3A}, C_{3C} = \left(\frac{1}{1+m}\right)\left(\frac{1}{\omega_c R_o}\right) \text{ farad}$$

$$C_{3B} = \left(\frac{1}{\omega_c R_o}\right) \text{ farad}$$

where,
- m equals 0.6,
- ω equals 2π fc;
- R_o is the circuit impedance.

Actually there is little difference in the design of four- or five-way networks, except for the *crossover* frequencies. A typical five-way network is given in Fig. 7-98D, with its frequency response shown in Fig. 7-98E. The circuit element values are calculated as for the three-way network (Fig. 7-98B) with additional high-pass sections. It will be observed the fifth section (10,000 Hz) consists of a single capacitor connected in series with the second high speaker unit. As this latter unit is confined to frequencies of slightly less than 10,000 Hz and above, a more elaborate network is not required. The frequency response for each section of the network is plotted above each configuration. The values of the series capacitor may be calculated:

$$C = \frac{79{,}600}{f \times R_o}$$

where,
- 79,600 is a constant,
- f is the cutoff frequency,
- R_o is the speaker impedance.

7.99 Can a crossover network be designed for 250- or 600-ohms impedance?—Yes, many theater installations use 250- or 600-ohm crossover networks because of the long transmission line between the amplifier system and the stage speakers. Autotransformers are used at the speaker end to convert the high impedance down to the voice-coil impedance.

By using a high-impedance transmission line, the voltage is high and the current low; thus, the power losses that would be encountered using a low impedance line are avoided. The configuration for a 250-ohm crossover network used in motion picture theater installations is given in Fig. 7-99.

7.100 What is a rumble filter?—A filter used in a record-reproducing sound system for removing rumble created by the turntable mechanism. Generally, these filters, because of their very low frequency characteristics, are composed of only capacitance and resistance, and are connected in the preamplifier stages of the amplifier system.

7.101 What is a line-noise suppression filter?—A filter used in the power line at the primary of a power transformer to prevent radio and line noises from entering the power snpply circuits and causing noise in amplifier systems. A typical circuit for such a filter is shown in Fig. 7-101.

Filters of this natnre may be used in either ac or dc lines. The coils are air core and wound with quite large wire capable of carrying the full current load of the equipment. The capacitors are either paper or oil insulated. (See Question 24.71.)

7.102 What method is used by the motion picture industry to identify the

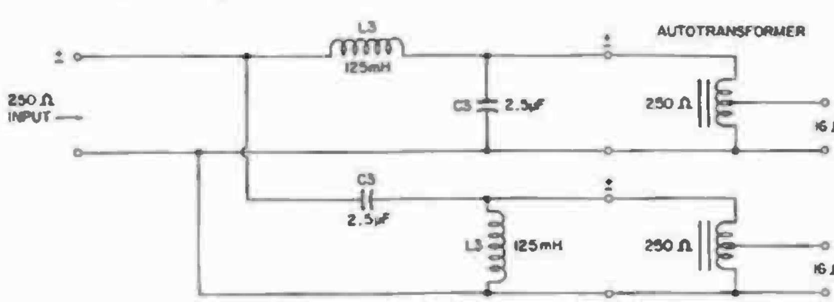


Fig. 7-99. Configuration for 12 dB per octave crossover network used in motion picture theater sound installations. Autotransformers are used at the speaker end to match the low impedance of the speaker voice coils. The network is mounted at the speakers on the stage.

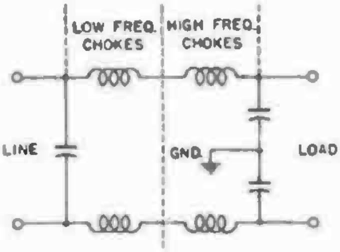


Fig. 7-101. Configuration for a line-noise suppression filter.

frequency characteristics of wave filters? —The method recommended by the Motion Picture Research Council for marking the nameplate of filters is as follows: Two frequencies which are down 3 and 10 dB with reference to 1000 Hz are stated. As an example: A high-pass filter designated 50-HI-42 would indicate that the frequency response at 50 Hz is down 3 dB and at 42 Hz 10 dB. A second example: An 80 Hz, high-pass filter is designated 90-HI-78. This would indicate that 90 Hz is down 3 dB and 78 dB is down 10 dB. The configuration and impedance are also stated.

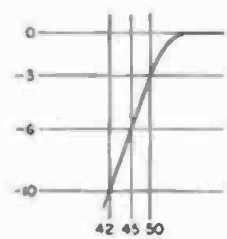
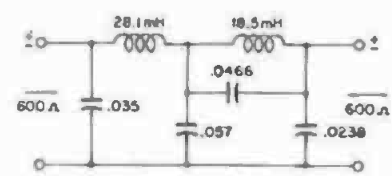


Fig. 7-102. The frequency characteristics of a 45-Hz high-pass filter designated 50-HI-42.

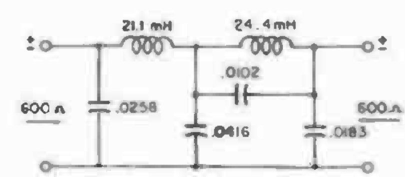
7.103 What are the impedance and phase-shift characteristics for constant-k and m-derived crossover networks?—In general, the impedance characteristics of a corresponding series or parallel networks are *inverse* to each other, and vary from one to the other. If a network is terminated at the output in R. ohms, the impedance at the input is the same. However, since the voice coil does not present a constant load impedance, this does not hold true. This effect is somewhat offset by the isolating effect of the intervening filter sections, particularly for the networks in Figs. 7-89A and C. For the m-derived networks in Figs. 7-89B and D, a slight improvement of the input impedance characteristic can be obtained by using the value of 0.45 for m. The improvement being small, little is gained by making this change.

As a rule, the designer of a crossover network finds little interest in the phase-shift characteristics; however, the following information is approximately correct. For the m-derived networks of Figs. 7-89A and C, the phase shift at the crossover is approximately 321°. For the networks of Figs. 7-89B and C approximately 221°. The phase shift at crossover for constant-k networks in Figs. 7-82A and C is approximately 90°, and for Figs. 7-82B and D it is approximately 180°.

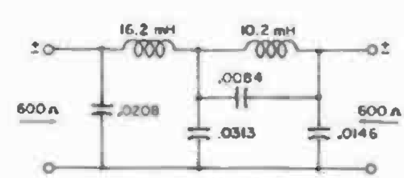
7.104 What are the circuit element constants for a 6-kHz, 8-kHz, and 10-kHz, 600-ohm low-pass filter having sharp cutoff characteristics? — Three low-pass filters that may be used for recording purposes are given in Fig. 7-104.



(*a*) *6000 Hz.*



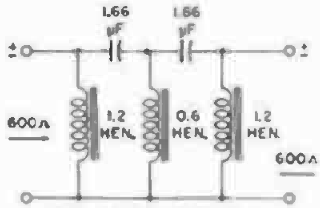
(*b*) *8000 Hz.*



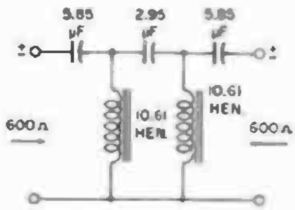
(c) *10,000 Hz.*

Fig. 7-104. Three 600-ohm low-pass filters suitable for recording purposes.

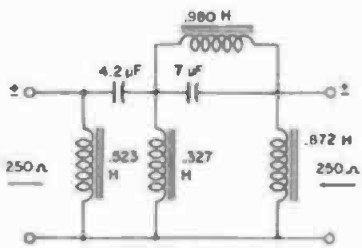
7.105 What are the circuit element constants for a 45 Hz and 80 Hz, 600-ohm high-pass filter?—Three high-pass filters suitable for recording purposes are shown in Fig. 7-105. The configuration shown at (c) has the sharpest cutoff characteristics.



(a) *80 Hz-600 ohm.*



(b) *45 Hz-600 ohm.*



(c) *80 Hz-250 ohm.*

Fig. 7-105. High-pass filters that are suitable for recording.

7.106 What is a wratten filter?—An optical filter used for filtering a given band of light. They are used in film recorder optical systems when recording directly on color film (direct positive) and are also extensively used in photography.

An optical filter may be likened to an electrical filter. A red optical filter appears red to the eye because it absorbs green and blue light rays, but passes red light freely. Such filters are made of glass and gelatin.

7.107 Give a schematic diagram for a 20,000-Hz low-pass filter for eliminating the high-frequency bias current in the output when making frequency-response measurements on a magnetic recorder.—When making frequency-response measurements on a magnetic recorder using the playback circuit, the bias current will often affect the frequency-response measurement. To eliminate this effect, a 20,000 Hz low-pass filter is connected in the output of the playback circuit which eliminates the effect of the bias current. Such a filter is shown in Fig. 7-107A, with its frequency-response curve shown in Fig. 7-107B. The filter configuration is designed for 550 ohms which will permit it to be operated in either a 500- or 600-ohm circuit. If a vacuum-tube voltmeter is used at the

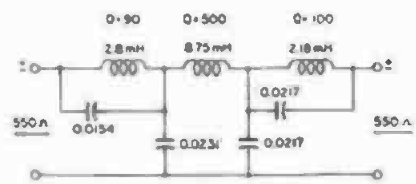


Fig. 7-107A. A 20-kHz low-pass filter for removing the high-frequency bias current in the output of a magnetic tape recorder when making frequency-response measurements.

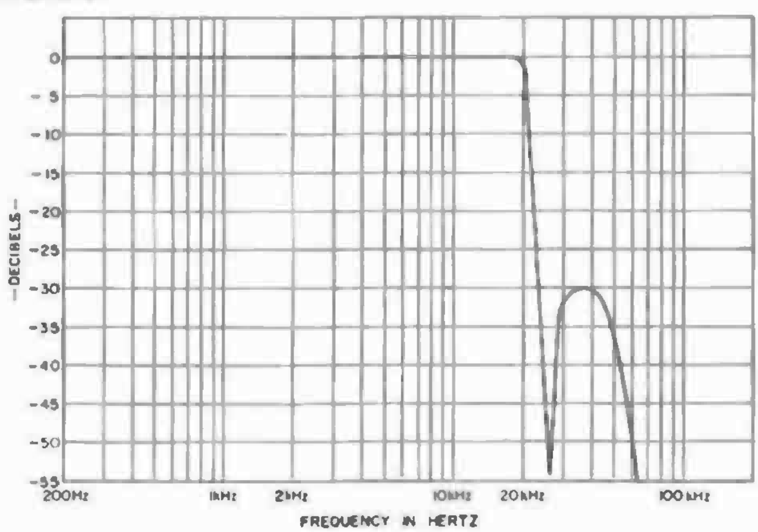


Fig. 7-107B. Frequency response of 20-kHz low-pass filter shown in Fig. 7-107A.

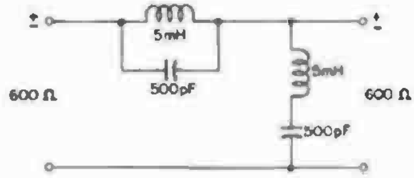


Fig. 7-107C. Low-pass filter for eliminating the effects of bias current when making measurements on magnetic recorders.

output of the filter, the filter must be terminated in either 500 or 600 ohms. If not, the filter frequency characteristics may be altered and the frequency response of the recorder under test may be affected. A second filter configuration, having similar characteristics to that in Fig. 7-107A, is shown in Fig. 7-107C.

7.108 What is an electronic crossover network?—An amplifier with variable RC circuits which are used to control the frequency of crossover, in a multiple loudspeaker system. Electronic crossover networks are connected after the preamplifier stage and feed two power amplifiers, one for the high-frequency speakers and one for the low-frequency speakers.

Electronic crossover networks are discussed in Section 20.

7.109 Describe the construction of a variable filter panel.—Variable filter panels are quite handy in making measurements on almost any type of electronic equipment, and they are used extensively in the laboratory and in sound recording systems. A variable filter panel, incorporating a high-pass and low-pass section, is shown in Fig. 7-109A. The two filter sections are completely isolated from each other, but may be connected in tandem by strapping the binding posts on the front of the panel. An octave bandswitch associated with each filter section, permits the cutoff frequency to be changed in octave steps, in conjunction with a multiplier dial for tuning the cutoff frequency over a range of one octave. The multiplication factor read on the dial multiplied by the octave band setting indicates the frequency, and when used with the tables at the upper portion of the front panel requires no calculations. As an example, setting of the octave bandswitch to 300 Hz and the multiplier to 1.8, results in a cutoff frequency of 540 Hz. The cutoff frequency is defined as the frequency attenuated 3 dB down from the maximum insertion loss.

By connecting the high-pass and low-pass sections in tandem, the device becomes a bandpass filter, without increasing the insertion loss which is so common to most bandpass filters with a bandwidth of ⅓ octave. Setting the high- and low-frequency sections to the same frequencies narrows the bandwidth to ¼ octave. Using an external 10-dB pad connected to the binding posts between the high- and low-frequency cutoffs will reduce the passband to approximately ⅙ octave.

Fig. 7-109A. Model 2AR variable LC filter panel manufactured by the Allison Laboratories, Inc.

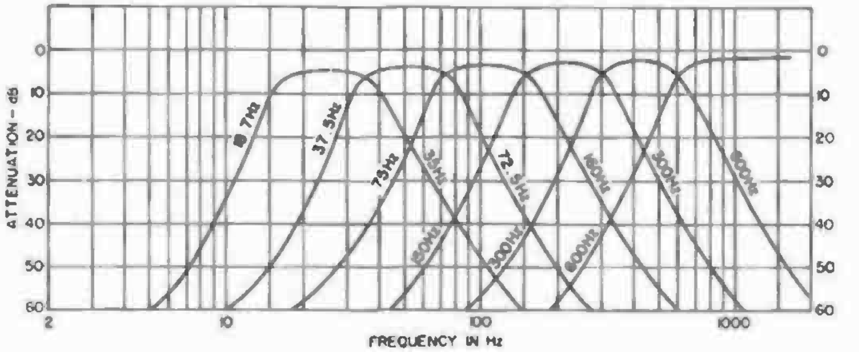


Fig. 7-109B. Frequency response for Allison Laboratories Model 2AR variable filter panel.

The normal rate of cutoff for both sections is 30 dB per octave, when properly terminated with an input and output impedance of 600 ohms. Measurements down to minus 120 dB below 1 volt are possible, without interference. The frequency response for bandpass operation below 1000 Hz is shown in Fig. 7-109B.

Variable filters are used extensively in rerecording sound tracks for the removal of unwanted sounds above and below the dialogue range. They are also used for creating narrow band response to simulate a telephone, dictating machine, radio, or an intercommunicating system. A variable sound-effects filter panel of this type is shown in Fig. 7-109C, with its many frequency characteristics shown in Fig. 7-109D. Two switches are provided for cutting the filter panel in and out of the circuit and for operating as a high-pass, low-pass or bandpass filter. The insertion loss is less than 0.25 dB. The coils are toroidal wound to eliminate stray magnetic fields. The filter configurations are of the constant-k type. For sound effects, the cutoff rate is generally not less than 12 dB per octave, and not more than 18 dB per octave.

7.110 Describe the basic configurations for double-m and double-m prime filter sections.—Although an m-derived filter section can be designed to provide an impedance match over about 85 percent of the passband, such a match may not be sufficient for certain filter applications. With further modification of the m-derived equations, it is possible to develop other m-derived sections from existing m-derived sections, or from their prototype, the constant-k section. Such sections are termed mm′ and mm″ filters, and can be designed to have a more constant impedance in the passband than that which can be obtained with the conventional m-derived section. Double m-derived sections provide an impedance match within 2 percent over 96 percent of the passband. However, their attenuation characteristics remain the same as for the m-derived section of a similar design.

Typical examples of the double-m or mm′ filters are shown in Fig. 7-110. The design procedure is somewhat similar to that of the m-derived. First the filter is designed as a constant-k section, then the circuit elements are modified, using the factors given in the diagram. The filter sections shown are termed termi-

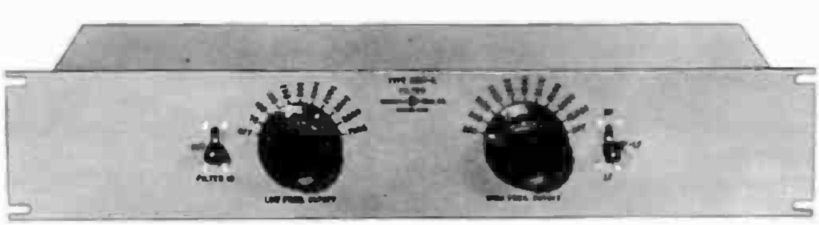
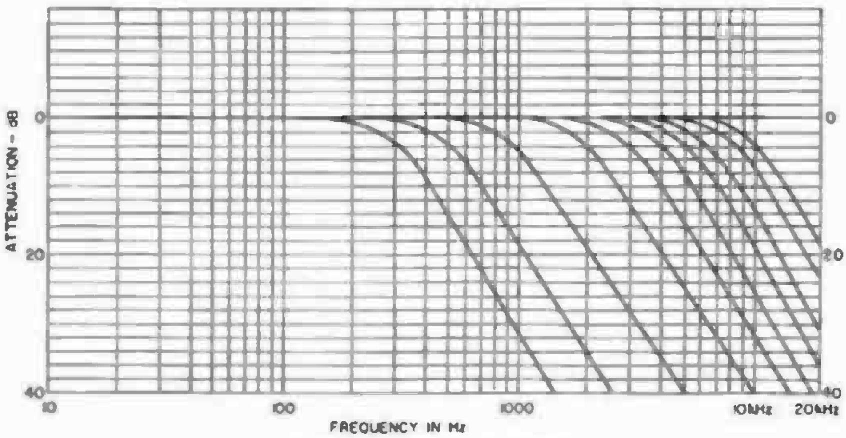
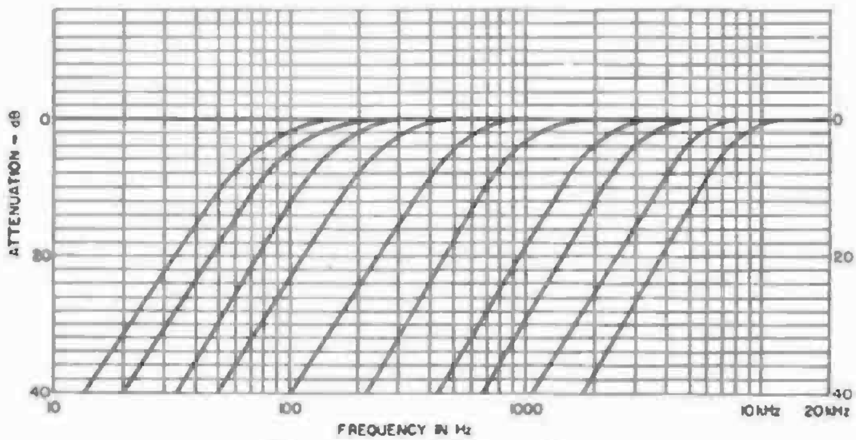


Fig. 7-109C. Model 6517-E variable sound-effects filter panel manufactured by Cinema Engineering Co. Frequency range, 70 Hz to 10,000 Hz.



(a) *Used as a low-pass filter.*



(b) *Used as a high-pass filter.*

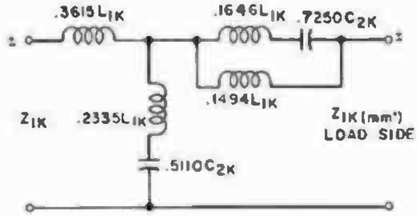
Fig. 7-109D. Frequency-response characteristics for Cinema Engineering Co., Model 6517-E sound-effects filter panel.

nal sections and are used as end sections. These sections may be used with other m-derived and constant-k sections, provided the image impedance match is satisfied between the sections. The mm'-derived filters have been included to show only the basic conception of design. The reader is referred to the references.

7.111 What are the design characteristics for filters to be operated in parallel or series?—Wave filters to be operated in either series or parallel require special treatment, because the impedances are joined to a common source impedance. Since filter impedances vary with frequency in the stop band, both the parallel impedance and the insertion loss are affected. The loudspeaker crossover networks of Fig. 7-111 are typical examples of parallel and series filter combinations.

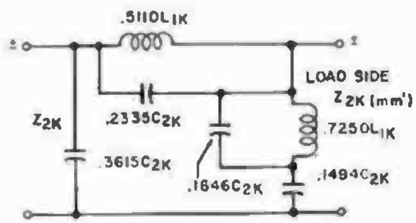
For parallel operation, the low- and high-pass sections must employ T-type intermediate sections with an L-type input half-section, designed for m equal to 0.6. With equal load impedances at the output terminals of each filter, a normal impedance match is achieved by omitting the shunt impedance (Z_4 and Z_3) at the input of each filter section. For series operation the filter sections are designed using pi intermediate sections, with pi input half-sections. Again using a value of m equals 0.6, and using equal termination impedances at the output, a normal impedance match is obtained by omitting the series impedance at the input of each filter section.

When low- and high-pass sections are connected in parallel, the circuit elements of the high-pass filter replace the shunt inductance and capacitance of the low-pass filter of the passband.



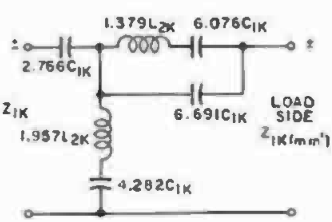
$L_{1K} = \frac{R_0}{\pi f_0}$

(a) *Series-derived, low-pass terminal section.*



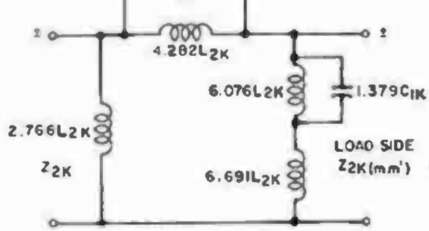
$C_{2K} = \frac{1}{\pi f_0 R_0}$

(b) *Shunt-derived, low-pass terminal section.*



$L_{2K} = \frac{R_0}{4\pi f_1}$

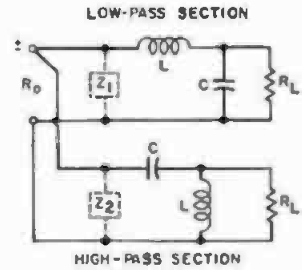
(c) *Series-derived, high-pass terminal section.*



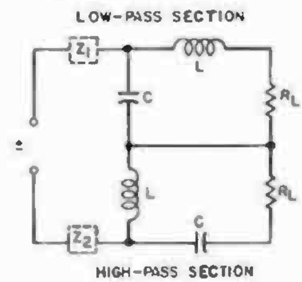
$C_{1K} = \frac{1}{4\pi f_1 R_0}$

(d) *Shunt-derived, high-pass terminal section.*

Fig. 7-110. Double m- or mm'-derived filter sections.



(a) *Parallel-connected.*



(b) *Series-connected.*

Fig. 7-111. Filters connected in parallel and series.

The circuit elements of the low-pass filter replace the inductance and capacitance of the high-pass filter of the passband. When connected in parallel, the two filters will exhibit about the same amount of attenuation and impedance characteristics as if the input elements Z_1 and Z_2 had not been removed. When connected in parallel, about 12 dB attenuation per octave may be expected.

In the series circuit, the procedure used in the parallel network is not necessary, as filters connected in series are inverse to each other. The foregoing design procedures are applicable in all instances where filters are complementary: that is, where frequencies that lie in the stop band of one filter are in the passband of the other.

7.112 Describe the basic principles of a phase-shift or phase-correction network.—An ideal transmission line would be one in which the received signal currents represent a faithful copy of the transmitted signal—in other words, a distortionless transmission system. For relatively short distances, the effect of phase distortion is not appreciable; however, this is not true for long-distance transmissions. Here the phase distortion can become serious enough to impair the commercial efficiency of the line. The correction of frequency dis-

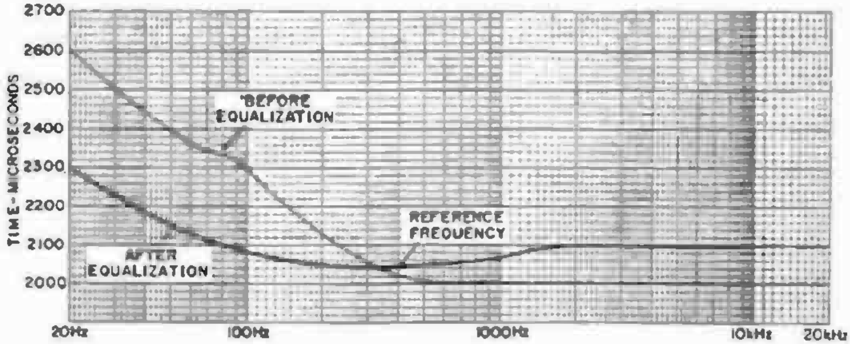


Fig. 7-112A. Time-delay characteristics of a transmission line equalized ±1 dB, 10 to 20,000 Hz.

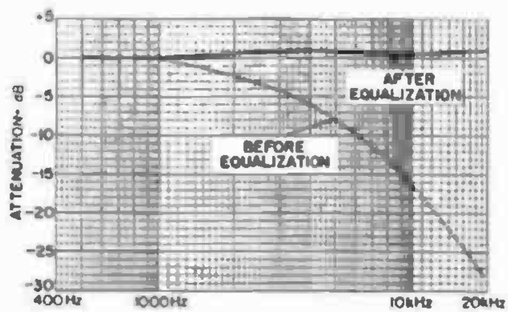


Fig. 7-112B. Frequency characteristics of the transmission line in Fig. 7-112A.

tortion by the use of equalizers is not enough because as the equalizatiou is increased, the phase distortion is likewise increased. For high-quality traus-missions, the phase distortion must be reduced to a negligible amount.

Distortion due to frequency variations and phase differences between the signal at the sending and receiving ends of the line gives rise to what is termed transient effects. The signal after arriving at the far end of the line requires an appreciable time, which varies with frequency, to build up, and may at times never build up to anything resembling the transmitted signal. To correct for these deficiencies, filter networks, termed phase-correction or all-pass filters, are used. Such networks have zero attenuation for all frequencies within their design range. They are used as phase-correction equalizers to induce a time delay for a given group of frequencies.

By applying a complex waveform consisting of many frequencies to a long transmission line, each frequency will be delayed in its transmission a different length of time, the delay time increasing with the length of the line and

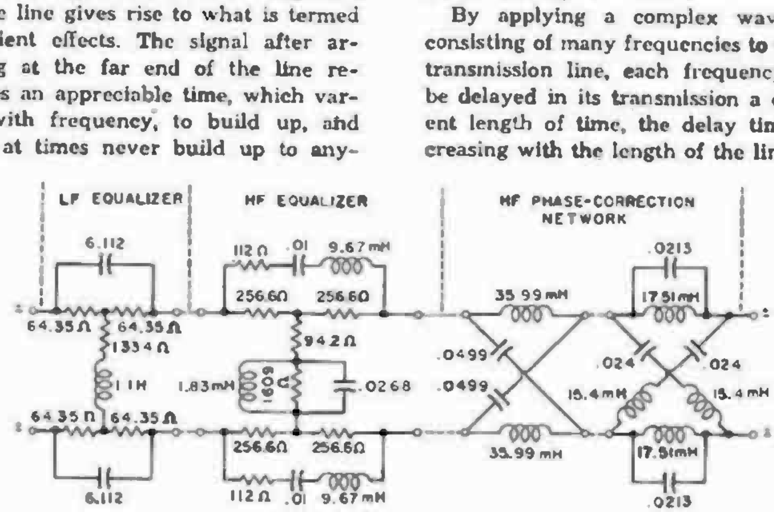


Fig. 7-112C. Low- and high-frequency equalizers and a phase-correction network.

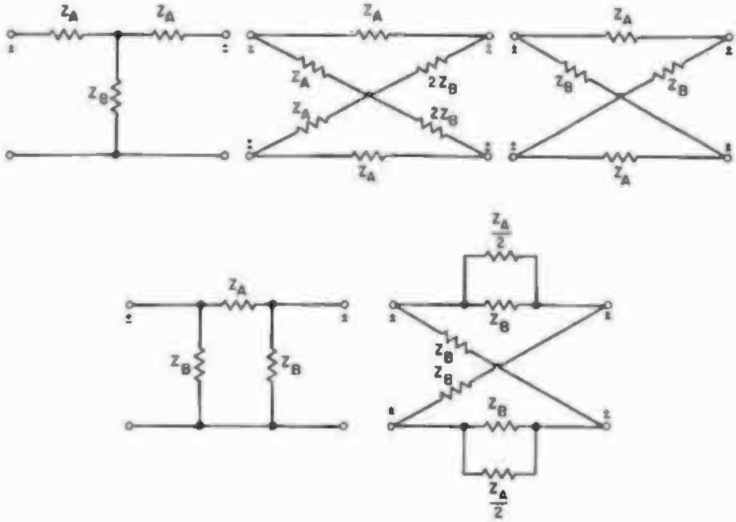


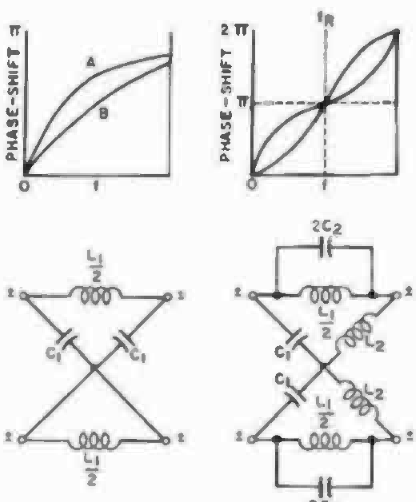
Fig. 7-112D. Converting T and pi networks to lattice networks.

the frequency. The actual time delay is not important, but the relative time delay between the frequencies is important. The difference in phase between the signal at the sending and at the receiving end of line can be expressed in terms of time through the line:

Phase-shift $= \gamma\omega + n\pi$ radians

where,

γ is the time delay of the network,
ω is equal to $2\pi f$,
and n is an integer.

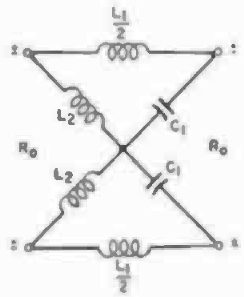


(a) *Simple network.* (b) *Complex network.*

Fig. 7-112E. Phase-shift characteristics for two different types of networks.

For a network to have zero phase shift γ must remain constant for all frequencies, or the curve of the phase shift, as a function of frequency, must be a straight line. When a complex waveform is applied to a system which has a linear phase shift, the transmitted signal is received at the far end of the line after a definite time delay, but all frequencies in the waveform will be received at the same instant in the same relationship as they were at the sending end of the line. A time delay at the high frequencies is more noticeable than at the low frequencies because the human ear depends on the higher frequencies for definition. In the range between 5000 and 8000 Hz, it has been determined that the time-delay should not exceed the 1000-Hz delay time by more than 10 milliseconds. The delay at 50 Hz may be up to 75 milliseconds more than that at 1000 Hz, without noticeable effects. The low frequencies are delayed by the presence of series capacitance and shunt inductance, the delay increasing with the value of the circuit constants. High frequencies are delayed by the presence of series inductance and shunt capacitance, the delay increasing with the increase of these values.

The effect of phase distortion for a transmission line 285 miles in length and terminated in 600 ohms at each end, is shown in Fig. 7-112A. This line, an experimental one, was designed to have a frequency response of plus or minus 1 dB, from 10 to 20,000 Hz. This required



$$f_C = \frac{1}{\pi\sqrt{L_K C_K}}$$

$$R_o = \sqrt{L_K C_K}$$

$$L_K = \frac{R_o}{\pi f_C}$$

$$L_1 = \frac{m}{2} L_K$$

$$L_2 = \frac{1}{2m} L_K$$

$$C_K = \frac{1}{\pi f_C R_o}$$

$$C_R = \frac{m}{2} C_K$$

$$\text{time} = \frac{2m}{\sqrt{1-(f/f_C)^2}\left[1-(1-m^2)(f/f_C^2)\right]}$$

Fig. 7-112F. Configuration for an all-pass network and its design equations.

that 20,000 Hz be equalized 28 dB, with reference to 500 Hz. After the line was equalized (Fig. 7-112B) it was discovered the high-frequency equalizer induced phase distortion, more than twice as great as that of the line itself. To correct for this deficiency, a phase-correction network of the design of Fig. 7-112C was connected in tandem with the equalizer networks. Time-delay networks can not be developed using the conventional ladder-type networks; instead, symmetrical m-derived networks of the T- or pi-type are used. Conversion of a T section and a pi section to a lattice network is shown in Fig. 7-112D. All-pass networks are developed by the use of lattice networks, in which the reactances Z_A and Z_B are reciprocal with respect to image impedances R_o. This results in a passband in which the attenuation is zero, and the image impedance is a constant resistance (R_o) for all frequencies.

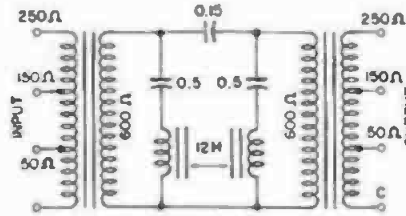
The phase-shift characteristics for two different types of networks are given in Fig. 7-112E. Fig. 7-112E part (a) indicates the phase-shift characteristic when the L/C ratio is large, and part (b) indicates when the L/C ratio is small. Phase-correction networks have been used with success in disc recording systems employing diameter equalization, and pre- and post-equalization. Fig. 7-112F gives the equations used in the design of all-pass networks.

7.113 Describe a low-frequency, high-pass filter for use with microphones.—A rather simple 100-Hz filter for use with a microphone to reduce the effects of wind noise is shown in Fig. 7-113, part (a). The device consists of a pi section, high-pass filter connected between two impedance-matching transformers. This filter is generally made and packaged in a metal box with standard microphone connectors at each end, for connecting directly into the microphone cable. The taps on the transformers permit the filter to be used with different impedances. The coils are toroidal wound, high-Q, and are encased in a mu-metal box to eliminate the effect of stray magnetic fields. The frequency response is shown in Fig. 7-113 part (b).

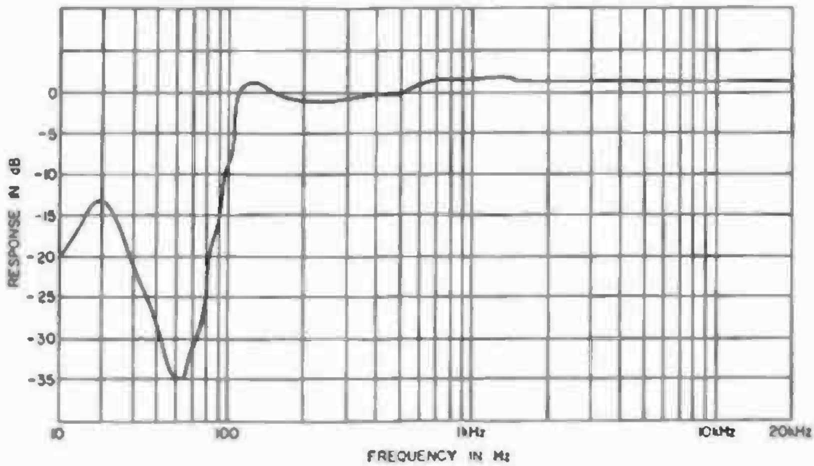
7.114 Describe the characteristics and use of a pink-noise filter.—A pink-noise filter (Fig. 7-114A) converts the output signal of a random-noise generator to a constant energy per octave generator which facilitates measurements to constant-percentage bandwidth analyzers. The circuit in Fig. 7-114B is an RC low-pass network with a slope of 3 dB per octave from 20 to 20,000 Hz, and a slope of 16 dB per octave at the higher frequencies shown in Fig. 1-114C.

If the output of the random-noise generator is connected to a constant-percentage bandwidth analyzer, the output-voltage characteristic appears as white noise (Fig. 7-114D). Connecting the pink-noise filter in the output produces an output signal as shown in Fig. 7-114E, termed pink noise. The white- and pink-noise spectrums shown were made with a one-third octave bandwidth analyzer and recorded using a graphic level recorder.

It is interesting to note that certain noises occurring in nature are closer in spectral characteristics to pink noise



(a) *Schematic diagram.*



(b) *Frequency-response curve.*

Fig. 7-113. A 100-Hz high-pass wind noise filter for a microphone.

Fig. 7-114A. General Radio Co. Model 1390 pink-noise filter.

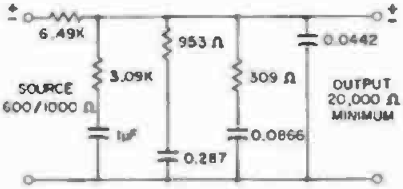


Fig. 7-114B. Circuit configuration for pink-noise low-pass filter.

than to white noise. This is also characteristic of low-frequency noise found in semiconductor devices and in certain acoustical background noises.

7.115 Describe an RC filter for making noise measurements.—When making conducting-noise measurements, particularly on magnetic tape or film equipment, the noise is measured while erasing the high-frequency bias current signal. To properly measure the noise, a filter (Fig. 7-115A) with the frequency characteristic shown in Fig. 7-115B is required to attenuate noises outside the audio spectrum. The filter shown is designed to operate in a 600-ohm circuit. Noise measurements are discussed in Section 23.

7.116 What are Butterworth, Chebyshev and elliptic filters?—Such filters are used for rf and crystal bandpass design. The equations used replace the

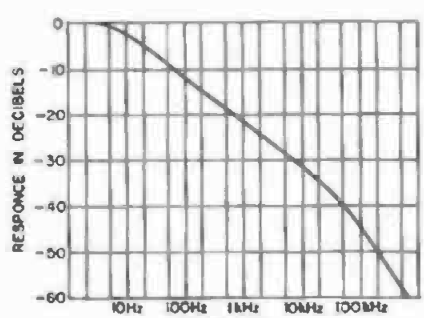


Fig. 7-114C. Frequency characteristic of pink-noise low-pass filter.

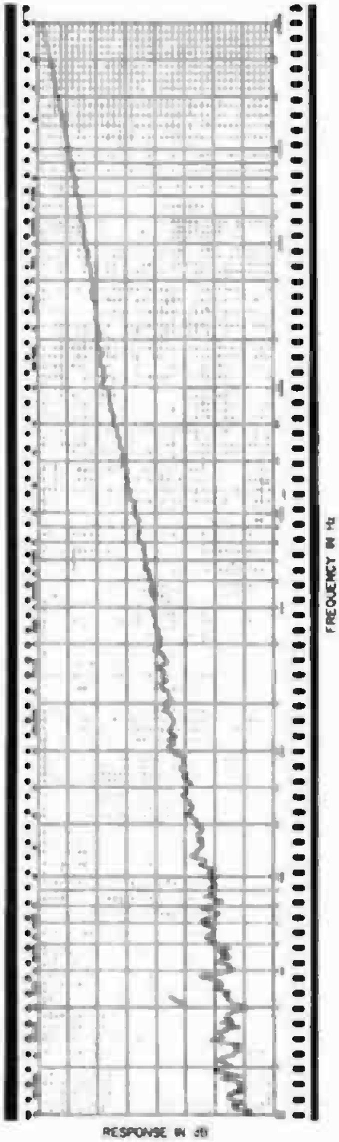


Fig. 7-114D. White-noise output of General Radio Co., Model 1390B random-noise generator without pink-noise filter.

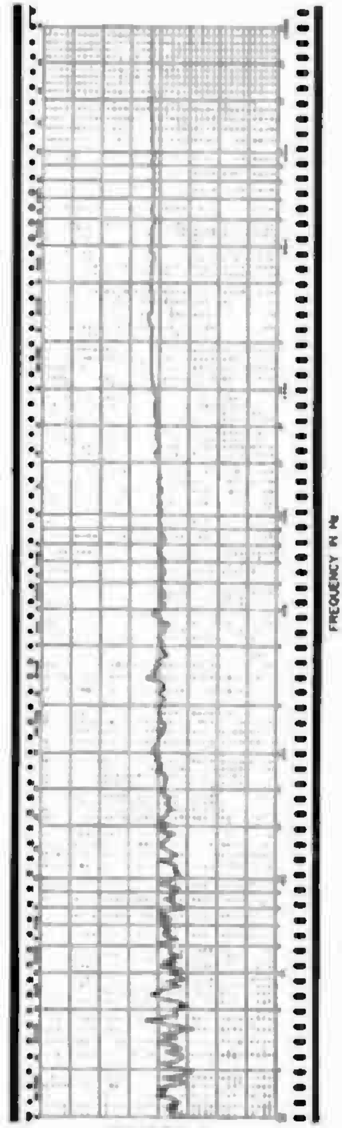


Fig. 7-114E. White-noise output of General Radio Co., Model 1390B random-noise generator with pink-noise filter.

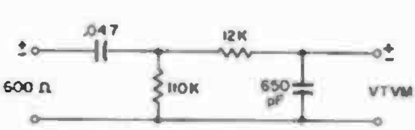


Fig. 7-115A. Filter configuration for measuring erasure noise in magnetic tape recorder.

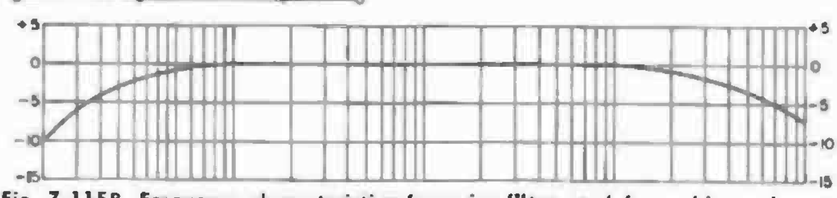


Fig. 7-115B. Frequency characteristics for noise filter used for making noise measurements on magnetic recording equipment and similar devices.

conventional equations used for the design of constant-k and m-derived filters.

Conventional filters are designed section by section, matching the impedance of each section to the other. Butterworth, Chebyshev and elliptic filters are designed as a whole which simplifies the matching of the sections, permitting the response to be optimized, and are expressed using transfer equations. As a rule, the configuration is either a ladder or lattice network.

Bessel filters, because of their excellent phase-shift linearity characteristics, are often referred to as linear phase- shift filters. Filters of this design are generally fabricated using solid-state techniques in the form of an operational amplifier. For mathematical treatment of such devices the reader is referred to the references.

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Section 8

Transformers and Coils

Transformers and coils play an essential part in the field of Audio Engineering. Types such as input, output, intermediate, repeat, power and modulation, to mention but a few, often serve as a key component in audio circuitry. Such devices are used for matching impedances, isolation, bridging low-impedance circuits, and in power supplies.

Work by men such as Clark, Lenz, Maxwell, and Steinmetz has added greatly to our understanding and usage of transformers and coils. The various segments, construction detail, and design characteristics of these devices, are discussed in this section, with charts and graphs for determining their characteristics, when used under conditions other than for which they were designed. Constant-voltage transformer, their design and use, are also covered.

8.1 What is a transformer?—An electrical device consisting of one or more coils wound on a magnetic material core. Electrical energy may be transmitted between one or several of the windings. Transformers designed for audio frequency use must be capable of transmitting currents over a wide frequency range, in contrast to a power transformer which transmits currents over a narrow frequency band. Transformers for either service may be designed to step-up or step-down the current. Typical schematics for an audio and a power transformer are shown in Figs. 8-1A and B, respectively.

8.2 Why are laminated cores used in transformers?—To reduce eddy-current loss in the core material, thereby improving the transformer characteristics. Hysteresis losses in the core are caused by molecular friction and are reduced by the use of alloy steel, such as silicon, hypersil, and hypernic. The latter two metals are manufactured by Westinghouse.

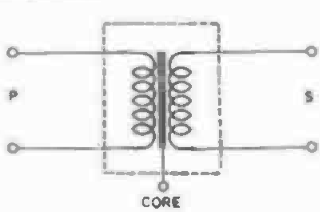


Fig. 8-1A. A typical audio transformer.

8.3 What takes place in a transformer when the core material is saturated?—When the core becomes saturated in a power transformer, it is no longer able to produce additional lines of force. Hysteresis losses occur, the coils become overheated, and damage to the winding may take place under these circumstances.

In an audio-frequency transformer, if the core cannot produce the required lines of force the inductance drops and, because of the core saturation, the core losses increase, and the frequency response is affected, particularly at frequencies below 100 Hz.

8.4 What is copper loss?—A power loss caused in a power transformer by the dc resistance of the coils. The loss varies as the square of the current.

8.5 What is iron or core loss in a transformer?—The loss of energy due to

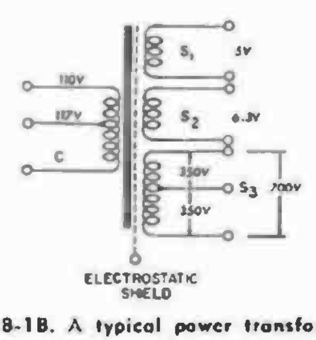


Fig. 8-1B. A typical power transformer.

eddy currents in the core material. In practical transformers, the core loss remains almost constant from no load to full load, under normal conditions. (See Question 23.158.)

8.6 ***What are eddy currents?***—Circulating currents induced in a conducting material as the result of the rapid reversal of the magnetic field. (See Question 8.2.)

8.7 ***What causes current flow in the primary of a power transformer operating with an unloaded secondary?***—The primary impedance has both an ac reactance and a dc resistance, which will cause a current flow. In well-designed transformers this current is small and is referred to as the mag-I current, or magnetizing current.

8.8 ***What is an autotransformer?***—It is a transformer with a single winding as shown in Fig. 8-8A. Such transformers may be designed for stepping the voltage up or down, and may be used with any type equipment. Autotransformers are also constructed to be continuously variable over their entire range. The interior construction of a variable autotransformer manufactured by General Radio Co. and sold under the trade name of *Variac* is pictured in Fig. 8-8B.

The autotransformer used in this device consists of a single layer winding on a toroidal core. As the shaft is rotated, a brush contact traverses the winding turn-by-turn; thus, the voltage may be varied from zero to full voltage. A tap on the winding permits the input voltage (line) to be changed, resulting in an overvoltage connection (zero to 140 volts for a line voltage of 117 volts). Autotransformers are sometimes used in audio amplifiers for interstage coupling and for impedance matching in loudspeaker systems.

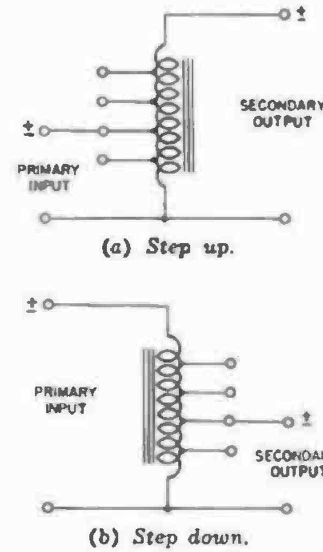


Fig. 8-8A. Two types of autotransformers.

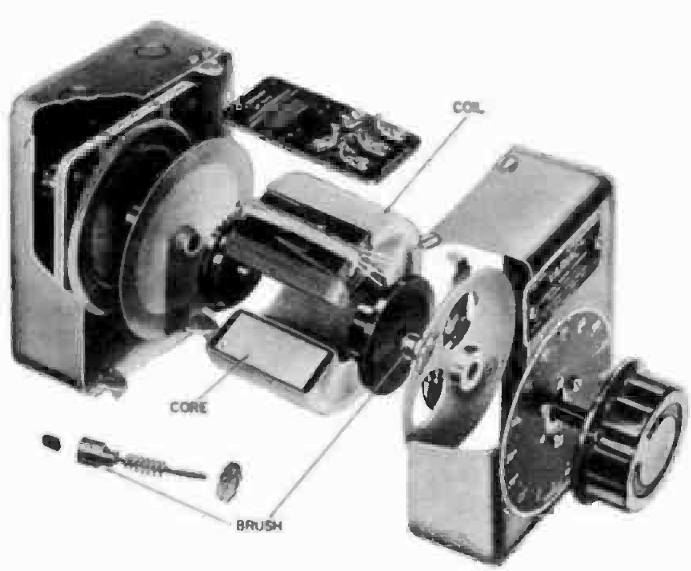


Fig. 8-8B. Interior construction of a General Radio Company Variac, showing the autotransformer winding.

8.9 What is a bifilar winding?—A coil is said to be bifilar wound when, to achieve a balance between windings, two conductors are wound side by side. (See Fig. 5-16B.)

8.10 What is a pie winding?—A thin sectional winding used in transformers and coils to reduce the effect of distributed capacity between the windings.

8.11 How is the wattage rating of a transformer determined if rated in volt-amperes?—The power rating of a transformer depends on the amount of heat the windings can stand. The power rating in watts is the product of the volt-ampere rating times the power factor. Small transformers employed in electronic equipment generally have efficiencies of 95 to 98 percent; therefore, efficiency as a rule is ignored. Assume a power transformer is rated 117 volts, 60 Hz, 100 volt-amperes, with a secondary of 6.3 volts. The rating of 100 volt-amperes means the product of voltage times the current ($E \times I$). Since the secondary can never exceed 100, and if it has a load of 12 amperes, the wattage will be 75.5 watts, ignoring the power factor. If the power factor is to be taken into consideration, the rating of 100 volt-amperes is multiplied by the power factor. For a power factor of 0.9, the maximum wattage at the secondary is $100 \times 0.9 = 90$ watts.

To convert the volt-ampere rating into kilovolt-amperes, divide the volt-ampere rating by 1000. The transformer would then be rated 0.100 kilovolt-amperes. *If the transformer has more than one secondary winding, the total load wattage cannot exceed the 100 volt-ampere rating.* Power factor is discussed in Questions 3.14, 3.58, and 3.62. Other formulas used with power transformers are shown in Fig. 8-11. For 3-phase circuits I and E are per phase.

8.12 Define an electrostatic shield and its purpose in a power transformer.—An electrostatic shield consists of a single turn of copper or brass covering the entire length of the primary winding. The ends of the shield are insulated from each other to prevent the shield from acting as a shorted turn. A connection to the shield is brought out to a separate terminal for grounding. The purpose of the shield is to prevent the passage of line noises and radio-frequency signals from the transformer into the power supply, and thence to other parts of the equipment. Electrostatic shields are also used in audio transformers. (See Question 8.42.)

8.13 What is the effect of connecting a power transformer designed for 60-Hz operation to a 25-Hz line?—The inductive reactance of the primary is considerably lower when connected to the 25-Hz voltage source. This causes the primary current to be more than double its normal value. The increased current through the primary winding will cause serious overheating and, eventually, will cause the primary to burn out.

To Find	Knowing	1-phase	3-phase
Amperes	E, hp, PF, Eff.	$\frac{hp \times 746}{E \times Eff. \times PF}$	$\frac{hp \times 746}{1.73 \times E \times Eff. \times PF}$
Amperes	kW, E, PF.	$\frac{kW \times 1000}{E \times PF}$	$\frac{kVA \times 1000}{1.73 \times E \times PF}$
Amperes	kVA, E.	$\frac{kVA \times 1000}{E}$	$\frac{kVA \times 1000}{1.73 \times E \times PF}$
Watts	E, I, PF.	$E \times I \times PF.$	$E \times I \times 1.73\ PF$
Kilowatts	E, I, PF	$\frac{E \times I \times PF}{1000}$	$\frac{E \times I \times 1.73 \times PF}{1000}$
kVA	I, E.	$\frac{E \times I}{1000}$	$\frac{E \times I\ 1.73 \times PF}{1000}$

where,
E is voltage,
I is current,
Eff is efficiency,
PF is power factor,
hp is horsepower,
kW is 1000 watts.

Fig. 8-11. Formulas used with power transformers.

Transformers designed for 25-Hz operation have about 50 percent more core material than do transformers designed for 50- to 60-Hz operation.

8.14 Can a transformer designed for 60 Hz be operated on 400 Hz?—If the core material is of a good quality, a transformer which has been designed for 60 Hz may be operated on 400 Hz. However, a transformer which is designed for 400 Hz operation cannot be operated on 60 Hz, as the core is smaller for a given power rating.

8.15 What is a filament transformer?—A transformer used for supplying a source of voltage to the filament or heater circuit of a vacuum tube.

8.16 What is the electrical center of a coil?—The point of equal impedance. In audio transformers this may not always indicate equal dc resistance from the center to the ends of the winding. The reason for this is the center tap is made on the basis of *impedance* rather than dc resistance. However, in high-quality audio transformers, the dc resistance is the same for both sides of the coils because the windings are split into several separate coils and then connected in series. The center tap is taken at the junction of the coils. The internal connections for a 1:1 repeat coil, Fig. 8-16, shows how the coils are connected in series, and the taps taken for the different impedances.

8.17 What are the essential differences between an audio and a power transformer?—Power transformers are designed to operate at one frequency or over a small range of frequencies. Audio transformers must operate over a frequency range of 20 to 20,000 Hz. Many audio transformers are designed to operate over a frequency range of 10 to 200,000 Hz. Such wide frequency characteristics require careful design as to the method of winding the coils, their connection, material for impregnating the windings, core types and material, and the method used for stacking the laminations.

8.18 Define ampere turns.—The equations for determining ampere turns are given below:

$$\text{ampere turns} = \frac{V}{R} \times T$$

$$\text{ampere} = \frac{AT}{T}$$

$$\text{resistance} = \frac{V}{R} \times T$$

$$\text{effective turns} = \frac{\text{total resistance}}{\text{resistance of coil}} \times \text{turns of coil}$$

where,

A is the current in amperes,
R is the dc resistance,
T is the number of turns in the coil,
V is the voltage applied to the coil.

The latter takes into consideration any resistance external to the coil. One ampere turn equals 1.257 gilberts.

8.19 What is a current transformer?—A special transformer designed to work with a current-indicating instrument. The primary of the transformer is connected in series with the load, and the indicating instrument is connected to the secondary, as shown in Fig. 8-19. In this manner, the transformer carries the load while the instrument is com-

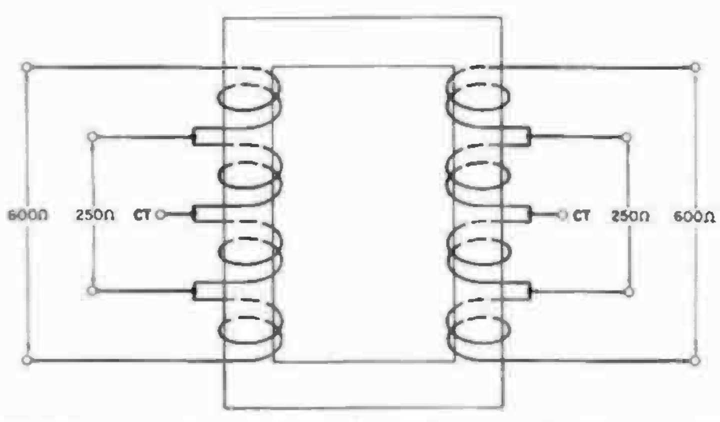


Fig. 8-16. Coil construction and winding for a 1:1 repeat coil with taps at 250 ohms and center.

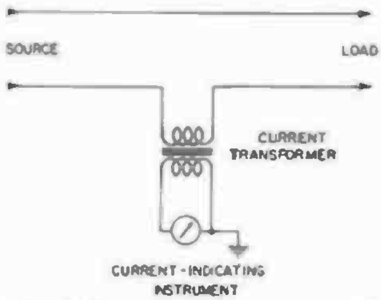


Fig. 8-19. Current transformer connected to indicate load current.

pletely isolated. The indicating instrument is actually a voltmeter, although calibrated in amperes.

8.20 How is the turns ratio calculated for a current transformer?

$$\frac{I_p}{I_s} = \frac{N_s}{N_p}$$

where,

I_p is the current in the primary,
I_s is the current in the secondary,
N_p is the number of primary turns,
N_s is the number of secondary turns.

8.21 How is the voltage regulation of a transformer calculated?—The voltage is first measured in a no-load condition, and then with a full load applied. This information is then substituted in the following equation:

$$\%\ \text{Regulation} = \frac{V_1 - V_2}{V_2} \times 100$$

where,

V_1 is the no-load voltage,
V_2 is the full-load voltage.

8.22 How are three-phase power transformers connected for utility service?—Several methods used for connecting 3-phase power transformers for commercial services are shown in Figs. 8-22A to F. Fig. 8-22A shows an open-delta connection and is the simplest of all 3-phase transformer connections as each winding consists of only two coils. In Fig. 8-22B is shown a six-winding transformer connected in delta. Figs. 8-22C and D illustrate two transformers connected in a star or wye configuration using a 3-wire and a 4-wire feed system. The 4-wire connection will permit the taking of 115 volts from any one winding to a neutral wire connected at the center of the three secondaries.

Fig. 8-22E shows a star-delta, or delta-wye, connection. Here, one side of the transformer winding is delta connected and the other side is star connected. In Fig. 8-22F is shown a 4-phase or quarter-phase connection. The various voltage combinations are shown relative to the common center tap and to the outside ends of the coils. (See Question 3.32.)

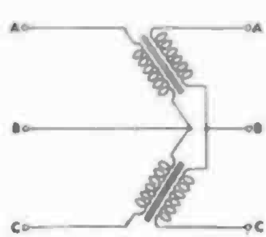


Fig. 8-22A. A three-phase transformer open-delta connection.

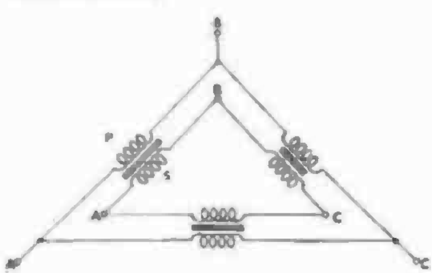


Fig. 8-22B. A three-phase transformer delta connected.

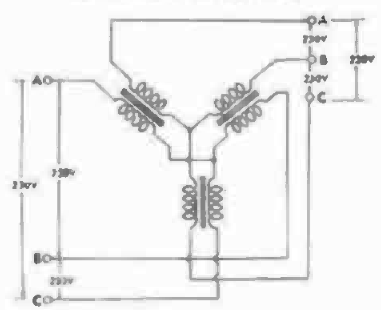


Fig. 8-22C. A star- or wye-connected three-wire, three-phase transformer.

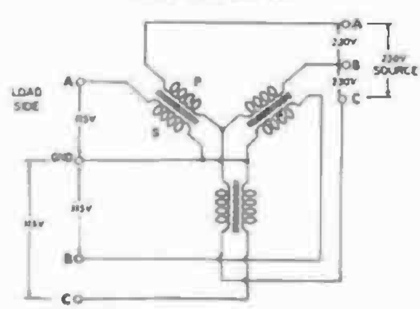


Fig. 8-22D. A star-connected, four-wire, three-phase transformer.

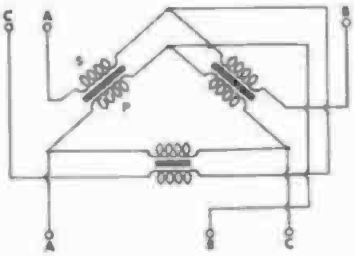


Fig. 8-22E. A three-phase transformer with windings connected in delta and wye.

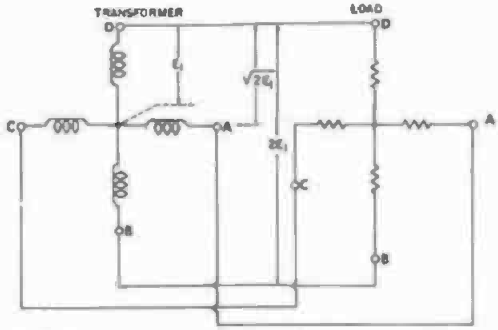


Fig. 8-22F. A four-phase or quarter-phase connected transformer.

8.23 What is a teaser transformer?—A power-transmission transformer used with 3-phase power-distribution systems, as shown in Fig. 8-23.

Assume a 1000-volt, 3-phase power source is to be stepped down to 100 volts. The primary of a 1:1 teaser transformer, T2, is connected in the center leg of the main power transformer as shown. The secondary of the teaser is connected to a center tap on the secondary of T1. The voltages developed by the system are given in the diagram. It will be noted the teaser transformer develops only 86.6 volts across both its primary and secondary windings.

The advantage of such a system is that moderate amounts of power may be transformed with only two transformers. About 58 percent of the system rating may be used with a teaser transformer, compared to one using a transformer in each phase. Because of dissymmetry, the system can easily become unbalanced. The phasing of the transformer windings must be carefully carried out to prevent damage.

8.24 Describe a repeat or isolation coil.—An audio transformer consisting of two windings with a one-to-one impedance ratio. However, repeat coils are generally designed with taps at various points on the coils to provide impedance matching between circuits of unequal impedance. A typical schematic of a repeat coil is shown in Fig. 8-24. It will be noted that although the coil is designed for an impedance of 600 ohms, ranging downward to 250 ohms, the center tap is balanced with respect to the center of the coil. High quality repeat coils used for transmission measurement work, as described in Question 23.14, employ an electrostatic shield between the two windings to prevent the transfer of noise by means of the capacitance existing between the two windings.

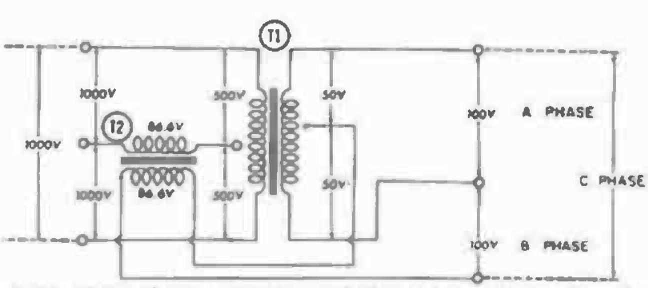


Fig. 8-23. Three-phase power transmission using a teaser transformer.

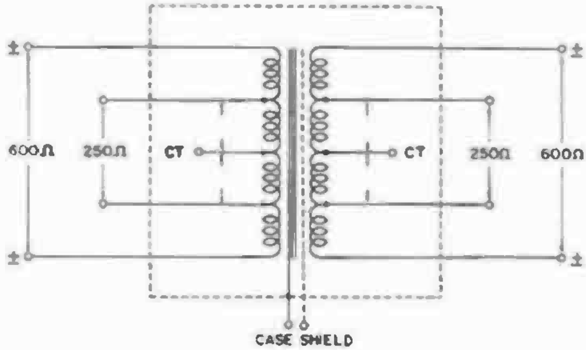
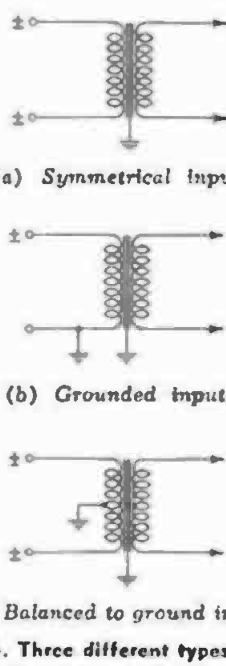


Fig. 8-24. Typical 600:600-ohm repeat or isolation coil. In-between impedances are indicated for ready reference.

Thus, a signal is transferred from one coil to the other coil inductively only.

The windings consist of several separate coils, each with the same number of turns and connected in series in such a manner that the capacitance between coils is reduced to a minimum. This method of construction permits the various impedance taps to be taken at the exact electrical center, resulting in a well balanced design. The insertion loss of the average repeat coil is about 0.10 to 0.5 dB.

Repeat coils are often referred to as isolation coils, when they are employed to separate an unbalanced circuit from a balanced one. This is demonstrated by the extensive use of such coils in mixer consoles.



(a) *Symmetrical input.*

(b) *Grounded input.*

(c) *Balanced to ground input.*

Fig. 8-25. Three different types of transformer-input circuits.

8.25 Define the terminology used with input transformers.—Terminology used with input transformers is often misused, especially when referring to grounded-input circuits. Fig. 8-25 illustrates three methods most commonly employed. At part (a) is shown an ungrounded symmetrical input, often erroneously referred to as a balanced input. A true balanced circuit is one using an input transformer with a center tap to ground, as shown in part (c). For the circuit of part (a) to be truly balanced (although no ground is used) the transformer windings must be wound so that the distributed capacitance to ground is the same for each coil. Also, each coil must have identical inductance and dc resistance, and be physically interleaved to reduce the effects of leakage inductance. If all these conditions are met, the coil may be said to be balanced. However, in the average coil these conditions do not always prevail, therefore this circuit is said to be symmetrical.

The circuit in part (b) is an unbalanced input with one side of the input winding connected to ground. This input is often used with circuits such as a high-impedance microphone employing a single conductor and shield. The shield is connected to the ground side of the transformer.

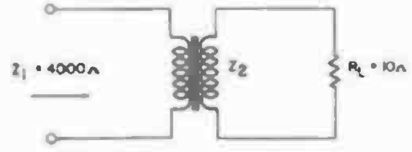
8.26 What is an impedance-matching coil?—A coil similar to a repeat coil, except it has more taps and a wider range of impedances. A typical impedance-matching coil may have a 600-ohm winding with taps at 500, 250, and 150 ohms; the other coil may be a 5000-ohm winding with taps at 1000, 500, 250, 150, 50, or 30 ohms. Such coils are not always balanced in their design and should be carefully checked for unbalance when connected in a circuit subject to leakage.

8.27 What causes the insertion loss in a transformer?—It is the loss caused by the shunt reactance, distributed capacitance, dc resistance of the coils, core material, and other losses due to the geometry of the coils. In the average audio transformer, this loss is not too important unless the power in the transformer is large, such as in an output transformer.

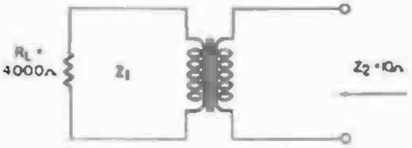
8.28 What are the basic principles of impedance matching?—The formula for impedance matching states: The impedance ratio of a transformer is directly proportional to the square of the turns. Frequency does not enter into this formula.

8.29 What is meant by the term impedance ratio?—Transformers are generally used to step up or step down a voltage or to transfer power from one circuit to an her. Because circuits differ in their in edance, if the maximum transfer of power is to take place, the turns ratio of the transformer must be such that an impedance match is effected between the two circuits, the source and the load. As an example, assume the plate circuit of an amplifier is to be connected to a loudspeaker.

As the plate resistance of a tube may run into several thousand ohms and the loudspeaker only a few ohms, the transformer will have to be of such a turns ratio that the impedance reflected back by the loudspeaker will cause the



(a) Step-up transformer.



(b) Step-down transformer.

Fig. 8-30. Reflected impedance by termination of a transformer winding.

primary of the transformer to present the correct load impedance to the plate of the tube. If the tube requires a load impedance of 5000 ohms and the speaker 16 ohms, the impedance ratio is 5000/16 or 312.

For a transformer of this design, if a 16-ohm resistance is connected across the output winding, the reflected impedance seen by the tube plate circuit will be 5000 ohms. The turns ratio of the coils is 17.66 or the square root of 312. Thus, it may be said:

$$\frac{Z_s}{Z_p} = \left(\frac{N_s}{N_p}\right)^2 \quad \text{or} \quad \frac{N_s}{N_p} = \sqrt{\frac{Z_s}{Z_p}}$$

where,

Z_s is the impedance of the secondary,
Z_p is the impedance of the primary,
N_s is the number of turns in the secondary winding,
N_p is the number of turns in the primary winding.

The foregoing relationship is generally stated: The impedance ratio is equal to the turns ratio squared, or the turns ratio is equal to the square root of the impedance ratio. The voltage ratio is equal to the turns ratio.

In a well-designed transformer the impedance ratio will remain the same throughout the normal frequency band. If a pure capacitance or indnctance is connected across the transformer rather than a resistance, the impedance seen by the primary will appear as a capacitance or inductance. Very often advantage is taken of this fact to transform small capacitors to a larger value for starting capacitor-start motors. (See Question 8.34, and Question 3.9.)

Because of the practical limits of a transformer, a given transformer cannot be used over a wide range of impedances. Attempting to use a transformer to obtain a given impedance ratio, but with the impedance diverging considerably from the original design, will result in frequency discrimination and a loss of power transfer.

8.30 What is reflected impedance?—The impedance seen when looking into a given winding of a transformer with one or more of the other windings terminated in a given load resistance.

To illustrate how the reflected impedance of a transformer winding is calculated, assume an ideal transformer (one with no losses) is available with a turns ratio of 20:1. The secondary winding (the smaller of the two) is terminated with a resistive load of 10 ohms as shown in part (a) of Fig. 8-30. The reflected impedance to the primary winding may be calculated:

$$Z_1 = N^2R_L \text{ or } Z_1 = N^2Z_2$$

where,

N is the turns ratio,
R_L the load resistance in ohms across the secondary,
Z_1 the primary impedance,
Z_2 the secondary impedance.

The impedance seen by the generator is the turns ratio (20) squared, or 400 times the impedance (10 ohms) of the secondary. The 4000 ohms seen at the primary winding is the "reflected impedance" caused by the 10-ohm secondary terminating resistance. The 4000 ohms will only be seen when the secondary is terminated in 10 ohms. If the primary impedance is measured with the secondary unterminated, only the actual inductive reactance of the primary coil will be measured. The reflected impedance of a step-up transformer of known turns ratio is calculated using the foregoing equation. For a step-down transformer, Fig. 8-30 part (b), the equation is:

$$Z_2 = \frac{Z_1}{N^2}$$

where,

Z_2 is the smaller impedance of the two windings,
Z_1 is the larger impedance of the two windings,
N^2 is the impedance ratio (turns ratio squared).

It should be understood that a transformer does not supply an impedance, but is designed to effect an impedance match between two circuits by supplying the correct turns ratio between the source and load impedances. Transformers will only present the correct impedance match when they are properly terminated. Transformers may be supplied with a solid termination on either winding to deflect a given impedance to the other winding. This method of operation is used with amplifiers employing beam-power tubes in the output stage. The plate resistance of such tubes is generally of the order of 25,000 to 90,000 ohms and is worked into a load impedance of 3000 to 10,000 ohms to reduce distortion. If the transformer is of the correct turns ratio, the correct load impedance will be reflected by the

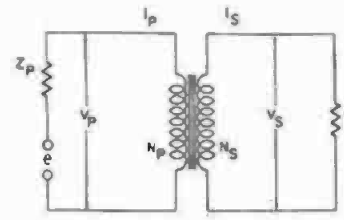
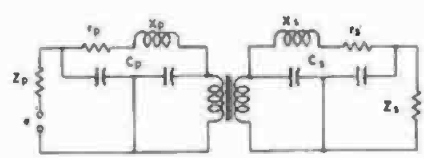


Fig. 8-31. The electrical circuit of an ideal transformer of 100% efficiency.

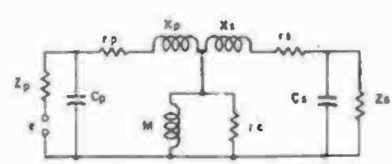
secondary winding load resistance. If the transformer has more than one winding, only one must be terminated, unless the transformer is designed to operate with a double termination. In the practical transformer, the distributed capacity and leakage are also reflected to the primary side. (See Questions 8.31, 8.32, and 8.33.)

8.31 ***What is an ideal transformer?***—A transformer which has no electrical losses. None of the electrical energy is lost in producing the magnetic field in the iron core, nor as heat in the windings. Furthermore, all the lines of flux of the changing current in one winding link all the turns of the other winding. (See Fig. 8-31.) Actually, this is not possible, as there are always some losses regardless of the design and materials used.

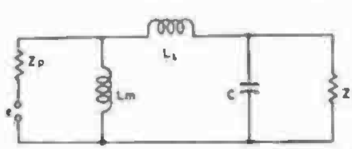
8.32 ***What are the losses of a practical transformer?***—A simple trans-



(a) Physical transformer.

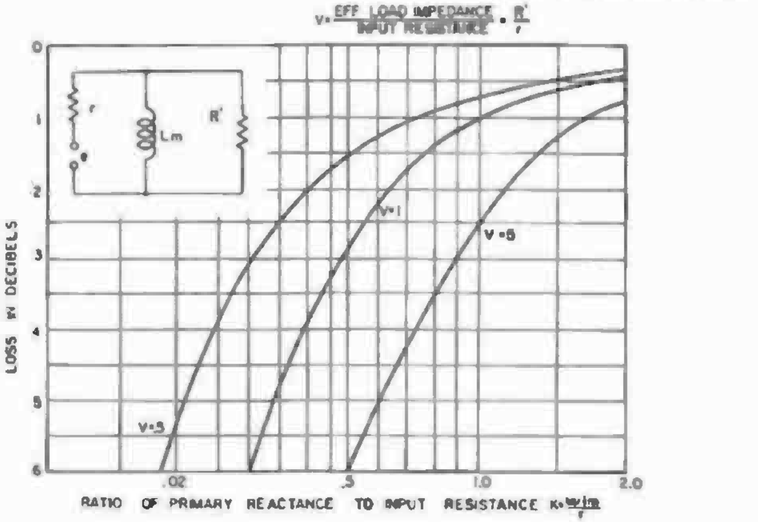


(b) Equivalent "T" network of (a).

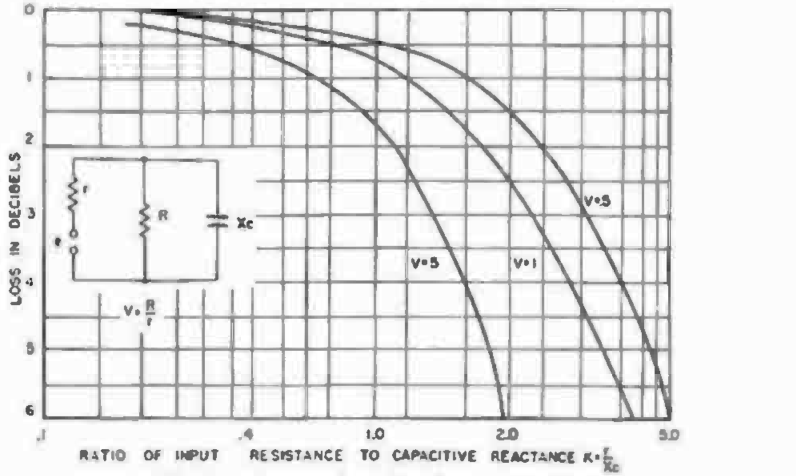


(c) Simplified equivalent circuit of (b).

Fig. 8-32A. Equivalent circuits of an audio transformer.



(a) Loss of power due to shunt effect of the primary.



(b) Loss due to distributed capacity.

Fig. 8-32B. Loss curves of an audio transformer.

former with its various perimeters is shown at parts (a) and (b) of Fig. 8-32A. In addition to the power transfer characteristics, the primary inductance L_m shunts the source impedance, as shown at part (c) of Fig. 8-32A. A leakage reactance L_L which in effect is in series with the load, and a distributed capacity C (due to the winding of the coils) are also in shunt with the source impedance.

As the frequency of the input signal decreases, the primary inductance also decreases. This decrease of inductance may reach a point where some of the power which is normally transferred from the primary to the secondary is shunted through this inductance, as shown at part (a) of Fig. 8-32B. If the frequency of the signal voltage is increased, the impedance of the distributed capacity decreases and shunts a portion of the power which would normally be passed to the load, as shown at part (b) of Fig. 8-32B.

8.33 *What Is leakage reactance?*—The transfer of electrical energy from the primary to the secondary of a transformer depends on the magnetic lines of force linking both the primary and secondary windings. In an ideal transformer, 100 percent of the lines of force would be utilized and the maximum transfer of energy would take place without loss.

The foregoing situation can never be achieved, as a certain amount of the magnetic lines of force leak off into the air. This leakage is called leakage reactance. Leakage reactance is reduced in a transformer by the core material, core design, and the geometry of the coils. (See Question 8.32.)

In conventional push-pull output transformers, the magnetic lines of force do not completely couple both halves of the primary winding. Lines of force coupled to one coil but not to the other cause a counter emf to be generated.

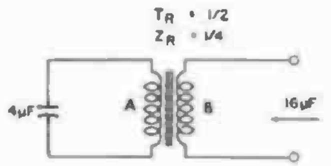


Fig. 8-34. Transformation of a capacitor.

Leakage reactance may be considered to be like an inductance connected in series with the plate of each tube, and is the result of insufficient magnetic coupling between the two halves of the primary winding. When the tubes are driven to cutoff at frequencies above 3000 Hz, the output waveform becomes distorted because of sudden changes in the plate current caused by transients. Output transformers of special design, such as the McIntosh, have been developed to reduce the effects of leakage reactance to where it becomes negligible. The McIntosh amplifier is discussed in Question 12.231.

8.34 Is it possible to transform capacitance?—Yes, this is often done in capacitor-start motor circuits. Referring to Fig. 8-34, a 4-μF capacitor is shown connected across the primary winding of a transformer having an impedance ratio of 1:4. Looking into the secondary winding, an image capacity of 16 μF is seen. (See Question 8.29.)

8.35 What are reflection losses?—The transmission losses resulting from a portion of the transmitted signal being reflected back towards the signal-source. This may be the result of mismatched impedances or of a discontinuity in the transmission line. In the case of an improperly terminated line, not all of the signal energy is transmitted to the load. The portion not absorbed by the load is reflected back towards the source and meets the energy from the source, and causes a loss of power. The greater the mismatch of the termination at the load end of the line, the greater will be the losses because of reflection.

8.36 How may the turns ratio of an unknown transformer be determined?—A known voltage is applied to one of the windings and the voltage is measured across the other winding. The turns ratio is equal to the voltage ratio.

8.37 What is the relationship of the impedance to the turns ratio of a transformer?—The impedance ratio is equal to the turns ratio squared.

8.38 What is the relationship of voltage to the turns ratio of a transformer?—The voltage ratio is equal to the turns ratio.

8.39 What is the relationship of turns ratio to the impedance ratio of a transformer?—The turns ratio is equal to the square root of the impedance ratio.

8.40 Does the mismatch of impedances affect the frequency response of a transformer?—A slight mismatch has little or no effect; however, it is best to measure the frequency response if more than a 15 percent mismatch is known to exist.

8.41 What is the maximum mismatch of impedances that may be tolerated?—Although circuit requirements will control the amount of mismatch that may be tolerated, the mismatch should not exceed 15 percent.

8.42 What is the purpose of a center tap, balanced-to-ground audio transformer?—To reduce the effects of longitudinal currents and also reduce the influence of external magnetic fields and line noises. In Fig. 8-42 is shown a balanced line using two center-tapped transformers balanced to ground. Longitudinal currents are the result of inductive interference induced into both sides of the line simultaneously, as shown by the arrows. The induced currents travel in the same direction in each side of the line. Arriving at the receiving end of the line, the current travels downward to the ground through the center tap of the transformer. As the field reverses, the same action occurs but in the opposite direction. During this action, the audio signals are traveling in opposite directions to their normal manner, as indicated by the solid arrows parallel with the lines. In addition to the center tap on the transformer, electrostatic shields, as

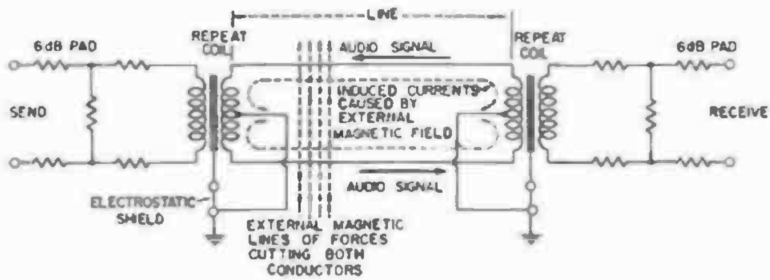


Fig. 8-42. Balanced transmission line using two balanced-to-ground repeat coils, one at each end of the line to reduce the effect of longitudinal currents caused by external magnetic fields.

explained in Question 8.12, are used to prevent capacity coupling between windings, thus, permitting only inductive coupling.

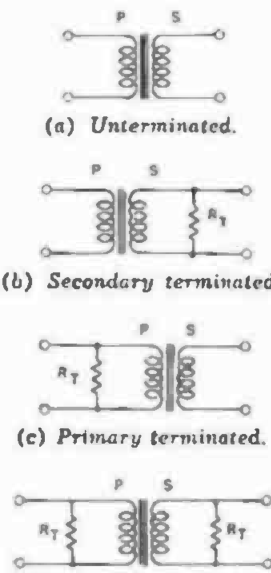
8.43 What is a terminated transformer?—A transformer operating with a resistive load connected across the primary or secondary winding in parallel with the normal load of the winding. Fig. 8-43 shows unterminated and terminated transformers operating with different types of terminators.

8.44 Where are unterminated transformers used?—To feed the grid of a vacuum tube. This connection is often referred to as an open-circuit connection and is generally employed in microphone and photocell preamplifiers. The source feeding the transformer is considered to be the transformer termination. An unterminated transformer should not be used to feed an equalizer, filter, or an amplifier requiring a solid termination.

8.45 Where are terminated transformers used?—In circuits requiring a solid termination such as a line, equalizer, filter, or similar devices.

8.46 What are the characteristics of an unterminated transformer?—The frequency characteristics of an unterminated transformer differ somewhat from those of a terminated transformer and are as follows: Assume a high quality line-to-grid transformer designed to operate from a source impedance of 500 ohms and having a secondary impedance of 50,000 ohms is connected to the grid of a vacuum tube operating as a class-A amplifier. The secondary will be left unterminated.

Under normal operating conditions, when the primary is terminated by a source impedance of 500 ohms, it will be assumed the frequency response



(a) Unterminated.

(b) Secondary terminated.

(c) Primary terminated.

(d) Primary and secondary terminated.

Fig. 8-43. Transformer terminations.

shows a loss of 1 dB at 40 Hz and 12,000 Hz. If the primary is now terminated by a line impedance of 100 ohms rather than 500 ohms, the loss of 1 dB will appear at $\frac{1}{5}$ ($\frac{100}{500}$) of 40 Hz or at 8 Hz. Under these conditions, the high end may resonate at some frequency around 12,000 Hz.

If the primary termination is now changed to 1000 ohms, the frequency response will show a loss of 1 dB at 80 Hz, or the lowest frequency multiplied by the new impedance ratio (1000/500). The high end will start to drop around 7000 Hz.

These values are only approximate but are generally true of unterminated transformers.

8.47 What are the characteristics of a terminated transformer?—Terminated transformers operate with resistive terminations across the primary or secondary, or both, as shown in parts (b) to (d) of Fig. 8-43. If the transformer described in Question 8.46 (500/50,000 ohms) is terminated on the secondary side with 50,000 ohms and connected to the grid of a class-A amplifier, looking into the primary will show an impedance of 500 ohms. The frequency response will show a loss of 1 dB at 80 Hz and 7000 Hz. If the transformer is terminated by a 1000-ohm source impedance, the low-frequency response will still show a loss of 1 dB at the low end, but the high-frequency response will be better than normal, or a loss of 1 dB at around 15,000 Hz. Before operating a transformer either terminated or unterminated, its characteristic should be checked by referring to the manufacturer's specifications. In any case, measurements should be made with the transformer terminated and unterminated to determine whether or not the frequency characteristics of the transformer are affected either by the termination or the method of operation.

8.48 What is a split-terminated transformer? — A transformer terminated on both the primary and secondary sides as shown in part (d) of Fig. 8-43. To illustrate how such terminations are used, assume an input transformer having an impedance ratio of 600/50,000 ohms is terminated on the secondary side with 100,000 ohms or double the secondary impedance.

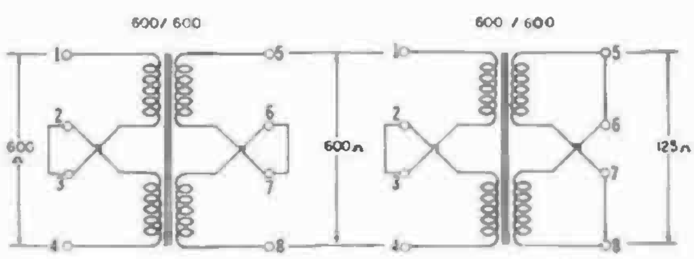
Looking into the normal 600-ohm primary shows an impedance of 1200 ohms. Now, if a 1200-ohm resistance is connected across the primary, the impedance looking into the primary is again 600 ohms. The advantage of such termination is that the transformer primary presents a solid termination to the source impedance. This type termination is of particular value if the primary is being fed from the output of another amplifier. When a transformer using split termination is first connected into a circuit, frequency-response measurements should be made to determine whether the frequency characteristics have materially changed from the original specifications. (See Question 6.113.)

8.49 How are the coils of a four-coil 600-ohm repeat coil connected externally?—A typical four-winding repeat coil is shown at part (a) of Fig. 8-49. If the terminals numbered 2 and 3 of the left-hand coils are connected in series, and terminals 6 and 7 of the right-hand coils are connected in series, the coil will have an impedance ratio of 1:1 or 600/600 ohms.

Leaving terminals 2 and 3 connected in series and connecting terminal 5 to 6 and terminal 7 to 8, as shown at part (b) of Fig. 8-49, the impedance ratio is changed to 600/150 ohms. Leaving the left-hand coils connected in series, either of the right-hand coils alone will present an impedance of 150 ohms. As the four coils are of similar design, any combination may be used by parallel or series connection. It will be observed the internal connections of the coils are crossed over inside the case, and that the ends of a given coil are either odd or even numbered. This is standard procedure for numbering coil terminals.

8.50 What is a nested shield?—An audio-transformer case consisting of alternate iron and copper shields separated by an air space, as shown in Fig. 8-50.

8.51 What is the theory of a nested shield?—Referring to Fig. 8-51, when a magnetic field encounters the outside high permeability iron case (A) of the transformer, it will travel through the case and be deflected around the trans-



(a) 600/600 ohms. *(b) 600/150 ohms.*

Fig. 8-49. Internal connections for a four-coil repeat coil.

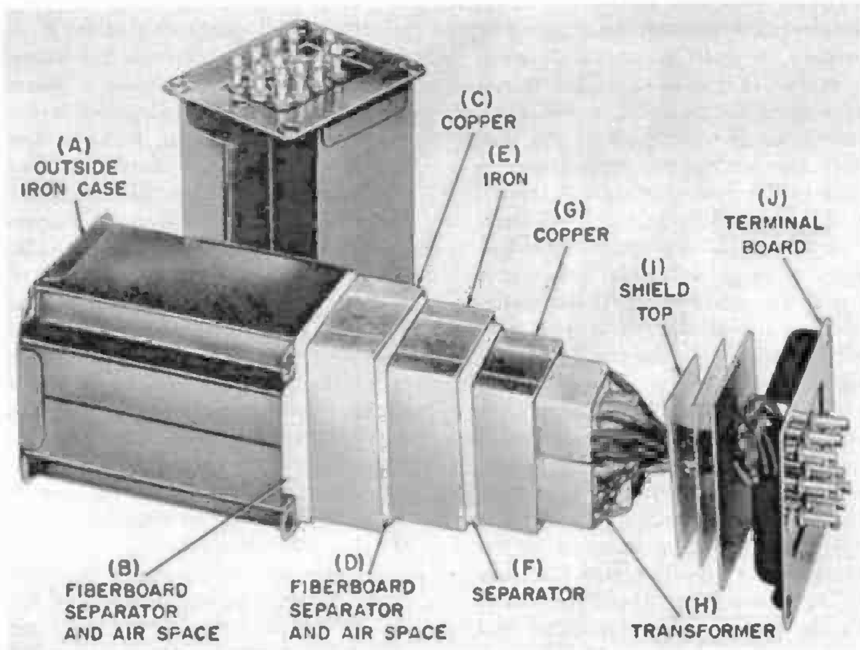


Fig. 8-50. Transformer encased in nested-shield. (Courtesy, Kenyon Electronics Inc.)

former (inside the case) because iron offers a path of lower resistance to magnetic lines of force than air. Also, in passing through the iron case a portion of the field is dissipated in the form of heat due to friction in passing through the iron and the short-circuiting effect of the case. The now reduced field continues through the air space filled by the fiberboard spacers (B), then through the copper shield (C). The field again passes through an air space (D) to a high-permeability iron shield (E) which again reduces the field strength. Leaving this shield, the field passes through air space (F) to copper shield (G) which houses the transformer (H). At the top of each shield are covers (I) (shown in Fig. 8-50) which, in effect, provide watertight shields around the transformer. The copper shields have little or no effect on the reduction of the magnetic field. Their purpose is to reduce the effects of electrostatic fields and disturbances caused by breaking of electrical circuits and radio transmissions.

To secure a further reduction from the effects of external magnetic fields, the coils of the transformer are wound hum-bucking, which neutralizes the effects of any magnetic field that may reach the transformer coils after passing through the multiple shielding.

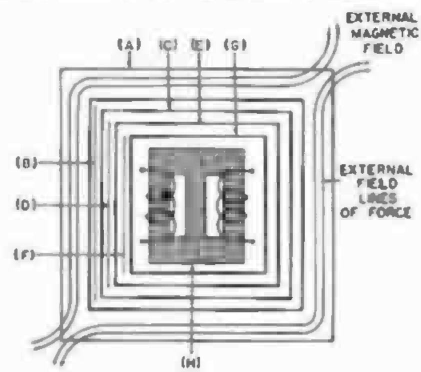


Fig. 8-51. External magnetic lines of force deflected by an iron case and a nested-shield around a transformer. Symbols correspond to those of Fig. 8-50.

Using the construction illustrated in Figs. 8-50 and 8-51 hum and noise pickup may be reduced 120 dB relative to the strength of the interfering magnetic field.

The core of the transformer is brought to a separate terminals for connection to a transmission ground system as described in Question 24.33. The outer iron case is completely insulated from the internal shields and transformer core.

8.52 ***Define the term 120 dB magnetic shielding.***—It means for a given external magnetic field surrounding the transformer, the strength of the magnetic field will be reduced 120 dB at the transformer windings. The amount of reduction is fixed and the actual hum pickup can vary with the change in the external magnetic field strength. However, with 120 dB of hum reduction, generally little difficulty is encountered.

Transformer hum-bucking windings are discussed in Question 8.98, and the measurement of shielding effectiveness is discussed in Question 23.162.

8.53 ***How much is an external magnetic field reduced by a single cast-iron case 1/8-inch thick?***—Approximately 12 dB.

8.54 ***How is the effectiveness of a magnetic field measured?***—By the method given in Question 23.162.

8.55 ***What type transformers require nested shielding?***—Low-level input transformers operating ahead of a considerable amount of gain, microphones, line, repeat coils and others of similar design, or any low-level transformer operating near ac operated equipment.

8.56 ***Should an output transformer be shielded?***—Yes, but an output transformer does not require as heavy a shield as does an input or interstage transformer. As a rule, an output transformer only requires 6 to 10 dB of shielding because it operates at a fairly high level and is not generally affected by extraneous magnetic fields. In many instances, the output transformer is only cased in light metal.

8.57 ***Should an interstage transformer be shielded?***—Yes, about 12 to 40 dB. It is a matter of circuit design and operating levels.

8.58 ***What is an input transformer?***—An audio transformer used at the input of an amplifier or similar device. Input transformers are generally better shielded from magnetic fields, as they are required to operate in low-level circuits. Input transformers, as a rule, are designed to match the low impedance of a transmission line to the grid of a tube. Typical impedance ratios are: 250 or 600 ohms to a 50,000-ohm secondary. (See Question 8.52.)

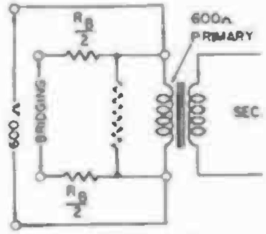
8.59 ***What is an interstage transformer?***—A transformer designed to couple the plate circuit of one vacuum tube to the grid circuit of another. It may be designed to couple either single-ended or push-pull stages.

8.60 ***What is an output transformer?***—A transformer used to transfer the power of an output stage to an external circuit. The primary is of higher impedance than the secondary because it must couple the high impedance of the tube plate circuits to a lower impedance. Output transformers have but one purpose—to transfer the power developed in the plate circuit by the vacuum tubes to an external circuit, without frequency discrimination or excessive power losses. A well-designed output transformer will have similar frequency characteristics, within plus or minus 1 dB, at both high- and low-level power outputs. The insertion loss must be negligible. This is particularly true if a considerable amount of power is to be transmitted.

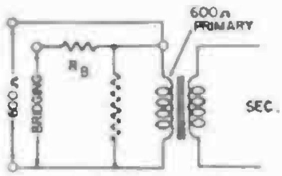
8.61 ***What is a modulation transformer?***—A transformer used for coupling a modulator tube in a radio transmitter to the tubes to be modulated. Such transformers have numerous taps to secure a proper impedance match between the two circuits.

8.62 ***What is a bridging transformer?***—A special transformer with a high-impedance primary designed to be operated from a low source impedance. Many amplifiers used in recording systems and broadcasting make use of bridging transformers, as these transformers permit several amplifiers to be connected across a line without upsetting the impedance match of other equipment. Bridging amplifiers are described in Question 12.87.

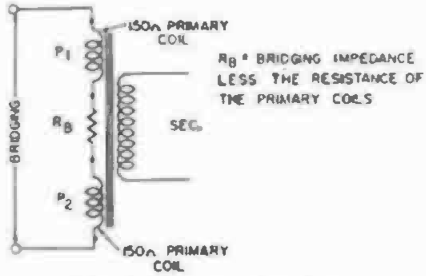
8.63 ***Can a transformer with a high-impedance primary be used for bridging purposes?***—No, uot as a rule. Bridging transformers are of special design. However, a transformer with a low-impedance primary may often be converted for bridging use by the addition of resistors as shown in Fig. 8-63. In part (a) of Fig. 8-63 two resistors R_B have been added in series with a 600-ohm primary, to raise the input impedance to the desired bridging impedance. In part (b) of Fig. 8-63 the same method has been used, except the circuit is unbalanced. In part (c) of Fig. 8-63, resistor R_B has been connected in series witb the return of a split-primary transformer. In this case the transformer may only be used for bridging purposes. Transformers converted to bridging inputs, as shown, will



(a) *Balanced.*



(b) *Unbalanced.*



(c) *Center-tapped primary.*

Fig. 8-63. Methods of converting low-impedance transformers to bridging inputs.

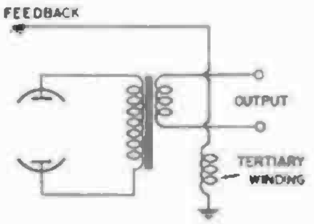
show a considerable insertion loss because of the resistance R_B in series with the windings. This loss will vary depending on the impedance ratio of the transformer and the value of the resistors. Transformers designed expressly for bridging service do not as a rule make use of resistors.

It may be necessary when using a transformer with a 600-ohm primary for bridging purposes to add an additional resistor in shunt with the primary winding as shown by the dotted line resistor in parts (a) and (b) of Fig. 8-63, to preserve the frequency response of the coil. The addition of this shunt resistor will increase the insertion loss of the transformer as a whole, the amount depending on the impedance ratio between the bridging source and the impedance of the transformer primary.

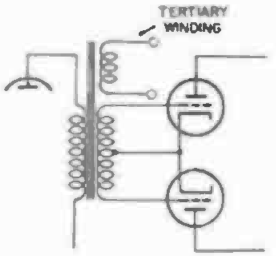
The three resistors, in the case of parts (a) and (b), form a bridging pad. The loss may be computed as described in Question 5.54.

8.64 ***What are the standard bridging input impedances used in the sound recording industry?***—7500, 10,000, 25,000, and 30,000 ohms.

8.65 ***What is a tertiary winding?***—It is a third winding on a transformer. In an output transformer it is generally used for obtaining a negative feedback



(a) *Negative feedback.*



(b) *Interstage transformer.*

Fig. 8-65. Tertiary windings.

voltage. In an interstage transformer, it is used to feed auxiliary equipment, as shown in Fig. 8-65.

8.66 ***What is a hybrid coil?***—A coil consisting of two or more windings, and used in telephone circuits, sound mixers, and test equipment. The geometrics of the coil design are such that the signal can only be transmitted in a given direction when the coil is properly terminated. A typical hybrid coil is shown in Fig. 8-66A. It will be noted the signal can only be transmitted between Z_1 and Z_4, and Z_2 and Z_4. When Z_1, Z_2, and Z_4 are terminated in their proper load impedances and two signals are applied simultaneously to Z_1 and Z_2, the signals do not react on each other and can only be transmitted in the direction of Z_4.

If a voltage is applied to Z_1 and Z_3 (plus-minus to C) by induction, a voltage is developed across Z_2. When Z_3 (a resistance) is adjusted to its correct value, the voltage developed across Z_3 will be *equal* to that developed *across* Z_2, *but of opposite polarity*. Since these two voltages are in *opposition*, the volt-

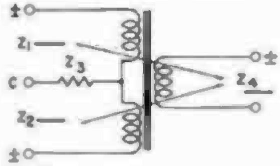


Fig. 8-66A. Conventional hybrid coil.

age applied to Z_1 will be *cancelled* and will not appear across Z_2. Normal transformer action occurs between Z_1 and Z_4, therefore the voltage applied to Z_1 appears across Z_4; the same holds true for a voltage applied to Z_2.

The electrical balance and isolation in a hybrid coil changes somewhat with frequency, the greatest unbalance occurring above 5000 Hz. This causes low-level mixing of the two signals applied to Z_1 and Z_2, and results in *low-level intermodulation*. As a rule, a well designed hybrid coil will indicate a balanced condition of 45 dB, or better, from 30 to 10,000 Hz. (See Ques. 23.72.)

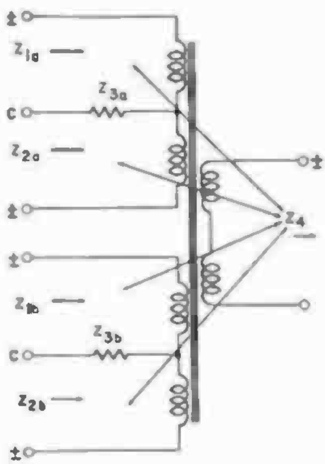


Fig. 8-66C. Sound-mixer hybrid coil.

If a signal voltage is applied to Z_4, it will be transmitted to both Z_1 and Z_2, but *not* to Z_3. The load terminators for Z_1, Z_2, and Z_4 are rather critical, as is the adjustment of the Z_3 resistor.

In Fig. 8-66B is shown how hybrid coils are employed in the telephone industry to permit a single line to be used for two-way conversation. It is the practice to use two amplifiers to make up a repeater station, one amplifier providing amplification in one direction and the second in the opposite direction. When the lines are properly terminated and the balancing networks adjusted, both amplifiers may be used without producing a howl or singing.

The sum of the gains of the two amplifiers must always be less than the sum of the losses across the two hybrid coils to prevent unbalance and singing. For telephone transmission the frequency bandwidth is limited to frequencies between 200 and 3000 Hz. Filters are employed to prevent singing at frequencies outside this band. The balancing networks consist of a resistance and a capacitor connected in series across one pair of terminals of the hybrid coil. In some instances an inductance is employed.

The action of a hybrid coil is somewhat similar to a Wheatstone bridge circuit as described in Question 22.21.

The hybrid coil shown in Fig. 8-66C is typical of such a coil used in sound mixer networks designed for rerecording purposes.

Inputs Z_{1a} to Z_{2b} are connected to the mixer controls. Winding Z_4 feeds equipment following the coil. Resistors Z_{3a} and Z_{3b} are the same as for the coil shown in Fig. 8-66A.

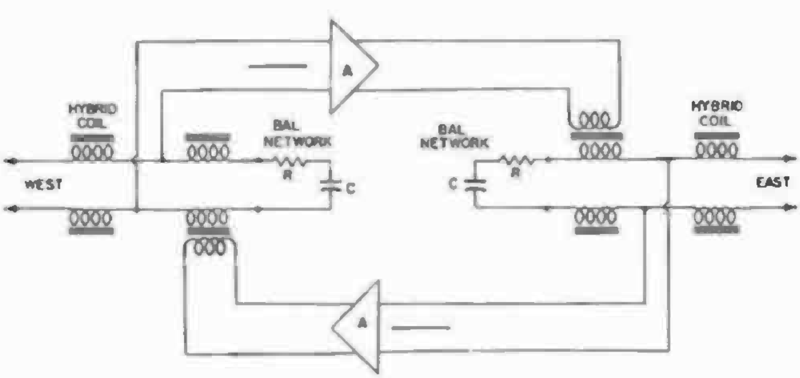


Fig. 8-66B. Hybrid coils used in a telephone-repeater station to permit using a single pair transmission line for East-West conversation.

Sound mixers using hybrid coil design are discussed in Question 9.23.

8.67 ***What are the transmission losses between the windings of a conventional hybrid coil?***—Refer to Fig. 8-66A.

Z_1 to Z_4 — 3 dB	Z_2 to Z_3 — 3 dB
Z_1 to Z_3 — 3 dB	Z_3 to Z_4 Infinite
Z_2 to Z_4 — 3 dB	Z_1 to Z_2 Infinite
Z_4 to Z_3 Infinite	Z_4 to Z_1, Z_2 — 3 dB

The losses given are ideal and will vary somewhat in a practical transformer. Losses between coils, given as infinite, in a practical transformer are approximately 45 to 60 dB. Other losses given are approximately correct.

8.68 ***What is the formula for calculating the balancing resistor Z_3 in a hybrid coil?***—For a mixer hybrid coil:

$$Z_3 = \frac{Z_1(N-1)}{2}$$

where,
- Z_1 equals the input impedance,
- N equals the total number of inputs,
- Z_3 is a noninductive balancing resistor.

8.69 ***How is the insertion loss of a hybrid coil calculated?***

$$dB = 10 \operatorname{Log}_{10} N$$

where,
- N is the total number of inputs.

8.70 ***What is the impedance ratio of each primary to the secondary in a hybrid coil?***

$$Z_r = \frac{2Z_1}{R_L}$$

where,
- Z_1 equals the input impedance,
- R_L equals the secondary load impedance.

8.71 ***What is trans-hybrid loss?***—The degree of isolation provided by the windings of a hybrid coil between the send and receive windings. For a perfectly balanced hybrid coil, this is infinite. However, in practice, this is generally not true as some leakage exists and is particularly noticeable at the high frequencies. Trans-hybrid loss is measured in decibels using a reference level of one milliwatt. The balancing of hybrid coils is discussed in Question 23.72.

8.72 ***Describe a method of measuring the "Q" of a coil.***—Using the circuit of Fig. 8-72 the "Q" of a coil may be easily measured for frequencies up to 1 MHz. Since the voltage across an inductance at resonance equals "QE" (where E is the voltage developed by the oscillator) it is necessary only to measure the output voltage from the oscillator and the voltage across the inductance.

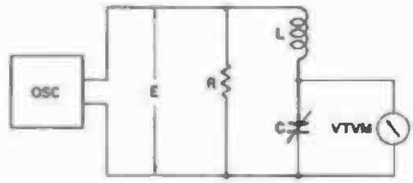


Fig. 8-72. Circuit for measuring the "Q" of a coil.

The voltage from the oscillator is introduced across a low value of resistance, R, about 1/100 of the anticipated radio-frequency resistance of the LC combination, to assure that the measurement will not be in error by more than 1 percent. For average measurements, resistor R will be on the order of 0.10 ohm. If the oscillator cannot be operated into an impedance of 0.10 ohm, a matching transformer may be employed. It is desirable to make C as large as convenient in order to minimize the ratio of the impedance looking from the vacuum-tube voltmeter to the impedance of the test circuit. The voltage across R is made small, an even value on the order of 0.10 volt. The LC circuit is then adjusted to resonate at the resultant voltage measured. The value of "Q" may then be equated:

$$Q = \frac{\text{Resonant Voltage Across C}}{\text{Voltage Across R}}$$

The "Q" of a coil may be approximated by the equation:

$$Q = \frac{2\pi fL}{R} \doteq \frac{X_L}{R}$$

where,
- F is the frequency,
- L is the inductance,
- R is the dc resistance (as measured by an ohmmeter),
- X_L is the inductive reactance of the coil.

8.73 ***What factors affect the Q of a coil?***—The principal factors are frequency, inductance, dc resistance, inductive reactance, and the type of winding. Other factors are the core losses, distributed capacity, and the permeability of the core material.

8.74 What does the term permeability mean?—It is a measure of the ease with which a magnetic material will pass lines of force. High-quality transformers, shields, and other electrical equipment use high-permeability magnetic iron made of various alloys.

8.75 What is the relationship of the turns of a coil to its inductance?—The inductance varies as the square of the turns.

8.76 What effect does a shorted turn have on the inductance of a coil?—The inductance is decreased. In an audio transformer the frequency characteristic will be affected and the insertion loss increased.

8.77 What is the effect of inserting an iron core in a coil?—The inductance is increased; hence, its inductive reactance is increased.

8.78 What is a retard coil?—A choke coil. This is a term used in the telephone industry.

8.79 What is coupled impedance?—The effect produced in the primary of a transformer by the influence of the current flowing in the secondary windings.

8.80 What is meant by the term unity coupling?—If two coils are placed so that the lines of force from one coil cut all the windings of the second coil, it is said that unity coupling exists.

8.81 Why is an air gap used in the core of a choke coil?—To prevent the saturation of the core material with an attendant loss of inductance. Generally, the gap is of the order of a few thousandths of an inch.

8.82 If an air gap is induced in an existing choke, how is the inductance affected?—The inductance will be reduced.

8.83 What is a saturable reactor?—A device similar to a transformer in construction with a third control winding connected to a source of variable direct current. Increasing or decreasing the direct current through the control winding causes a change in the inductance of the coils on the core. In this manner, control of ac devices may be had by the application of direct current to the control winding.

8.84 What is meant by the term flywheel effect?—It is the effect noted in a tank circuit (resonant circuit) which will cause the circuit to continue to oscillate after the excitation voltage has been removed. The higher the Q of the tank circuit, the longer the oscillations will persist and the longer it will take to dissipate the energy and for the circuit to come to complete rest.

8.85 What is the incremental inductance of a coil?—When an alternating current is applied to a coil in which a given amount of direct current is flowing, the inductance will be increased by the superimposition of the alternating current. The inductance will increase (within limits) as the ac component becomes larger.

8.86 Can intermodulation distortion be induced in an audio transformer by the manner in which the core laminations are stacked?—Yes. The manner in which the core laminations are stacked can have quite an effect on the intermodulation distortion induced by a transformer. By the proper design of the coils, the method of stacking, and the use of grain-oriented steel, the intermodulation distortion can be reduced to a negligible amount.

8.87 How may an interstage transformer without a center tap be used for phase-splitting?—In the manner illustrated in Fig. 8-87. Two resistors R_1 and R_2 are connected across the secondary winding to provide the missing center tap. The terminating resistors are selected on the basis of the reflected impedance to the plate of the preceding tube, and the maximum value of resistance permissible for the grid circuit of the particular tube in use. Thus, if a transformer designed for an impedance ratio of 15,000 to 86,000 ohms is to be center-tapped, resistors should be selected which, when connected across the secondary, will not greatly affect the

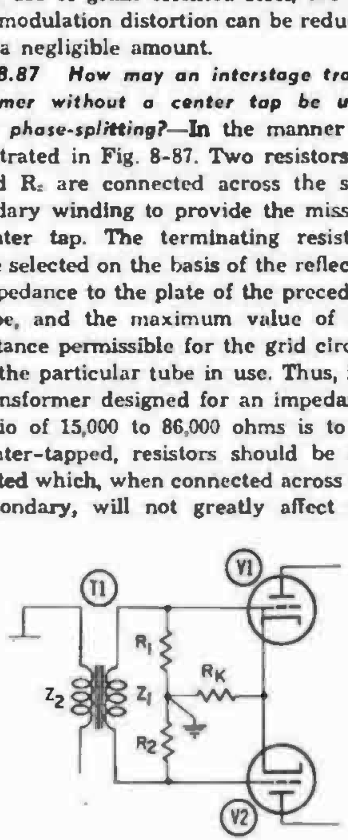


Fig. 8-87. Interstage transformer with resistors connected across the secondary to supply a center tap for push-pull operation.

Z	2	3	4	5	6	7	8	9	10	15	25
2	—	.097	.330	.670	1.070	1.520	2.000	2.520	3.080	6.00	12.900
3	.097	—	.064	.250	.500	.840	1.200	1.600	2.000	5.00	10.600
4	.330	.064	—	.058	.190	.410	.660	1.000	1.340	3.60	9.000
5	.670	.250	.058	—	.050	.180	.340	.580	.840	3.00	7.700
6	1.070	.500	.190	.050	—	.049	.120	.290	.510	2.00	6.600
7	1.520	.840	.410	.180	.049	—	.084	.140	.280	1.40	5.500
8	2.000	1.200	.660	.340	.120	.034	—	.029	.110	1.08	4.600
9	2.520	1.600	1.000	.580	.290	.140	.029	—	.023	.81	3.900
10	3.080	2.000	1.340	.840	.510	.280	.110	.023	—	.49	3.400
15	6.000	5.000	3.600	3.000	2.000	1.400	1.080	.810	.490	—	1.210
25	12.900	10.600	9.000	7.700	6.600	5.500	4.600	3.900	3.400	1.21	—
30	16.500	14.000	12.100	10.500	9.180	8.080	7.020	6.150	5.380	2.60	.230
33	18.500	16.000	13.700	12.300	10.200	9.000	8.400	7.000	6.000	3.20	.500
50	32.000	29.000	25.000	23.000	22.000	20.000	18.000	16.000	15.000	10.00	4.200
64	44.000	40.000	36.000	34.000	30.000	28.000	27.000	25.000	23.000	16.80	9.000
100	73.000	68.000	64.000	61.000	57.000	54.000	51.000	48.000	47.000	37.20	25.000
125	95.000	88.000	85.000	80.000	77.000	72.000	70.000	68.000	63.000	53.30	39.000
150	117.000	116.000	104.000	100.000	94.000	90.000	88.000	84.000	81.000	69.00	52.000
200	161.000	154.000	146.000	142.000	135.000	130.000	128.000	123.000	119.000	104.00	83.000
250	208.000	198.000	191.000	184.000	181.000	175.000	168.000	164.000	160.000	142.00	116.000
333	282.000	272.000	262.000	256.000	247.000	240.000	237.000	231.000	225.000	205.00	174.000
500	436.000	426.000	417.000	405.000	399.000	385.000	381.000	375.000	368.000	342.00	301.000
600	534.000	520.000	506.000	497.000	484.000	475.000	471.000	462.000	454.000	424.00	380.000
$\sqrt{N}$	1.41	1.73	2.00	2.24	2.45	2.65	2.83	3.00	3.16	3.87	5.00

Fig. 8-88B. Impedance between

15,000-ohm impedance of the primary winding. As an example, selecting two 246,000-ohm resistors and connecting them across the secondary winding will reflect to the primary an impedance of approximately 15,00 ohms.

The reflected impedance may be calculated as follows:

$$Z_p = \frac{R_1 + R_2}{Z_R^2}$$

Z→ ↓	30	33	50	64	100	125	150	200	250	333	500	600
2	16.500	18.50	32.0	44.0	74.0	95.0	117.0	161.0	208.0	282.0	436.0	534.0
3	14.000	16.00	29.0	40.0	68.0	88.0	116.0	154.0	198.0	272.0	426.0	520.0
4	12.100	13.70	25.0	36.0	64.0	85.0	104.0	146.0	191.0	262.0	417.0	506.0
5	10.500	12.30	23.0	34.0	61.0	80.0	100.0	142.0	184.0	256.0	405.0	497.0
6	9.180	10.20	22.0	30.0	57.0	77.0	94.0	135.0	181.0	247.0	399.0	484.0
7	8.800	9.00	20.0	28.0	54.0	72.0	90.0	130.0	175.0	240.0	385.0	475.0
8	7.020	8.40	18.0	27.0	51.0	70.0	88.0	128.0	168.0	237.0	381.0	471.0
9	6.500	7.00	16.0	25.0	48.0	68.0	84.0	123.0	164.0	231.0	375.0	462.0
10	5.380	6.00	15.0	23.0	47.0	63.0	81.0	119.0	160.0	225.0	368.0	454.0
15	2.600	3.24	9.6	16.8	37.2	53.3	68.9	104.0	142.0	205.0	342.0	424.0
25	.230	.50	4.2	9.0	25.0	39.0	52.0	83.0	116.0	174.0	301.0	380.0
30	—	.67	2.5	6.3	20.4	32.5	45.8	75.0	106.7	164.0	290.0	361.0
33	.676	—	1.7	5.3	18.5	30.0	42.0	71.0	102.0	156.0	282.0	353.0
50	2.500	1.70	—	1.0	8.4	17.0	27.0	50.0	77.0	125.0	233.0	306.0
64	6.350	5.30	1.0	—	4.0	10.2	17.6	37.0	61.0	104.0	207.0	272.0
100	20.400	18.50	8.4	4.0	—	1.4	4.8	17.0	33.0	67.0	154.0	210.0
125	32.500	30.00	17.0	10.2	1.4	—	1.0	8.4	21.0	50.0	125.0	177.0
150	45.800	42.00	27.0	17.6	4.8	1.0	—	3.6	13.0	36.0	104.0	150.0
200	75.000	71.00	50.0	37.0	17.0	8.4	3.6	—	2.9	16.8	69.0	108.0
250	106.700	102.00	77.0	61.0	33.0	21.0	13.0	2.9	—	5.8	42.0	76.0
333	164.000	156.00	125.0	104.0	67.0	50.0	36.0	16.8	5.8	—	17.6	40.0
500	290.000	282.00	233.0	207.0	154.0	125.0	104.0	69.0	42.0	17.6	—	4.4
600	361.000	353.00	306.0	272.0	210.0	177.0	150.0	108.0	76.0	40.0	4.4	—
$\sqrt{N}$	5.48	5.74	7.07	8.0	10.0	11.2	12.25	14.14	15.8	18.25	22.36	24.5

taps on audio transformers.

where,

R_1 and R_2 are the terminating resistors,

Z_R is the impedance ratio of the transformer,

Z_p is the reflected impedance of the primary winding.

8.88 ***What is the equation for calculating the impedance between taps of a transformer?***—The unknown impedance Z_3, between taps on a transformer (Fig. 8-88A), may be calculated as follows:

$$Z_3 = Z_2\left(\sqrt{\frac{Z_1}{Z_2}} - 1\right)^2$$

$$\text{or } Z_3 = (\sqrt{Z_1} - \sqrt{Z_2})^2$$

where,

Z_3 is the unknown impedance,
Z_2 is the lower impedance tap,
Z_1 is the higher impedance tap.

As an example; assume it is desired to know the impedance (Z_3) between the 250- and 500-ohm taps on the transformer illustrated in Fig. 8-88A.

$$\begin{aligned} Z_3 &= 250\left(\sqrt{\frac{500}{250}} - 1\right)^2 \\ &= 250\,(\sqrt{2} - 1)^2 \\ &= 250\,(1.41 - 1)^2 \\ &= 250 \times .41^2 \\ &= 250 \times .1681 \\ &= 42\Omega \end{aligned}$$

Thus, the impedance between the 250- and 500-ohm taps is 42 ohms. A chart for determining the impedance between taps of a transformer without calculation is given in Fig. 8-88B. The known taps are plotted across the top and the left side of the chart. Assume it is desired to know the impedance between a 600-ohm and a 30-ohm tap. Entering the chart at 30 ohms at the left and reading to the right to the 600-ohm column the impedance is found to be 361 ohms. The square root of various impedances from 2 to 600 ohms is given at the foot of the chart for convenience when calculating an unknown impedance.

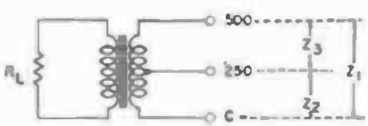


Fig. 8-88A. Transformer for example in Question 8.88.

In Fig. 8-88C is shown a typical output transformer with taps at 15, 250, 500, and 600 ohms, with the impedances between windings indicated, for ready reference.

8.89 ***How are the terminals of a transformer or a coil identified?***—Standard terminal terminology is used to identify the terminals. In Fig. 8-89A is shown the terminal markings used on a typical two-coil repeat coil or isolation transformer. An electrostatic shield is brought out to a separate terminal for connection to a transmission ground system. The core may be brought to a separate terminal or tied internally to the case, or in some instances, the core is left floating. In Fig. 8-89B is shown a coil or transformer employing a single primary center-tapped with multiple secondaries, two of which are center-tapped. The coils may be connected in series or in parallel depending on the impedance requirements.

In Fig. 8-89C is shown a coil consisting of a primary and a secondary with taps at various impedance points. In this design, the taps are unbalanced relative to ground; no center tap is provided.

Fig. 8-89D is that of a four-winding repeat coil. Each coil is identical with the others as to its turns, distributed capacity, inductance, dc resistance, and Q. The four coils are assembled in such a manner that the distributed capacity between them is at a minimum. The coils are generally connected hum-bucking. (See Question 8.98.)

A variety of impedance relations may be obtained by connecting the primaries in series or parallel and the secondaries in a similar manner. Coils of this type are generally cased in a multiple shield (see Question 8.51) to prevent pickup of noise or external magnetic fields. An electrostatic shield prevents the transfer of noise between windings while the signal currents are passed inductively.

The coil shown in Fig. 8-89E is a tapped choke or reactor with one terminal connected as a center tap.

8.90 ***How may the electrical polarity or phasing of transformer windings be established?***—By the use of the test circuit given in Fig. 8-90. Here an oscillator or a source of 60-Hz voltage is connected through a half-wave diode rectifier to the primary of the transformer to distort the waveform of the applied voltage. The oscilloscope is first connected to the primary and the attitude of the waveform is noted. The oscilloscope is then connected across each secondary winding and the attitude again noted.

If the waveform has the same attitude, it is of the same electrical polarity or in-phase with the primary wind-

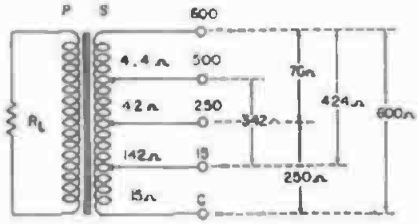


Fig. 8-88C. A typical output transformer with taps at 15, 250, 500, and 600 ohms. The in-between impedances are indicated for ready reference.

ing. If the attitude is reversed by 180 degrees, it is of opposite electrical polarity or 180 degrees *out of phase*. The in-phase terminals are marked plus-minus (±) for identification. This procedure may be used for identifying the polarity of either power or audio transformers.

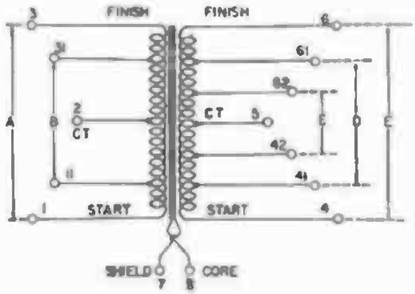


Fig. 8-89A. Configuration of two-winding coil with center tap.

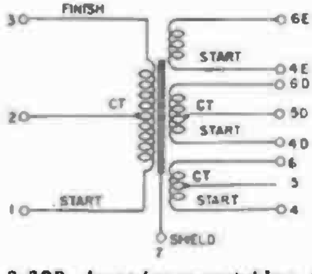


Fig. 8-89B. Impedance-matching transformer with multiple secondaries.

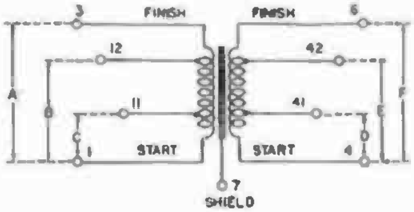


Fig. 8-89C. Configuration of a two-winding coil, no center tap.

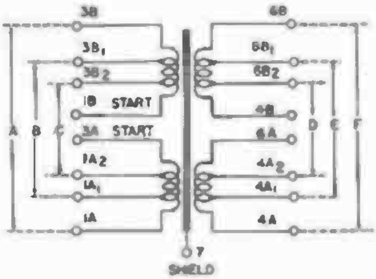


Fig. 8-89D. Impedance-matching transformer or repeat-coil terminal identification. The above coil contains four identical windings. For series connection, 3A to 1B, 6A to 4B. For parallel connection, 1A to 1B, 3A to 3B, 4A to 4B, 6A to 6B.

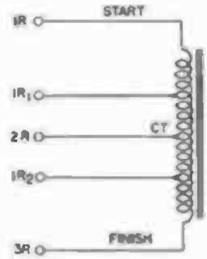


Fig. 8-89E. Center-tapped reactor.

8.91 ***What is the formula for calculating power transfer?***—A 6-volt generator with an internal impedance of 600 ohms connected to a load of 200 ohms is shown in Fig. 8-91A. Applying Ohm's law, the current is:

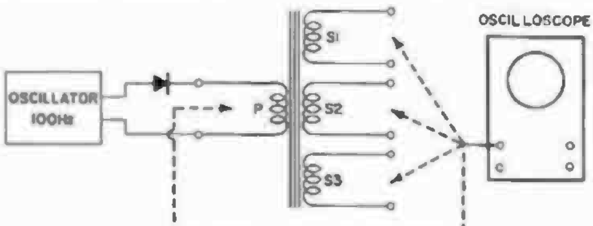


Fig. 8-90. Test circuit for identifying the electrical polarity or phasing of transformer windings.

$$I = \frac{E}{R}$$
$$= \frac{6}{600 + 200}$$
$$= 0.0075 \text{ amp}$$

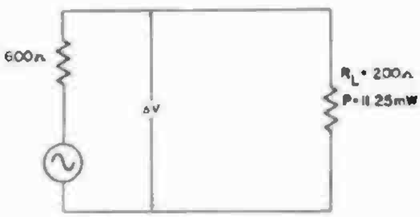


Fig. 8-91A. A 600-ohm generator working into a 200-ohm load.

The power delivered to the load may be calculated:

$$P = I^2R$$
$$= (0.0075)^2 \times 200$$
$$= 0.01125 \text{ or } 11.25 \text{ milliwatts}$$

In Fig. 8-91B the generator is connected to a load of 1000 ohms. The power delivered in this example is 14 milliwatts. In Fig. 8-91C the generator is connected to a load of 600 ohms which is equal to its internal impedance. Solving for I:

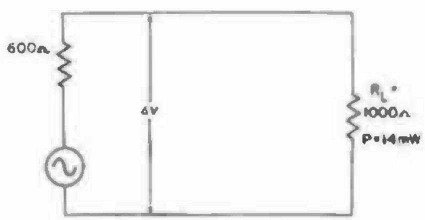


Fig. 8-91B. A 600-ohm generator working into a 1000-ohm load.

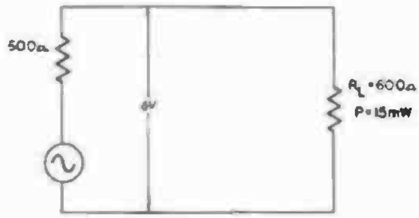


Fig. 8-91C. A 600-ohm generator working into a 600-ohm load.

$$I = \frac{E}{R}$$
$$= \frac{600}{1200}$$
$$= 0.005 \text{ amp}$$

$$P = I^2R$$
$$= (0.005)^2 \times 600$$
$$= 0.015 \text{ or } 15 \text{ milliwatts}$$

The same theory may be applied to a vacuum tube with the plate resistance of the tube acting as the generator impedance and the plate-load resistance acting as the load resistance.

Fig. 8-91D illustrates graphically the power-transfer characteristics of a generator terminated in loads varying from a lesser to a greater amount than the internal impedance of the generator.

At the point where the load resistance equals the internal impedance of the generator, a maximum transfer of power is effected.

8.92 ***What is a toroidal coil?***—This type of coil construction is explained in Questions 6.68 to 6.71.

8.93 ***What is a geoformer?***—A special type transformer designed for use with geophysical measuring equipment. The frequency range of such transformers is limited, generally to the

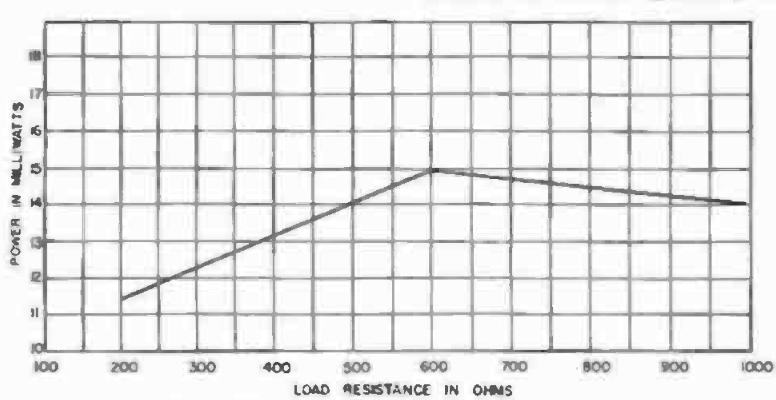


Fig. 8-91D. Power-transfer characteristics.

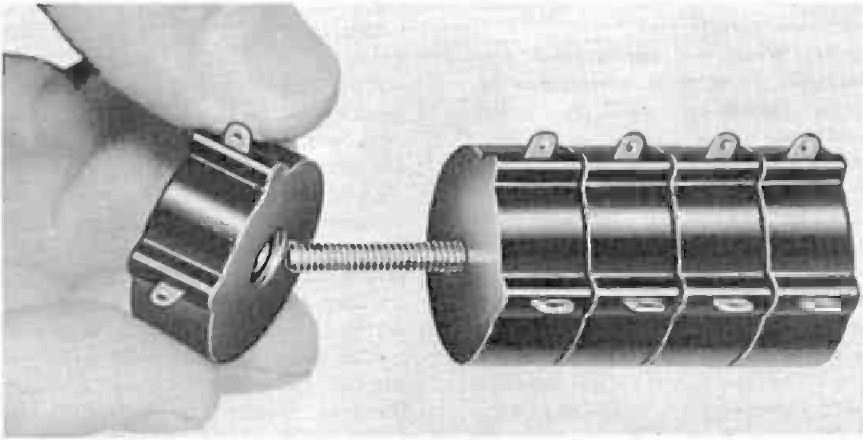


Fig. 8-94. Encapsulated toroidal coils. Because of their design, toroidal coils may be mounted one over the other without fear of intercoupling. Several encapsulated coils are shown using a single nonmagnetic machine screw ready for mounting.

range of 5 to 500 Hz, tapering off slowly after 500 Hz.

8.94 What Is an encapsulated coil?—A coil which after being wound is completely enclosed in a moulded plastic case as shown in Fig. 8-94.

8.95 What are the formulas for calculating the inductance of a single layer, spiral, and multilayer coils?—They may be calculated by the use of either Wheeler's or Nagaoka's formulas. The accuracy of the calculation will vary between one and five percent. Using Wheeler's formula for the single layer coil shown in part (a) of Fig. 8-95, the inductance may be calculated as follows:

$$L = \frac{B^2N^2}{9B + 10A} \text{ microhenries.}$$

For the multilayer coil in part (b) of Fig. 8-95,

$$L = \frac{.8B^2N^2}{6B + 9A + 10C} \text{ microhenries.}$$

For the spiral coil in part (c) of Fig. 8-95,

$$L = \frac{B^2N^2}{8B + 11C} \text{ microhenries,}$$

where,

N is the number of turns in the coil,
A is the length of the winding,
B is the radius of the winding,
C is the thickness of the winding.

8.96 What is a buck-boost transformer?—A transformer used in power-transmission work to increase or decrease the line voltage where a conventional transformer would be impractical or too costly. The connections for boosting the line voltage are shown in Fig. 8-96A and the connections for decreasing the line voltage are shown in Fig. 8-96B. It will be noted that, when reducing the line voltage, the secondary coil is connected to buck the line voltage, thus reducing it. The coils of the booster transformer must be capable of carrying the full-load current. For light loads, a filament or bell-

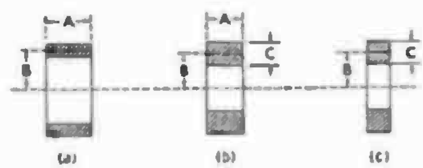


Fig. 8-95. Single- and multiple-layer inductors. (a) Single layer. (b) Multilayer. (c) Spiral.

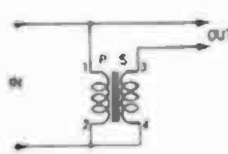


Fig. 8-96A. Booster transformer connections to increase line voltage.

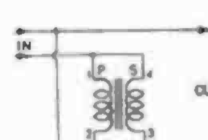


Fig. 8-96B. Booster transformer connections to buck line voltage.

ringing transformer may be used as a buck-boost transformer.

8.97 ***What are the general considerations given to a transformer regarding operating load?***—A power transformer shonld, if possible, be operated at about 90 percent of its normal rated load because the current in the secondary winding or windings will bring the voltage and the current in phase, reducing the phase angle, which results in a power factor closer to unity. The power factor of an ac circuit is the cosine of the phase angle, or the ratio of the true power to the apparent power. Therefore, it is desirable to keep the true power close to the apparent power to reduce excessive voltages and currents in the circuit.

A power transformer loaded to one-fifth its normal load-carrying capabilities can cause arcing and break down the insulation between layers of the coils. Under full-load conditions, the inductive reactance of the primary winding is almost completely cancelled by the opposing magnetizing force caused by the flow of current in the secondary windings. Thus, the phase angle between the true and apparent power is small and the power factor brought close to unity. As a rule, power transformers will show a power factor of 0.90 or more, with efficiencies of 95 percent when properly operated.

8.98 ***How is a hum-bucking audio transformer constructed?***—In modern high-gain amplifiers, the input transformer is generally required to work from a very low-level signal input. Under these conditions of high gain and a low-level signal, it is necessary to make use of specially designed input transformers which will not be affected by extraneous magnetic fields and strong electrostatic fields caused by switching circuits and radio transmissions.

Hum pickup may be reduced to a great extent by the design of the coils, core, and general construction of the transformer.

Two identical coils are wound on opposite legs of the transformer core. The direction of one of the coils is reversed from the other. The two coils are then connected in series by reversing their connections. This type connection does not cause bucking of any current generated by the windings. However, if any hum voltage is induced in the core from an outside source, such as through the chassis or from an adjacent transformer, this will in turn induce a hum voltage in the windings. With the transformer constructed as described, equal hum voltage is induced in the coils on each leg but one voltage will be 180° out of phase with the other, and thus be cancelled out. In addition to the winding construction described in the foregoing, the whole transformer is encased in a nested shield which further reduces the effect of extraneous magnetic fields. Nested shields are discussed in Question 8.51.

8.99 ***Describe how single-phase power may be converted to three-phase power operation.***—Several circuits have been developed for this purpose. The circuits to be described may be used for recording and reproducing equipment motor circuits, or for any other purpose where only a source of single-phase power is available and three-phase power is required.

The basic circuit used in the Westrex RA-1511-A converter, a nonrotating device devised for converting 115-volt

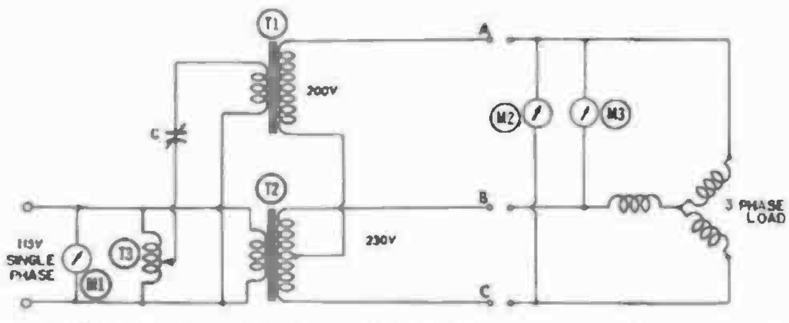


Fig. 8-99A. Basic circuit of Westrex RA-1511-A converter, used to convert single-phase power to three-phase power.

single-phase power to 230-volt three-phase power for driving camera motors, is shown in Fig. 8-99A. The circuit elements provide a means for correcting phase unbalance when motors are changed, and to compensate for changing load conditions.

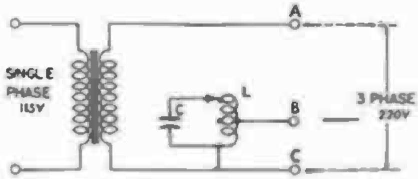


Fig. 8-99B. Schematic diagram showing how single-phase current may be converted to three-phase current by the use of a resonant circuit in the secondary circuit.

Basically, the circuit consists of a Scott transformer connection developed originally for the conversion of two-phase power to three-phase power. Using the circuit shown, the enrrent in the primary of transformer T1 is shifted 90 degrees with respect to the primary of transformer T2 by the connection of a capacitor C of the proper value in series with the primary of transformer T1. Thus, two-phase current is provided from a single-phase source. Three-phase current will only exist when the secondary sides of T1 and T2 are connected to an external three-phase inductive load, such as a motor.

The circuit is adjusted for a given set of load conditions by selecting a value of capacitance which will bring meters M2 and M3 to the same voltage reading (300 V), which is twice the line voltage read on meter M1 (150 V).

A continuously variable autotransformer T3, connected across the primary of transformer T1, permits the incoming line voltage to be adjusted to set the three-phase output voltage to its proper value. Voltmeters across the single- and three-phase voltage circuits permit the voltage to be monitored during actual operation.

In the commercial model, relays and switches are provided for protection to the capacitors and transformers in the event that trouble should develop in the load circuits.

In Fig. 8-99B is shown another circuit developed for converting single-phase current to three-phase current. In this circuit a parallel resonant circuit is connected in series with one leg of the developed three-phase current. This resonant circuit shifts the phase of the output current between 82 and 90 degrees, thus creating a third phase. For proper operation the frequency of the resonant circuit is adjusted under a given set of load conditions to provide the proper voltages to the external load circuits.

A third circuit, developed by William Ashworth, is given in Fig. 9-99C, and is suitable for driving 3-phase Y-connected motors of 1 to 3 horsepower. Here, one side of the single-phase power A is connected directly to phase A of the three-phase motor circuit. The initial current flowing from line B of the single-phase power source flows through a spring-loaded relay winding to phase B of the motor circuit closing the relay contacts and inserting start-

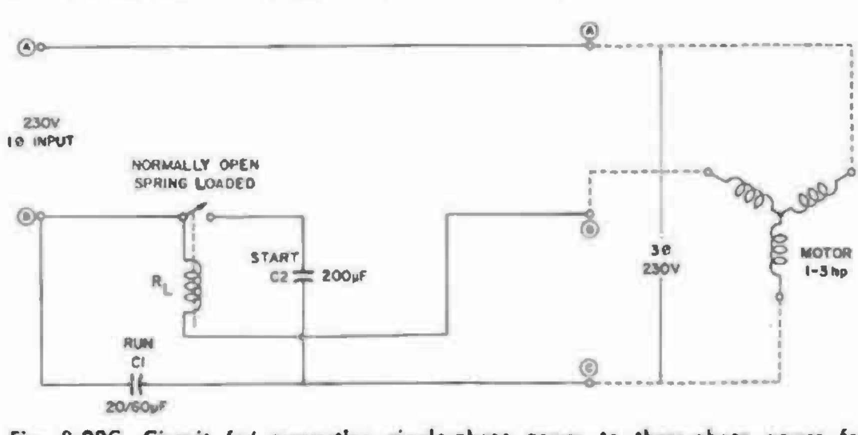


Fig. 8-99C. Circuit for converting single-phase power to three-phase power for motor circuits.

ing capacitor C2 into the circuit. This causes a reduction in the current flow through the relay winding. The spring-loaded action opens the contacts, disconnecting starting capacitor C2, leaving the relay winding connected in series with phase B. With the relay contacts open, running capacitor C1 is in the circuit creating a path from input line "B" of the single-phase, to phase C of the three-phase motor circuit.

Approximately 200 μF of starting capacitance is required, with 20 to 60 μF for the running capacitor C1, or about 20 μF per horsepower. An ammeter should be inserted in each leg of the motor circuit when it is operating under load conditions and the capacitance adjusted for equal current in the three legs.

The most critical part of this circuit is the relay, since its winding must be capable of carrying the motor current, and yet be able to open and close within the normal operating current range of the motor. In some instances the motor may not draw sufficient current when starting to actuate the relay, therefore the motor will not start. In this instance, some experimenting may be required to adjust the relay for the required balance between the starting and running currents. The described circuit will approximate a phase difference of 120 degrees between the three motor leads.

8.100 Describe the construction and operation of a constant-voltage transformer.—Constant-voltage transformers are special type transformers, containing voltage-regulating circuits, which automatically correct for line-voltage

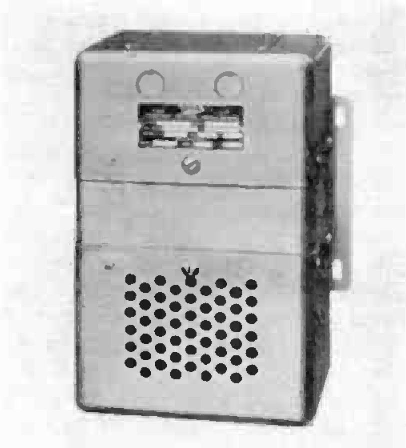


Fig. 8-100A. Constant-voltage transformer type CVS. (Courtesy, Solo Electric, Division of Sola Basic Industries)

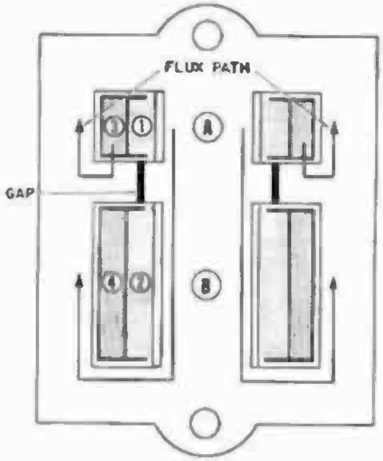


Fig. 8-100C. Cross-sectional view of the core and coil construction of a Sola Electric Co. constant-voltage transformer.

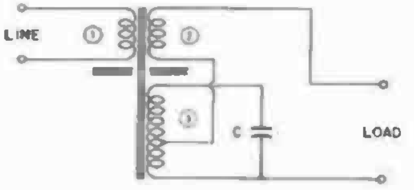


Fig. 8-100B. Basic circuit for Sola Electric Co. constant-voltage transformer using three coils.

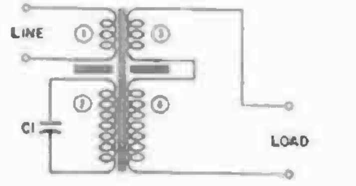


Fig. 8-100D. Basic circuit for Sola Electric Co. constant-voltage transformer using four coils.

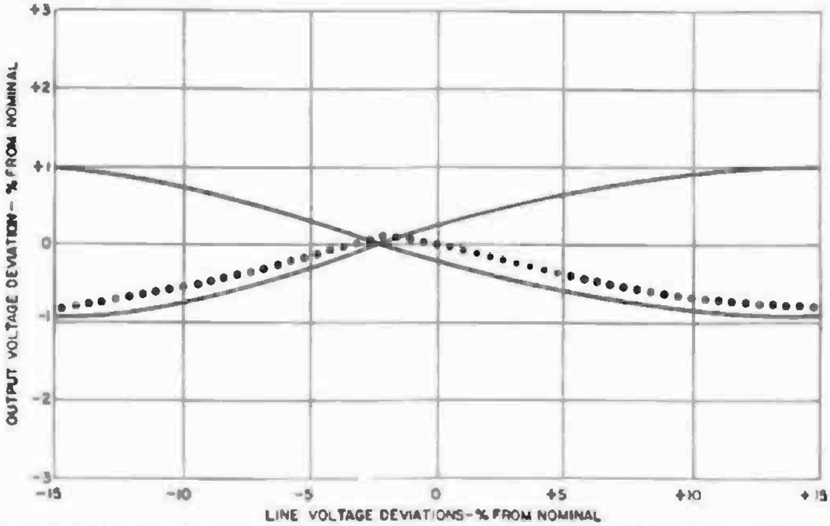


Fig. 8-100E. Residual variations within 1-percent limits, for Sola Electric Co. constant-voltage transformer.

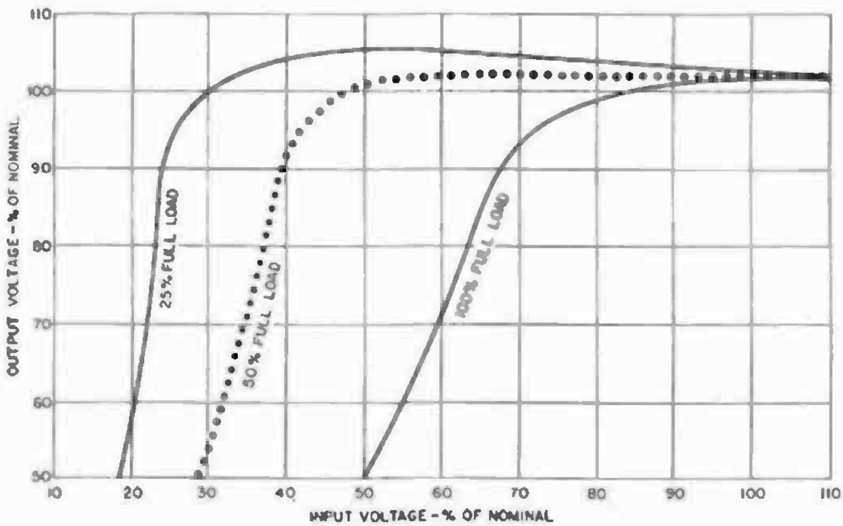


Fig. 8-100F. Characteristics for loads less than 100 percent.

changes, thus maintaining the load voltage constant. These transformers have no tubes or mechanical parts, and except for a slight hum, are noiseless. It should, however, be pointed out, these devices do have fairly strong magnetic fields and when installed, should be placed at some distance from equipment sensitive to such magnetic fields, *particularly magnetic-recording equipment.*

Constant-voltage transformers will be of interest to the audio engineer since many pieces of equipment in a sound installation require voltage regulation to within 1 percent of their normal operating voltages. Among these are transistor amplifiers, vacuum-tube heater and plate voltages, exciter lamps for optical film recorders, projectors, reproducers and similar devices. It is a well-known fact that when vacuum tube heater voltages are maintained within 1 percent, tube life is lengthened and the signal-to-noise ratio is increased. A model CVS constant-voltage transformer, manufactured by Sola Electric, is shown in Fig. 8-100A, and is typical of the type used for the above purposes.

Constant-voltage transformers by Sola are manufactured in two types, the *sinusoidal* type CVS and the *normal-harmonic* type CVN. Both transformers will regulate the output voltage to within 1 percent or less for a line voltage variation between 95 to 130 volts at the primary. The type CVN has about 14 percent THD and is used where line-frequency harmonics are of little consequence. Type CVS contains, besides the voltage regulating circuits, a harmonic-suppression circuit which reduces the internally generated harmonics to 1.5 to 3 percent (depending on the load conditions). Since the basic principle of operation is the same for both type regulators, the normal harmonic type (CVN) will be discussed first.

Referring to Fig. 8-100B, the voltage-regulating circuit involves the use of three coils, a primary (1), compensating winding (3), and a resonant winding (2), connected in parallel with a capacitor (C). A cross-sectional view of the internal construction appears in Fig. 8-100C. Primary winding (1) and compensating coil (3) are layer wound one over the other on part (A), of the center leg of the core. Resonant winding (2) is wound on part (B) of the same leg, but isolated from the primary and compensating winding by a magnetic shunt.

When an alternating current of low voltage is impressed across the primary winding, the resulting magnetic flux,

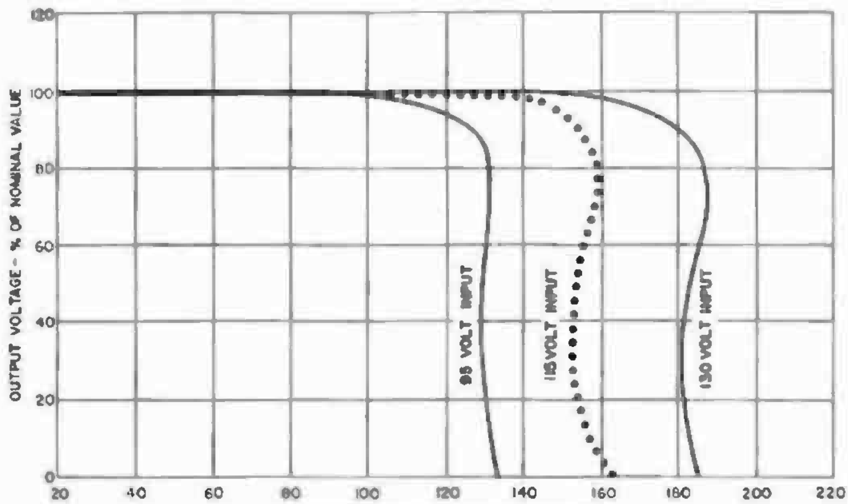


Fig. 8-100G. Effect of increasing the load on the output voltage for Sola constant-voltage transformer.

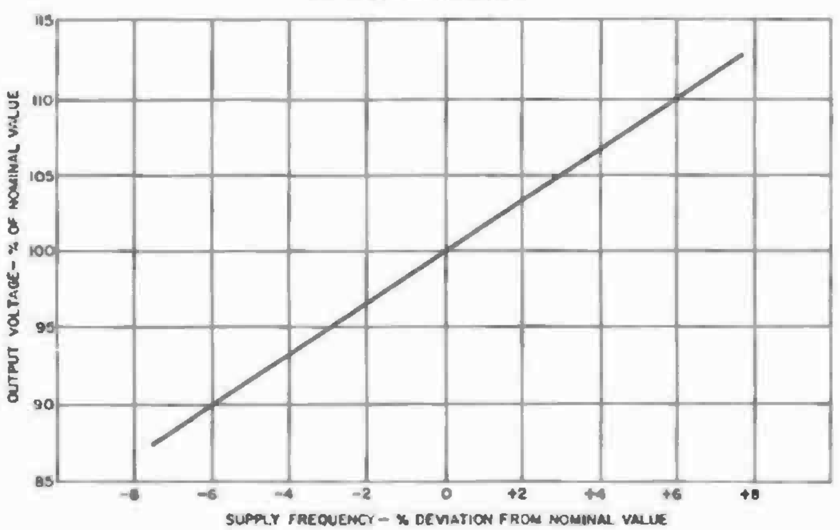


Fig. 8-100H. Effect of frequency on normal output voltage.

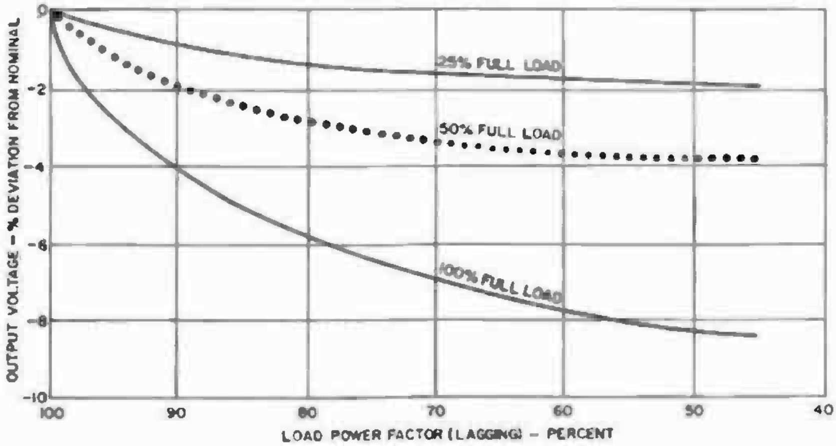


Fig. 8-100I. Effect of the change in median voltage level with changes in load and load power factor.

induces a voltage in winding (2). Because of the reluctance of the air-gaps in the shunt path, this voltage roughly approximates the turn-ratio voltage. As the voltage across the primary is increased, more flux threads through part B of the core structure. When this flux density becomes such that the inductive reactance of winding (2) approaches the value of the capacitive reactance of C at the frequency of the exciting voltage, this circuit becomes resonant and the voltage appearing across winding (2) rises rapidly to a stable, predetermined value which is higher than the calculated turn-ratio voltage. This has the effect of increasing the magnetic density in that portion of the magnetic circuit on which the resonant winding is wound (part B), and of greatly reducing the relative reluctance of the shunt system so that subsequent variations in flux, produced by changes in the primary circuit, are largely absorbed by the shunt system, and the voltage change in the resonant circuit is small. It remains only to compensate for this small change, and that is accomplished by winding (3).

The shunt system also operates to loosen the effective coupling between the resonant circuit and the primary winding so that once resonance is attained, the primary is required to supply only enough energy to overcome the iron and copper losses of the system in order to maintain the oscillation at no load.

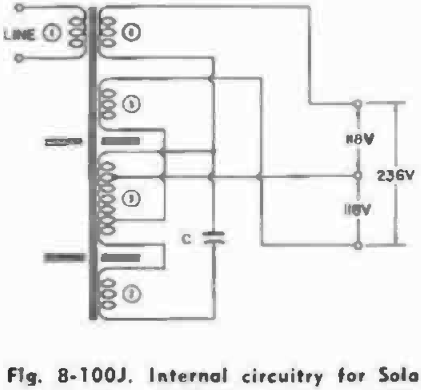


Fig. 8-100J. Internal circuitry for Sola Electric Company three-wire, constant-voltage transformer.

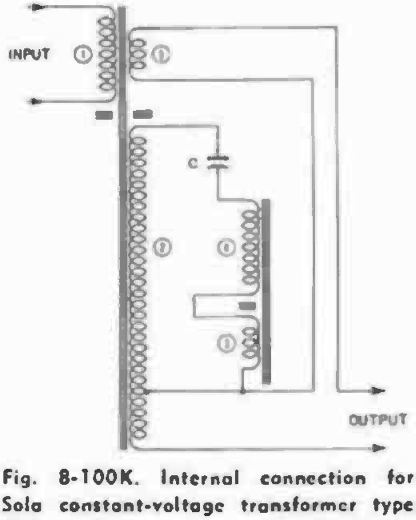


Fig. 8-100K. Internal connection for Sola constant-voltage transformer type CVS, with harmonic-neutralizing circuit.

Since the voltage across winding (2) is stable once resonance is attained, it may be used as a basis of constant-output voltage of the transformer. The output voltage is obtained by tapping across a portion (or all) of the resonant winding coil (2), Fig. 8-100B, or from an additional winding (4), wound over the resonant circuit (Fig. 8-100D). In either case, the compensating winding (3) is connected in series with the output load opposed to it. When this winding is so proportioned that the change in voltage induced by a change in primary voltage is roughly equal to the change appearing in the output, the resultant output voltage remains constant and independent of the voltage variation in the primary circuit.

Once resonance has been established in winding (2), the balance between the magnetic flux in part (A) and (B) of the core is maintained by the buffer action of the magnetic shunt system. Therefore, any load applied to a winding on part (B) will result in a passage of greater amount of useful flux through (B) from (A), exactly compensating for the energy consumed, and at the same time maintaining the oscillation of the circuit (2). The transformer then delivers regulated voltage within its rated capacity.

A regulation curve of 1-percent output voltage is shown in Fig. 8-100E, with regulating curves for loads less than 100 percent shown in Fig. 8-100F. The graph in Fig. 8-100G gives the effect of increasing the load and its effect on output voltage. The effect of changing frequency on a normal output voltage is shown in Fig. 8-100H. The effect of changes in the load power is given in Fig. 8-100I. The circuitry for a 3-wire constant-voltage transformer is shown in Fig. 8-100J. This transformer will supply the same regulation for the following conditions: from the 236-volt terminals alone, from either of the 118-volt legs, from a combination of the 236-volt and 118-volt loads, or from unbalanced 118-volt loads.

Voltage regulation may be obtained from loads other than unity power factor; however, only at the expense of lower output voltage. If the lower voltage under the lagging power factor is objectionable, correction may be made with capacitors connected at the load, or by means of the buck-boost transformer described in Question 8.96. The efficiency of these transformers is on the order of 75 to 90 percent. Power-factor correction is discussed in Questions 3.58 to 3.60. As a rule, constant-voltage transformers are *not* used on motor circuits supplying sound recording and reproducing equipment.

Voltage-regulating transformers of the CVS type, have 1 to 3 percent harmonic distortion, compared to 14 to 20 percent for the standard CVN type. To reduce such distortion, a harmonic-neutralizing circuit (Fig. 8-100K) is used. The primary coil (1), resonant circuit winding (2), buck-coil (5), capacitor (C), and the main core structure all function in the same manner as described for the CVN. Coils (3) and (4) of the neutralizer circuit are wound at opposite ends of the common core, and separated by a high reluctance magnetic shunt. The neutralizer assembly is connected into the resonant circuit, as most of the distortion (third harmonic) is generated in the particular portion of the electromagnetic circuit.

Coil (4) in series with the resonant circuit, when considered alone, will have *three times* the impedance for third harmonic currents as for the fundamental frequency. It thus offers considerable filtering action by its mere presence. The neutralizer circuit also reduces smaller harmonics, consisting of 5th, 7th and higher odd harmonics. Coil (3) connected in parallel with one section of the resonant winding (2), develops voltages proportional to all unwanted harmonics by the selection of a suitable winding and proper polarities. Since the above voltage-regulating transformer isolates the load portion of the circuit from the main ac source, a *separate* ground must be supplied. This is often a distinct advantage for eliminating heavy ground currents.

8.101 What precautions should be observed when using constant-voltage transformers?—They should be installed at least 50 to 100 feet distant from amplifier equipment having a considerable amount of gain. This is particularly true for photocell and microphone preamplifiers. Magnetic recording and reproducing equipment due to its nature should not be placed within less than 50 feet of such transformers, and farther if practical. Voltage readings at the load side must be made using either a *dynamometer* or *thermocouple*-type meter (such as a Weston-662). *Rectifier-type meters or vacuum-tube voltmeters should not be used, as the readings may vary several volts,* depending on the percent harmonic distortion of the voltage. It is possible for a vacuum-tube voltmeter to read 120 to 127 volts for an actual voltage of 117 volts. (See Question 22.103.)

The change in output voltage resulting from a resistive load is usually small, running to less than 1 percent. The power-factor will cause the output voltage to vary from the normal rating of the transformer if the load circuit has a power-factor other than that specified on the transformer data plate. Load regulation will also be relatively greater as the inductive load power factor is decreased.

8.102 Can constant-voltage transformers be connected in tandem to improve regulation?—Yes. However, when two units are connected in tandem, the output of the second unit will show little or no detectable change arising from supply-line variations up to about 15 percent. Cascade or tandem operation is recommended for special applications where the regulation must be in the region of 0.25 percent.

Operating sinusoidal voltage transformers in tandem may increase the harmonic distortion up to about 5 percent. Also, the transformers become more frequency sensitive because of the two resonant circuits, one in each transformer. This alone can cause about 0.25-percent variation in regulation. The first, or driver, transformer must be slightly larger in capacity to overcome the losses of the driven unit. The first transformer may be of the nonsinusoidal type, and the second a sinusoidal type if necessary.

8.103 Can constant-voltage transformers be connected in parallel?—If the transformers are of the same voltage rating and capacity, the primaries or secondaries may be connected in parallel to obtain a greater power output. However, regulation may suffer.

8.104 Describe an electromechanical voltage regulator.—Electromechanical line-voltage regulators combine the power-handling capabilities of a motor-driven variable autotransformer with the fast response and accuracy of an electronic feedback loop. This type regulator introduces no harmonic distor-

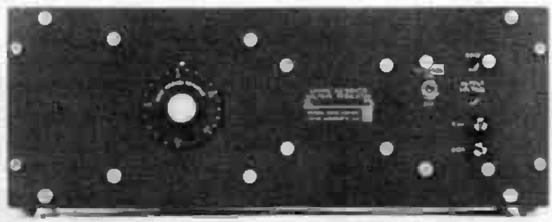


Fig. 8-104A. General Radio Co. Model 1581-A electromechanical voltage regulator.

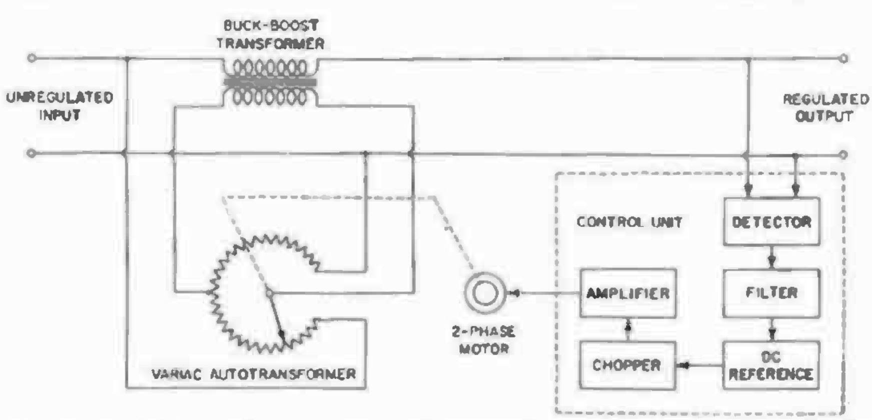
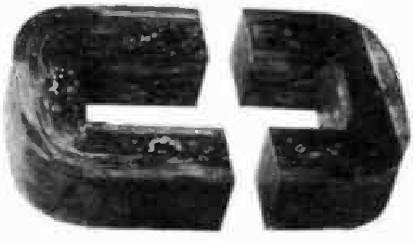


Fig. 8-104B. Block diagram of the electromechanical voltage regulator Model 1581-A manufactured by the General Radio Co.

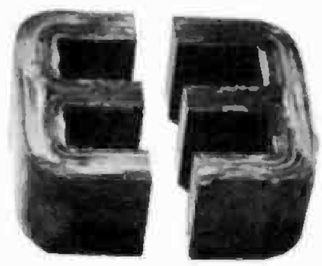
tion, and is totally unaffected by changes in load power factor, from zero leading, to zero lagging, and is insensitive to load currents. With proper design it can hold the output voltage constant with wide swings of line frequency, a factor most important in the regulation of voltage from portable and emergency generators. It is well suited for applications involving heavy starting currents and can withstand overloads up to 10 times the rated output current. Such regulators may be obtained with power-handling capabilities up to several kVA, for either single or three-phase operation. Such a voltage regulator is pictured in Fig. 8-104A, manufactured by the General Radio Co., with a block diagram of its circuitry shown in Fig. 8-104B. Voltage deviations at the output activate a servo-feedback loop consisting of a control unit, a two-phase motor, a Variac autotransformer, and a step-down buck-boost transformer. The deviation is thus translated into a correction voltage that is added to, or subtracted from, the input to bring the regulated voltage to the correct value. The solid-state control unit converts any small deviation in output voltage into a proportional electrical signal to drive the motor. The deviation is first sensed by an rms detector whose dc output, after filtering, is compared with a 9-volt reference voltage derived from a zener diode. The resultant difference voltage is chopped into an ac error signal, with a magnitude proportional to the output deviation; the phase is determined by the direction of the deviation.

The ac error voltage from the control unit drives the two-phase servo motor, with the phase and magnitude determining the direction of rotation and the motor speed. The motor drives the autotransformer through a gear train. Because the output voltage of the autotransformer is stepped down by the action of the buck-boost transformer, the full adjustment range of the autotransformer can be used to produce a relatively narrow range of control. Remote circuits are provided for connections to the load-line at some selected point. Thus, corrections are made for the voltage drop in the supply line.

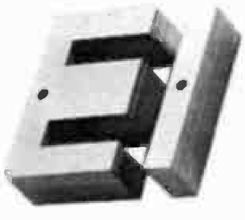
Two screwdriver controls on the front panel provide for adjustment of the amplifier gain and output voltage. The dial indicates the percent difference between the input and output voltages and permits manual voltage adjustment ranges from 0.25 to 0.50 percent of the output voltage, with response speed of 20 to 160 volts per second, and a correction range of 82 to 124 percent, depending on the model.



(a) Silectron grain-oriented "C" cores.



(b) Silectron grain-oriented "E" cores.

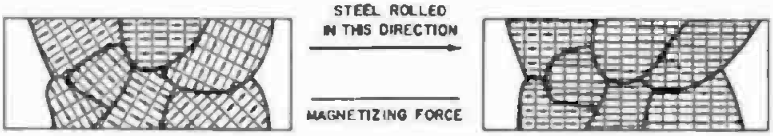


(c) "E" and "I" laminations.

(d) Powder core.

(e) Toroidal-coil core.

Fig. 8-105A. Typical cores used for power and audio transformer construction, and manufactured by Arnold Engineering Co.



(a) *Ordinary silicon steel.* (b) *Grain-oriented Hipersil Steel.*

Fig. 8-105B. Comparative grain arrangements. (Courtesy Westinghouse Electric and Manufacturing Co.)

8.105 Show the different type cores and laminations used in the construction of audio transformers. — Transformer laminations and cores are many and varied. It is the trend in the manufacture of high-quality audio transformers to use the "C" and "E" type, grain-oriented steel cores, as shown at parts (a) and (b) of Fig. 8-105A. The particular group of cores shown are manufactured by Arnold Engineering Co., and marketed under the trade name of *Silectron*. Grain-oriented steel is a cold-rolled grade of 3-percent silicon steel, manufactured by the Allegheny Ludlum Steel Corp. It has high-saturation flux density, lower core losses, and lower exciting volt-amperes than regular silicon steel. This high degree of orientation is preserved by the cutting of the core and the *gapless* construction. The steel strip is coated on both sides, which provides good interlamination resistance with a negligible effect on the space factor. The coated strip is slit to the proper width, and wound on a mandril to make a gapless core. The core is then annealed to relieve winding stresses, and cut to produce two core halves, with a highly polished surface. The effective air gap, when the surfaces are placed together, is 0.001 inch.

Similar cores manufactured by Westinghouse are known under the trade name of *Hipersil*. The magnetic properties of grain-oriented steel are achieved by the orientation of the steel crystals through rolling and heat treatment. Individual crystals line up with their edges essentially parallel to each other and parallel to the direction of the rolling sheets. Where crystals are oriented to all face in one direction, the steel has a much higher magnetic permeability in that direction than does steel with crystals pointing at random. Because of this arrangement of the crystals, grain-oriented steel requires a smaller external magnetizing force to produce a given flux than does unoriented steel. This results in a very high high-density permeability, high low-density permeability, high incremental permeability, and very low losses in the direction of the rolling. For comparative grain arrangement of ordinary silicon steel and grain-oriented silicon steel, refer to Fig. 8-105B. Each arrow represents the direction of easiest magnetization of the individual crystals forming the steel.

The "E" shaped cores at Fig. 8-105A, part (b) are used for shell-type transformers, where the coil is placed over the center leg. The "E" and "T" laminations shown in part (c) are used for both audio and power transformers. The powder core shown in part (d) is designed for two halves, one over the other, completely enclosing the coil. The toroidal core shown in part (e) supports the coil, which is shuttle wound around the core. The "C" and "E" cores are also used for pulse transformers and for 400- to 3000-Hz power transformers.

8.106 Describe a plug-in audio transformer.—Because of the small size of solid-state amplifiers, and the fact that they can be plugged into mounting trays, it has become necessary that audio transformers in some instances take the same form. Fig. 8-106 pictures such a transformer, manufactured by McCurdy Radio Industries Ltd, of Canada, designed in module form. The transformer is assembled using a contemporary dual glass-epoxy card design with gold-plated printed circuitry. Internal electrostatic shields are provided with a heavy external magnetic shield, to reduce pickup from stray magnetic fields. The transformer shown is a repeat coil, employing four windings, that may be connected for 150- or 600-ohms impedance.

8.107 How may three-phase loads be controlled using two variable autotransformers?—Two 120-volt autotransformers are connected in open-delta, as seen in Fig. 8-107. Maximum voltage

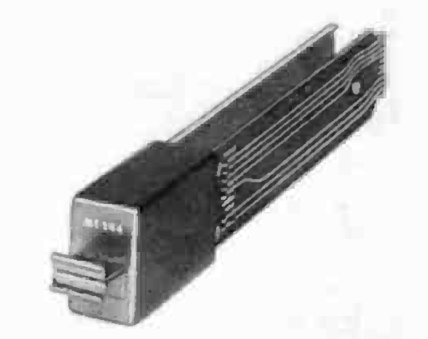


Fig. 8-106. Plug-in repeat coil manufactured by McCurdy Radio Industries Ltd. of Canada.

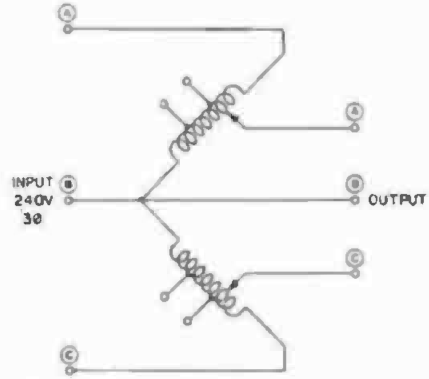


Fig. 8-107. Controlling a three-phase load using two 120-volt autotransformers connected open-delta.

attainable can be either the line voltage, or 17 percent above the line voltage by using the over-voltage connection. The load rating for a single unit is $\sqrt{3}$ or 1.732 times that of a single unit. If 240-volt autotransformers are used, an output voltage of more than double the supply voltage is possible, although the current and power ratings are halved.

8.108 How are three autotransformers connected for three-phase operation? —Three 120-volt autotransformers are wye-connected, as in Fig. 8-108. This connection is possible because the voltage from the line to neutral of the wye-connected assembly is the line voltage divided by the $\sqrt{3}$. Thus, in the case of the 240-volt three-phase line, the voltage across each unit will be 138 volts. Since each single unit is wound for 140 volts across the whole winding, three 120-volt units can be wye-connected if the over-voltage connection is omitted. Although the over-voltage connection cannot be used, the kVA is increased by the ratio of 138/120. The load rating of the wye-connected assembly is 3.47 times that of a single unit. Similarly, 240-volt units can be used on 480-volt three-phase lines.

8.109 Describe the construction of a unity-coupled driver transformer for a class-B transistor push-pull amplifier.— In Figs. 12-255 and 12-256, are shown two transistor amplifiers capable of producing 100 watts of audio power, with low distortion. To drive the output stages, a driver transformer of *low* turns ratio and *low* impedance, having as close to *unity* coupling between windings as possible, is required. To acquire such coupling, the primary and secondary windings must be wound *bifilar* style, that is—the wires must be wound parallel, or side-by-side. (See Question 5.16.) The primary is split into two or more coils and wound along with the secondaries. These windings are termed *bifilar*, *trifilar* and *pentafilar*, depending on the number of wires concerned.

A typical bifilar-wound transformer that may be used as the driver transformer for the amplifier of Fig. 12-255 is wound, using a split-primary bifilar wound with two secondaries. A 600-turn primary is first wound on a nylon bobbin with a ¾-inch square hole, using Number 30 enameled wire. Next,

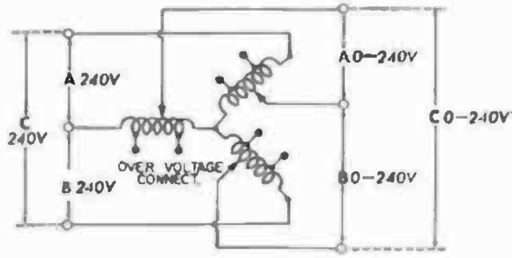


Fig. 8-108. Three 120-volt autotransformers connected for 240-volt, three-phase operation. The over-voltage connection cannot be used.

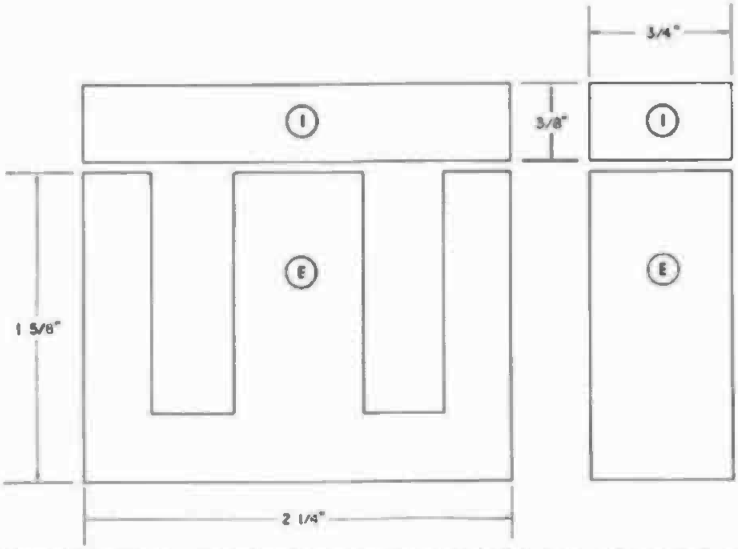


Fig. 8-109A. E and I core laminations used in class-B driver transformers for transistor power amplifier.

two secondaries are wound, using two simultaneous windings (bifilar) of Number 27 enameled wire for 200 turns. The primary is then continued by winding another 600 turns over the top of the other windings. The ends are brought out and the starts and endings indentified. The approximate dc resistance of the primary is 45 ohms, and of the secondaries, 3.3 ohms each. The core is made of EL-75, grade M-19 silicon steel "E" and "I" laminations, using a 1-mil air gap (see Fig. 8-109A). Enough laminations are used to fill the hole in the bobbin (¾-inch). This results in a transformer with a turns ratio of 6:1:1. In using the transformer, the proper polarities must be observed. This may be determined with the use of a compass and a dry cell, or as described in Question 8.90.

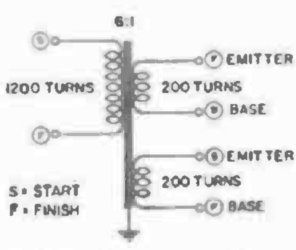


Fig. 8-109B. Driver transformer coil configuration for transistor class-B push-pull amplifier stages.

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Section 9

Sound Mixers

The design of todays sophisticated mixing console is quite complex, due to the multiplicity of devices associated with the mixing network. Many will recall when paralleled battery rheostats were used as mixer controls for broadcasting and motion picture sound recording, and will view the modern console as an outgrowth of the basic principles applied at that time. With miniaturization, many components are now included in the mixer network, rather than at a remote position. Printed circuit plug-in designed components facilitate replacement in the event of failure. Any of the vacuum tube designs may be used with solid-state components, with certain considerations.

Several different type solid-state and vacuum-tube mixers are discussed. A twelve-position, three-section, four-bridging bus dubbing mixer is described, with its associated monitoring, talk-back, reverberation, and signal circuitry. Solid-state devices employed in mixing consoles are discussed in Section 12.

9.1 ***What is a sound mixer?***—A resistive network designed to provide a means of combining several separate audio signal sources into one composite signal. The signal sources may consist of dialogue, music, and sound effects. The sources of signal may be from a broadcast line, optical or magnetic film sound tracks, records, a live pickup, or any combination of these sources.

The network is designed so that changing the level of any one of the individual signal sources has no effect on the level or frequency characteristics of the other signal sources in the network. For broadcasting and recording purposes, at least six positions are required. For recording purposes, equalizers, filters, and other devices are included in the mixer console but do not form a part of the mixer network. A mixer console designed for recording and a-m and fm broadcasting is shown in Fig. 9-1.

9.2 ***What is low-level mixing?***—A mixing network similar to that shown in Fig. 9-2 which uses no amplification

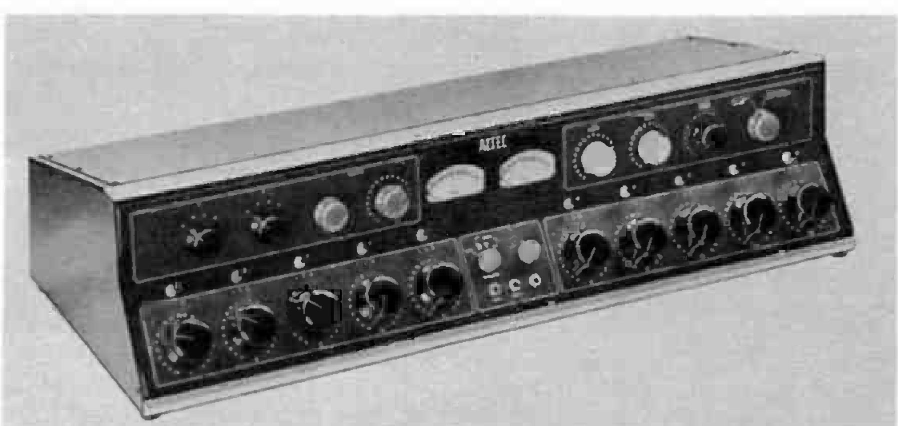


Fig. 9-1. Ten-position mixing console Model 250-SU, manufactured by Altec-Lansing.

between the signal source and the mixer control. The signal-to-noise ratio is low for this system and it is not used in professional installations. This method of mixing is now obsolete.

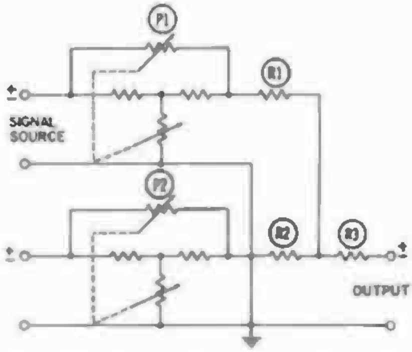


Fig. 9-2. Two-position low-level mixing network. Signal source is fed direct to mixer controls without amplication.

9.3 *What is high-level mixing?*—A mixing network which uses a preamplifier between the signal source and the mixer control as shown in Fig. 9-3. The advantage of high-level mixing is that the signal-to-noise ratio is increased in proportion to the gain of the preamplifier.

9.4 *What is the gain of the average mixer-preamplifier?*—About 40 to 60 dB. In some of the older installations a slight amount of equalization is included in the preamplifiers to compensate for microphone and recording-channel characteristics. However, in modern equipment this is the exception rather than the rule. Preamplifiers are designed to have uniform frequency characteristics from 20 to 20,000-Hz, within plus-minus 0.5 dB or less, with reference to 1000 Hz. This is true for both transistor and vacuum-tube types. Preamplifier design is discussed in Question 12.72.

9.5 *What are the principal components of a mixer console?*—Mixer networks consist of a group of variable controls, building-out resistors, a submaster and overall master control, and isolation coils or hybrid coils. In mixer networks specifically designed for motion picture rerecording, the network will also include low- and high-frequency attenuators (see Questions 6.80 and 6.81). For motion picture rerecording at least 8 mixer-control positions are required, and preferably 12. These are split into groups of four, with a submaster control for a given group of controls. A master control is sometimes included for fading all positions simultaneously; however, this is generally confined to broadcast-type mixer circuits. In addition to the mixer controls, the networks will include equalizer circuits operated by key switches or push buttons, which are not really a part of the mixer network, but are often included as a component part. A typical low-level mixer network is shown in Fig. 9-2, and a high-level mixing circuit is shown in Fig. 9-3.

The minimum requirements for a console might consist of the following components:

4 input circuits.
4 fixed attenuators to prevent overloading of microphone preamplifiers during the recording of high-level program material.
4 mixer controls, straight-line or rotary.

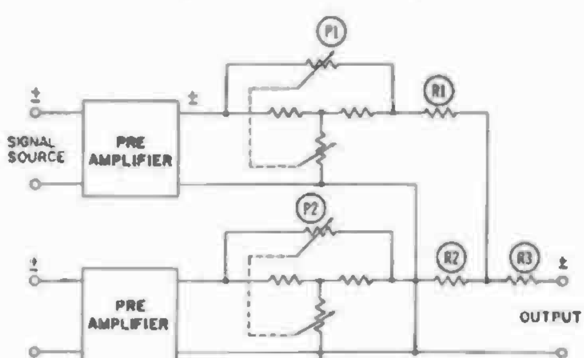


Fig. 9-3. Two-position high-level mixing network. Signal source is amplified then fed to the mixer control.

1 master control.
1 program equalizer (variable).
*1 graphic equalizer.
*1 compressor amplifier.
1 VU meter and attenuator.
* Intercommunication to other parts of the installation.
Monitor gain control.
Headphone jack and control.
Necessary amplifiers and attenuator networks.

*These items may be omitted; however, for good recording they should be included if possible. See Question 9.52.

9.6 *Describe the construction of typical recording and broadcast mixer consoles.*—A small six-position recording console, designed and built by RCA is pictured in Fig. 9-6A. The mixer network consists of two sections, each section containing three vertical ladder configuration mixer controls, and three graphic equalizers that may be connected into the mixer positions by means of push buttons. The outputs of the two mixer sections are combined in a hybrid coil. Two program equalizers, preamplifiers, compressor-limiter amplifiers, dB compression and VU meter, high- and low-pass filters complete the principal components. The system employs a single bridging bus. All amplifiers pertinent to the console are of the plug-in type, accessible at the rear of the console. Remote control for house lighting, ventilation and other functions of the studio are provided on a panel at the left end of the console. Two electrically driven footage counters mounted at the right, and operated from projection machines indicate the footage for both 16 and 35mm film. The operating levels shown on the block diagram of Fig. 9-6B are only approximate, and will vary with the type equipment and installation requirements. The monitor loudspeaker behind the screen is driven from a 40-watt power amplifier in the projection booth.

A mixer console, designed and manufactured by Altec-Lansing, for recording or a-m and fm broadcasting was shown in Fig. 9-1. The console contains 10 plug-in preamplifiers, 3 booster amplifiers, 3 program amplifiers (miniature plug-in), 10 mixer controls, along with the necessary networks, key-switches, and 2 VU meters with their auxiliary controls. The block diagram for this console appears in Fig. 9-6C. At the left are ten external input circuits, which may be from any source of sound, such as a turntable, magnetic tape recorder, telephone lines, microphones, etc. The output of each preamplifier is fed to a separate mixer control, then through combining networks, 3-position key switches, and to three bridging busses. The output circuitry provides for single (monophonic), two-channel stereo, or three-channel/two-channel operation.

Three-channel/two-channel operation consists of using the center position

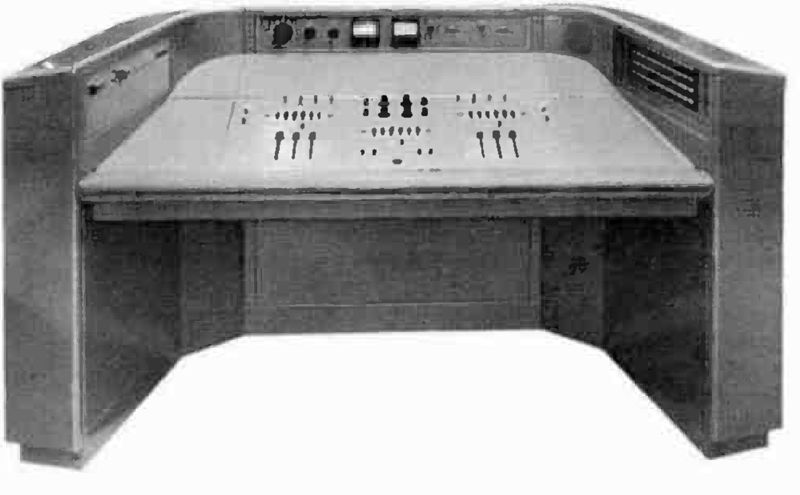


Fig. 9-6A. Small rerecording console which was designed and built by RCA for single track recording.

of the bus-selector key switch to feed a third, or center-channel, mixing bus. The output from this bus is amplified, then divided by means of a splitting pad and introduced into the left and right channels. For stereophonic use, this permits the vocal or dialogue to be picked up on a single microphone and evenly divided between the left and right channels. This reduces the need

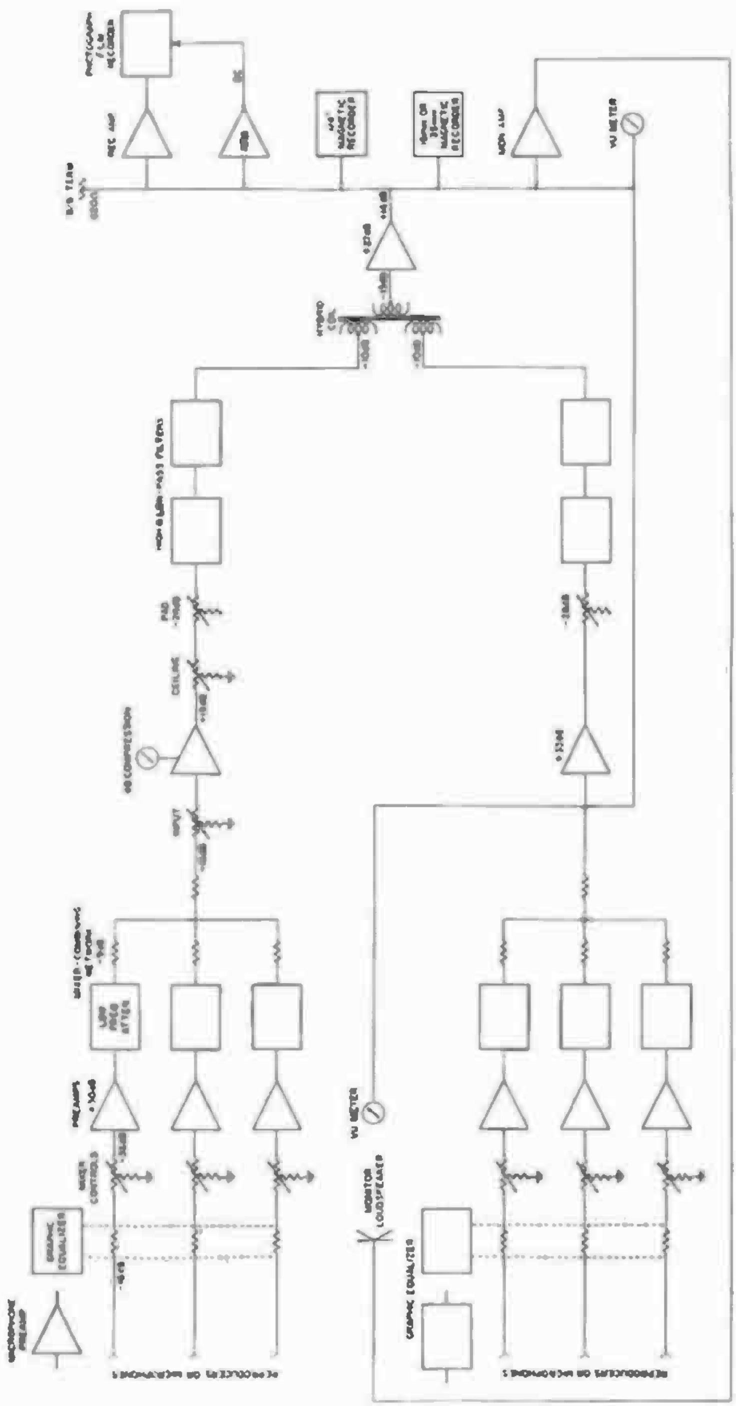


Fig. 9-6B. Block diagram and approximate operating levels for the mixing console which is shown in Fig. 9-6A.

for balancing and microphone matching. In this mode of operation, the variable equalizer is connected in the circuit for dialogue equalization (see Question 18.81). Provision is also made for splitting the center of the channel left and right for use in recording master tapes, where it is recorded on three channels, then reduced to two channels for release on disc or tape. Monitor am-

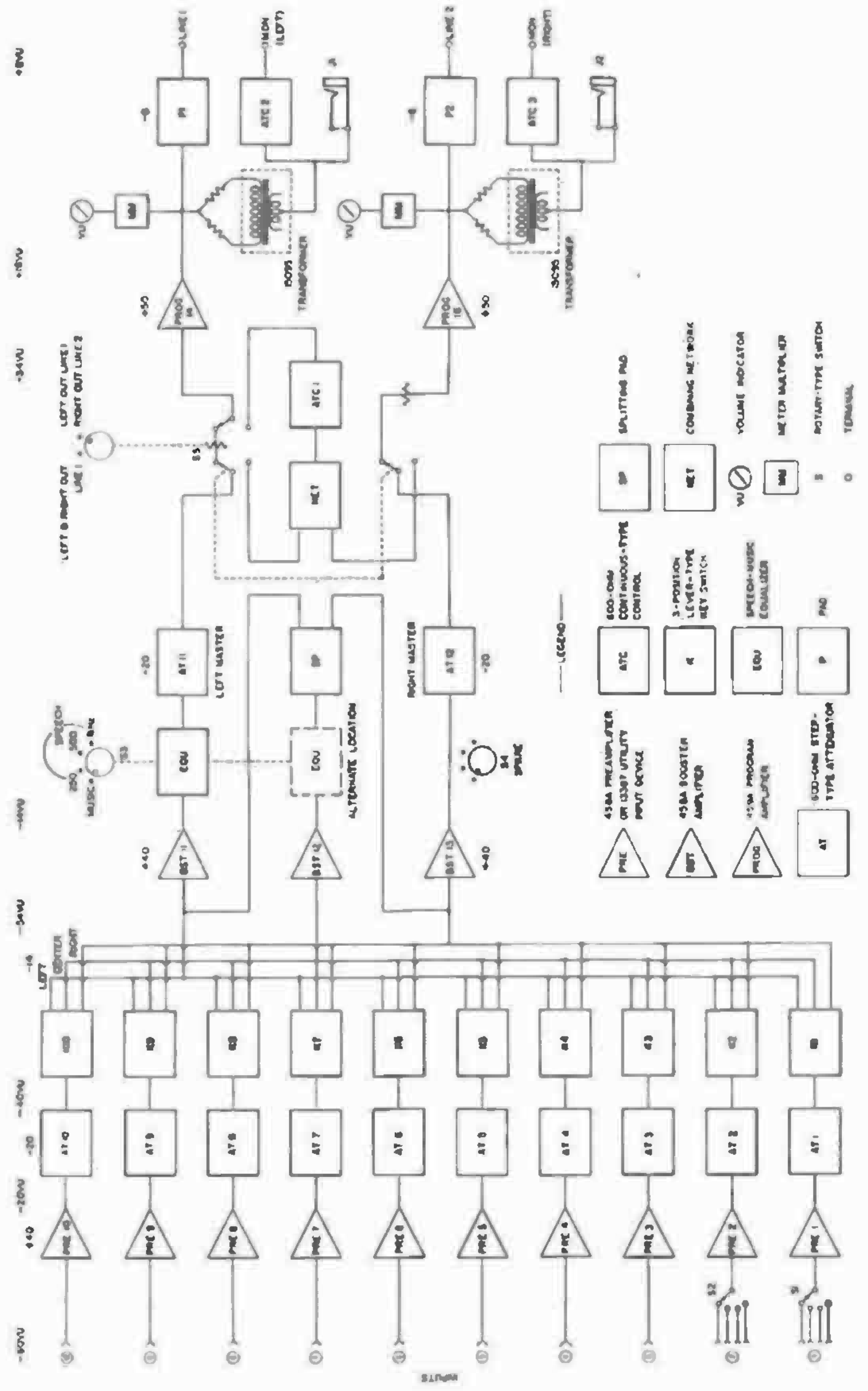


Fig. 9-6C. Block diagram for the Model 250-SU Altec-Lansing mixer console.

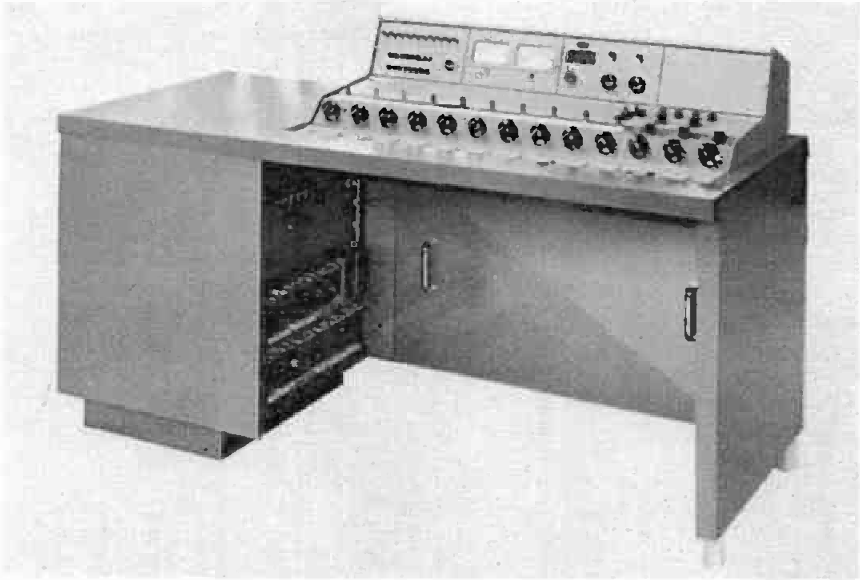


Fig. 9-6D. Ten-position mixing console Model SS4400 manufactured by McCurdy Radio Industries Ltd. Canada.

plifiers are not included in this console, since these items vary with the individual installations. A separate power supply provides the necessary plate and heater voltages. The frequency response is plus-minus 1 dB, 30 to 15,000 Hz. Distortion is 1.0 percent at plus 24 dB, signal-to-noise ratio, minus 70 dB referred to plus 18-dB output, with minus 50-dB input.

The Model SS4400 console, pictured in Fig. 9-6D, manufactured by McCurdy Radio Industries Ltd. (Canada), is designed for recording and a-m and fm broadcasting use. The basic circuit consists of 10 mixing positions (with cueing positions), solid-state preamplifiers, equalizers, VU meters, power supplies, module-type transformers, monitor amplifiers, and the necessary auxiliary controls. A total of thirty input circuits are provided, switchable to the 10 mixer positions, via the three-position key switches. After leaving the key switch, the signal is applied to a microphone preamplifier or a matching transformer. In addition to the thirty inputs, seven remote line inputs are available by push buttons, providing two switchable outputs that may be wired to any of the thirty input circuits. The ten mixer controls are ladder configuration, and feed 10 booster amplifiers. The mixer network outputs are selected by means of key switches for the two program busses. Two of the mixer circuits in-

Fig. 9-6E. Electrodyne Model ACC-1204 12-position mixing console.

corporate relays for remote control of the announcing microphone.

The two program channels employ line amplifiers to raise the level sufficiently to provide two output circuits at plus 18 dBm. Ten-dB pads are used between the output of the line amplifier and the output to provide a solid termination and isolation. An additional output is provided at each output for

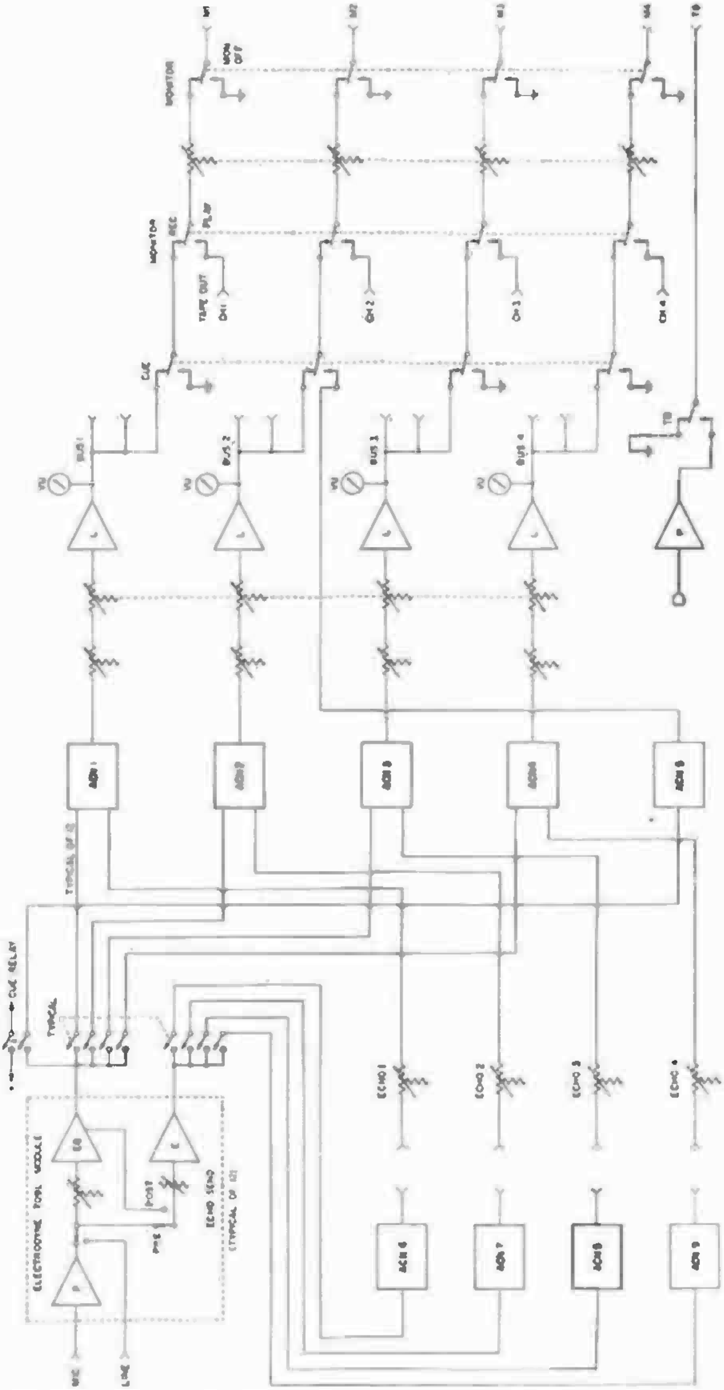


Fig. 9-6F. Block diagram for Electrodyne mixer console. This console was especially designed for use in recording phonograph records.

headphones. One VU meter is connected permanently across program amplifier 1, and the second may be switched to program channel 2 or across four auxiliary VU meter inputs for observing external program levels.

A complete talkback system is included, with push buttons for the selection of the console microphone, or inputs from announce turrets or remote cue and talkback circuits. Although the console shown has only one program equalizer, space has been provided over each mixer control for installing additional equalizers. The equalizers are similar to those discussed in Question 6.100. Circuitry is also supplied for connection to an external reverberation unit, and pre- and post-fades. Shown at the lower left in 9-6D are plug-in transformers (see Question 8.106), preamplifiers, power supplies, and monitor amplifiers.

The overall frequency response of this unit is plus-minus 0.5 dB, 30 to 15,000 Hz. Distortion is less than 0.5 percent for plus 18-dB output. Cross talk between the two program channels is not less than 65 dB. Signal-to-noise ratio is 64 dB for 78-dB gain at plus 18-dB output.

Pictured in Fig. 9-6E is a 12-position mixer console, manufactured by Electrodyne, employing integrated circuit amplifier modules discussed in Question 6.132. Module-design concept is used throughout the console, where practical. The mixing network uses active combining networks as discussed in Question 5.98, thus eliminating the need for booster amplifiers.

This console has been especially designed for recording, using a four-track tape recorder (or other types). Each of the 12 input circuits contain a 6-position high- and low-frequency equalizer-amplifier module, with a straight-line mixer control. Four separate output circuits, each with its own VU meter, provide visual monitoring facilities. A block diagram of its internal circuitry appears in Fig. 9-6F.

9.7 Describe the construction of a single- and multiple-mixer control.—The internal construction of a Model K-1 Gotham Audio Corp., single-element control is pictured in Fig. 9-7A. The control element is straight-line operated, although the attenuator element is a rotary one. The configuration of the control is the ladder attenuator discussed in Question 5.51. The attenuator element is hermetically sealed, with the contacts encapsulated in liquid silicone oil to reduce contact noise. The length of travel is 5¾ inches, employing a total of 56 contacts. In the upper 60 percent of travel (the usual operating range), the attenuation is 0.5 dB per step, with an increasing taper up to 85 dB before reaching infinity.

The attenuator element is actuated by

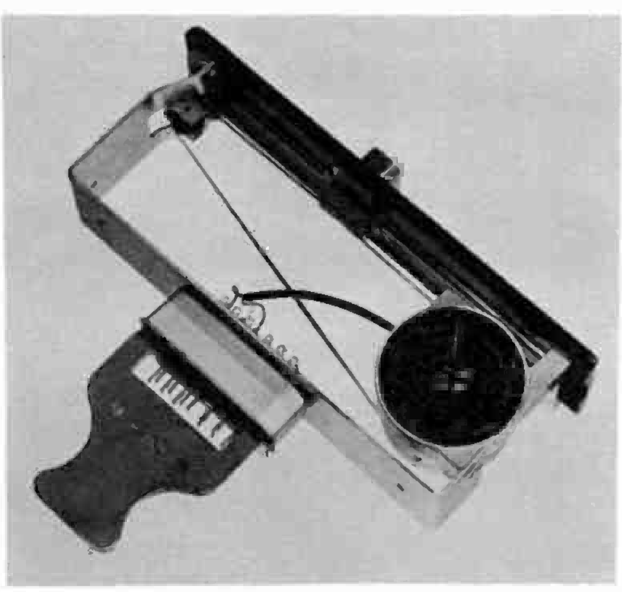


Fig. 9-7A. Interior view of Model K-1 Gotham Audio Corp. vertical mixer control.

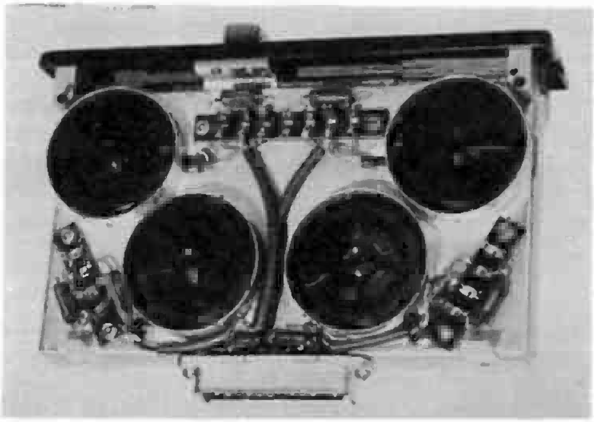


Fig. 9-7B. Interior view of Model K-4 Gotham Audio Corp. vertical mixer control.

a knob connected by a toothed belt imbedded with steel strands, and requiring only 180 to 240 grams of pressure for operation of 1 to 4 controls. A cable connector is provided for easy removal from the console. A Model K-4 mixer control, a four-element version using similar attenuators, is shown in Fig. 9-7B. Mixer controls of this design are quite common in stereophonic mixing consoles. Because of the additional three attenuators in the model K-4, ball bearings are used for ease of operation.

Fig. 9-8. Slide-wire ladder mixer control.

9.8 Describe the construction of a ladder slide-wire mixer control.—Slide-wire mixer controls find their greatest use in rerecording consoles, and are designed using the equation in Question 5.51. Such attenautors may be designed mechanically for either rotary or vertical usage. A typical rotary type is shown in Fig. 9-8. Vertical attenuators are similar in design, except the resistor card is straight, and a finger contact is used. Ladder attenuators have a fixed insertion loss of 6 dB, which must be taken into consideration in designing a mixer network. The signal-to-noise ratio of a ladder attenuator increases with the increase of loss.

9.9 Are balanced configurations used in mixer-pot design?—Yes. In special cases balanced mixer pots are used. These are generally of the balanced bridged-T type as shown in Fig. 5-38. Balanced-H configurations may also be used; however, because of the cost and the fact a balanced-H configuration requires six rows of contacts compared to four for the balanced bridged-T pot, the latter is preferred. A bridged-T pot also has a lower noise level and requires less maintenance.

9.10 What is the maximum attenuation required for a bridged-T mixer control used in a rerecording mixer?—At least 45 dB or greater, in steps of 0.5 dB. After passing 45 dB, the taper increases quite rapidly to afford a fast cutoff. Generally, the last steps are 6, 9, 12 dB, and infinity (see Question 5.80).

9.11 What is the minimum acceptable level of leakage that may be tolerated in a given position of a mixer network?—A minimum of 70 dB. The leakage is measured at a frequency of 10,000 Hz as shown in Question 23.67.

9.12 What does the term fade mean?—To attenuate a signal slowly to infinity.

9.13 What is cross-fading?—The gradual attenuation of one signal as another is gradually brought up to normal level. This is accomplished by closing one mixer pot as another pot is being opened.

9.14 *What is a board fade?*—An expression used in the broadcast industry which means to fade out all signals on the mixer by means of a master control.

9.15 *What is a submaster control?*— A control at the output of a group of mixer controls such as a split mixer. A

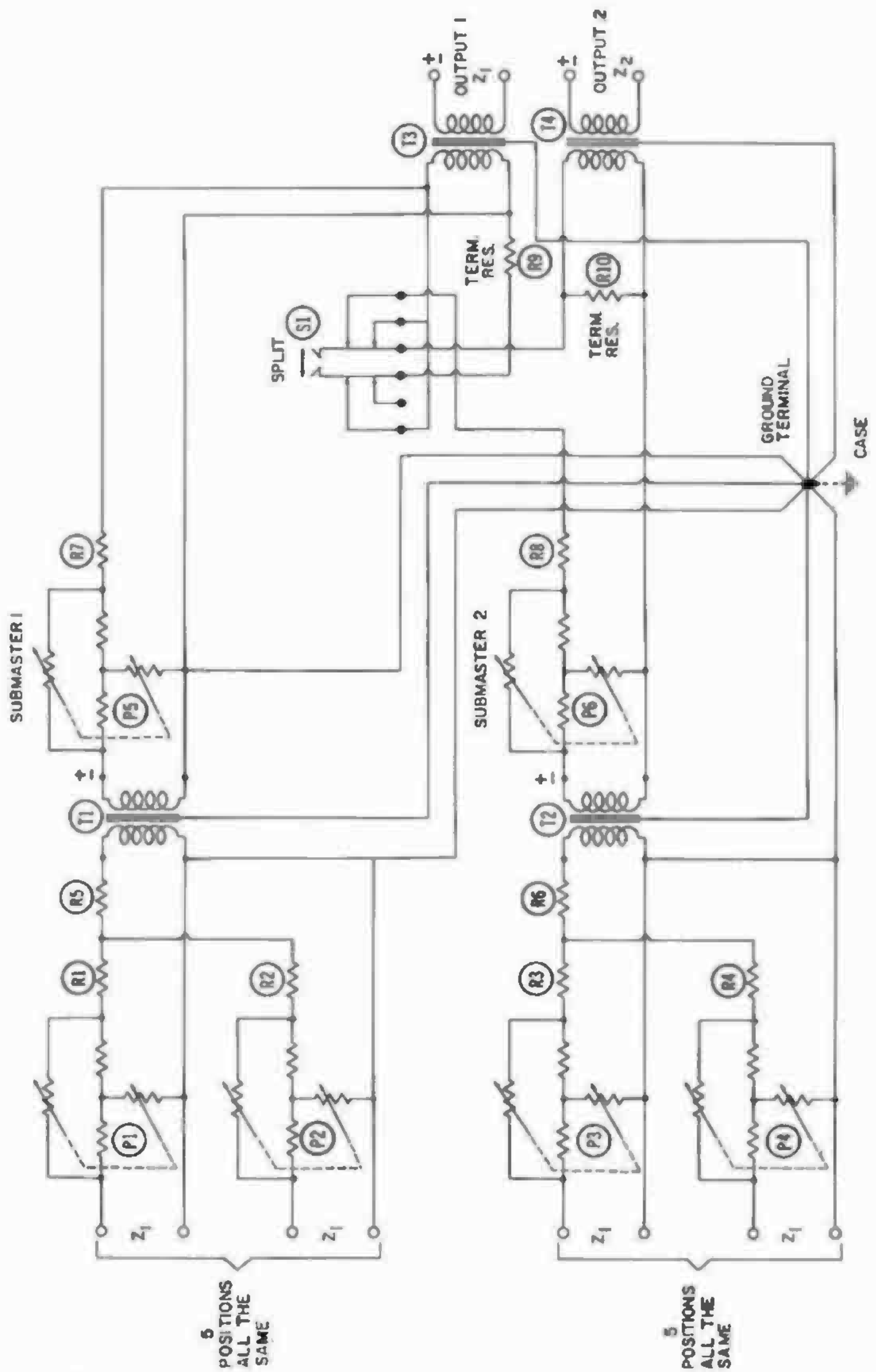


Fig. 9-18. A 10-position split-mixer network with a key switch for combining the two sections into a single network.

submaster control affects only a particular group of controls.

9.16 ***What is an overall master control?***—A control that is located at the output of the mixer network for the purpose of controlling all of the mixer positions simultaneously.

9.17 ***What is a grand master control?*** — It is the same as an overall master control.

9.18 ***What is a split mixer?*** — A mixer network split into two or more sections whereby a given group, or groups, of mixer controls may be controlled individually, yet their outputs may be combined into one composite signal. A typical split-mixer network is shown in Fig. 9-18. Ten positions are shown split into two groups of five each. At the output of each group are building-out resistors (R1 to R8) and a submaster control P5 and P6 for controlling the output level of each group. The two sections of the network may be combined into a 10-position mixer or split by throwing the key switch S1. The output windings of the transformers T3 and T4 are run to separate channels or to a single recording channel.

9.19 ***What is a parallel-connected mixer network?*** — A configuration in which the mixer pots are connected in parallel as shown in Fig. 9-19. It will be noted that the building-out resistors R1 through R4 are connected in series with the output of each mixer control and an extra one (R5) is located at the input to the output coil. The purpose of building-out resistors is to provide an impedance match and isolation between pots. The low potential side of each pot is returned to a common ground connection at the low potential side of the output coil, T1. The purpose of the coil is to provide an impedance match to the line feeding the recording channel and to isolate the mixer ground from the recording circuits.

9.20 ***Show a schematic diagram for a series-connected mixer network.*** — A four-position series-connected mixer network is shown in Fig. 9-20. It will be noted the pots are all above ground. As the circuit is floating, serious leakage at the high frequencies can take place. This circuit has not been used for years and is not recommended.

9.21 ***Show a schematic diagram for a three-channel stereophonic dialogue***

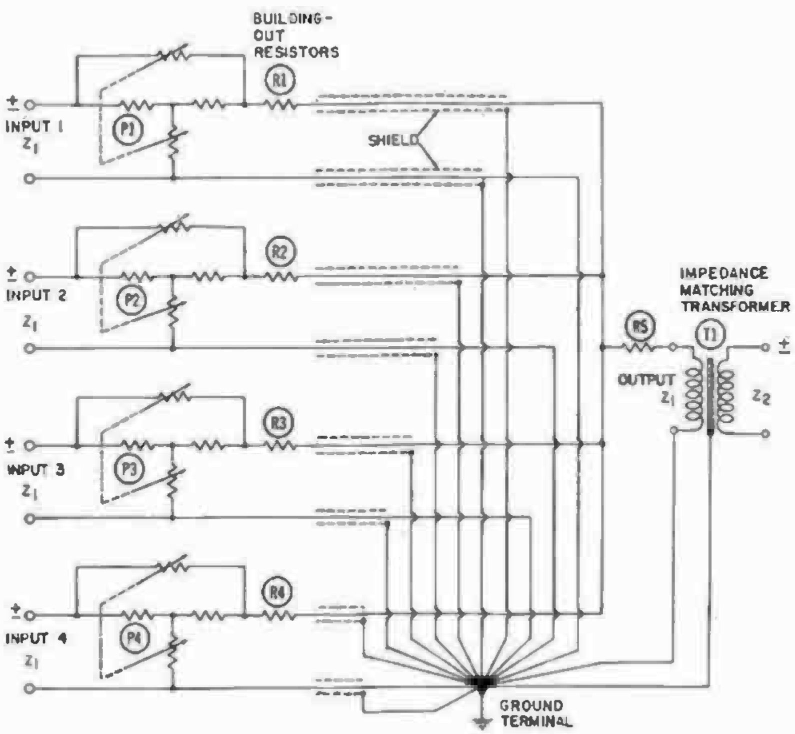


Fig. 9-19. A 4-position parallel-connected mixer network. The ground side of each mixer control is brought to a common ground point.

recording channel.—An elementary circuit diagram for a three-channel stereophonic mixer suitable for motion picture recording is shown in Fig. 9-21A. The mixer is split into three sections: left, center, and right. Each section consists of a conventional parallel-connected two-position mixer which feeds a recording amplifier connected to a magnetic recorder.

At the output of each recording amplifier is a resistive network which ties the three outputs together and feeds two monitor amplifiers connected to a pair of split headphones for the mixer. The left earphone is connected to the left channel and the right earphone to the right channel. Both earphones hear the center channel. Balancing pots are provided for balancing the sound level to each earphone.

Monitoring headphones are connected across the output of each recording channel for the microphone boom oper-

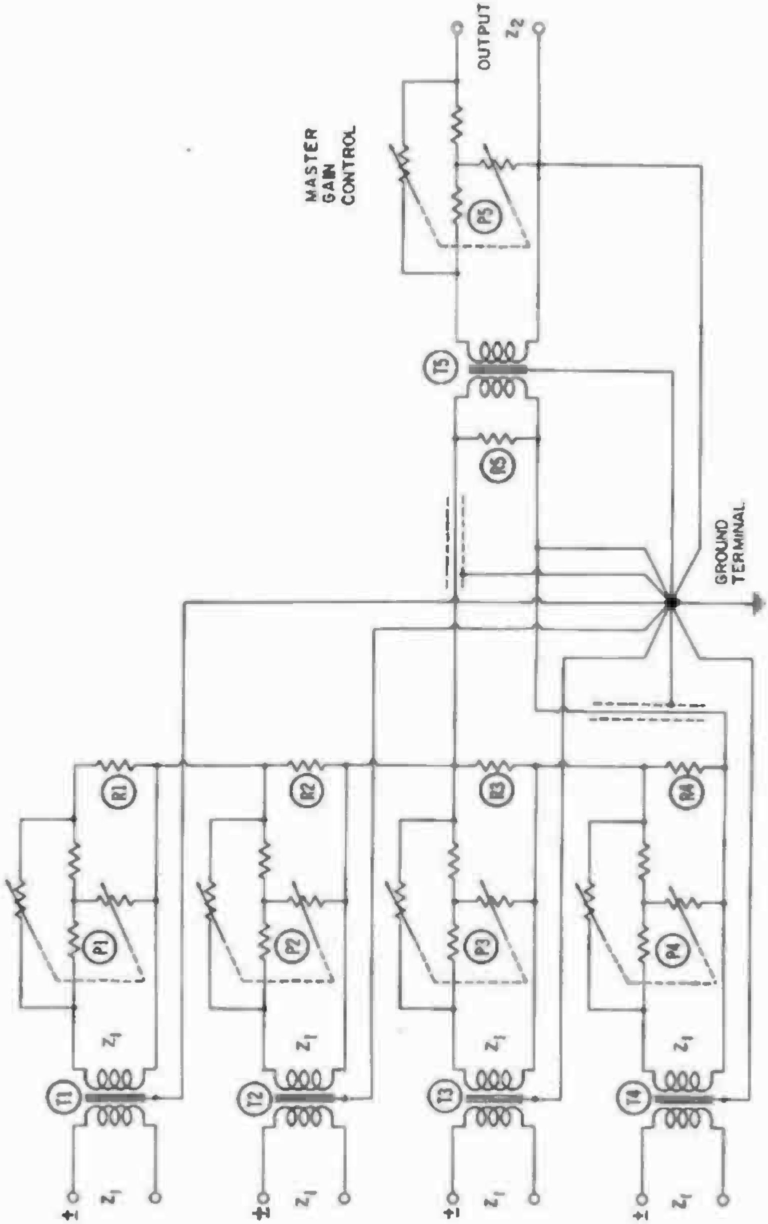


Fig. 9-20. A 4-position series-connected mixer network, not recommended for critical work.

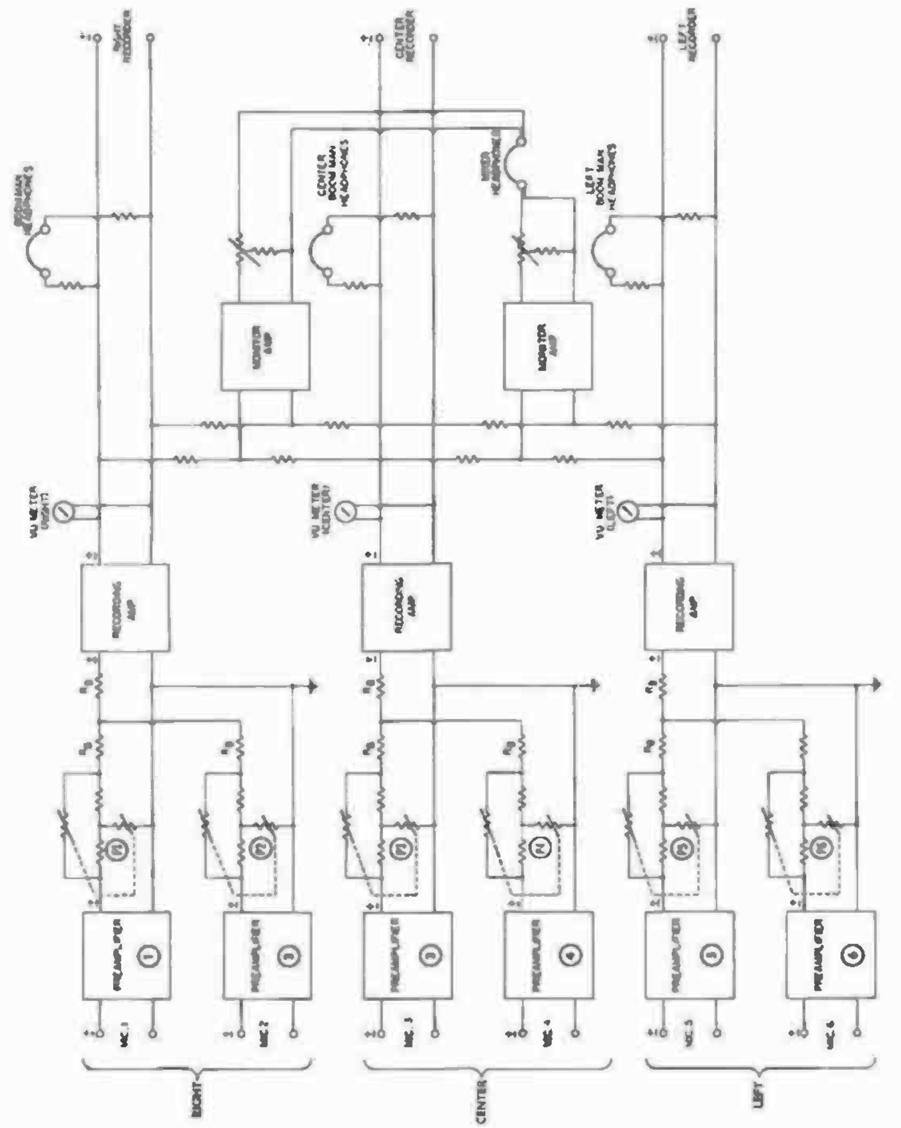


Fig. 9-21A. Simplified diagram for a 3-channel (6 microphone) stereophonic motion picture production mixer. Split headphones are used for the mixer, and individual headsets for the boom operators.

ator. It is highly important that all amplifying equipment be phased relative to the other channels, from the microphone to the magnetic recording head at the recorder. A typical stereophonic production mixer having six microphone inputs is shown in Figs. 9-21B and C.

It will be observed in Fig. 9-21C, that although the mixer controls are of vertical design, they are really two rotary controls ganged together. These controls consist of standard rotary attenuators operated by a system of cords and pulleys. The cords and pulleys may be viewed just below the edge of the panel. At the bottom of the case are microphone preamplifiers and recording and monitor amplifiers.

9.22 ***Show a block diagram for a four-position parallel connected stereophonic rerecording mixer using pan-pots.*** —An elementary block diagram for such a mixer network is shown in Fig. 9-22A. Although only four mixer controls are shown, any number may be connected in the network as long as the proper building-out resistors are used.

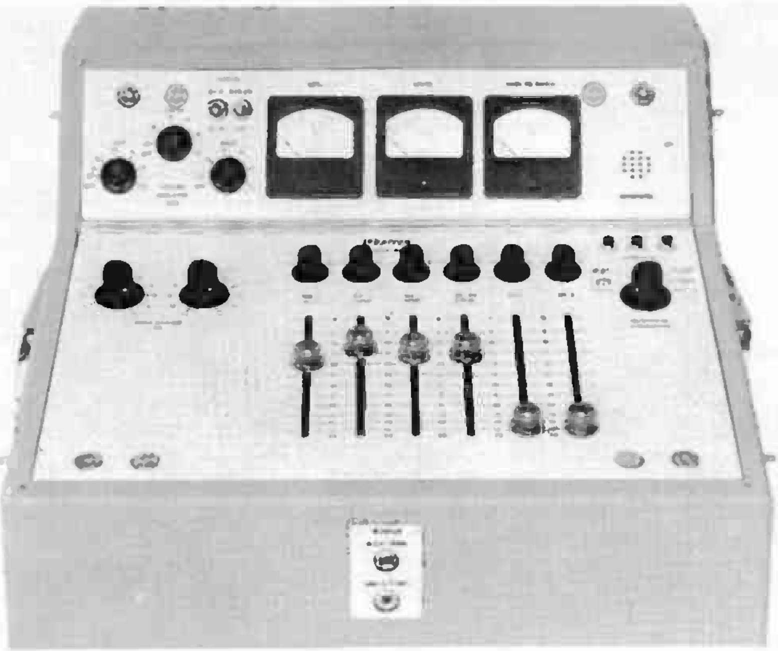


Fig. 9-21B. Exterior view of a Westrex portable stereophonic mixer.

The sound tracks to be spread are connected to the input of the regular monaural mixer network, P1 through P4. At the output of each mixer pot is a booster amplifier to compensate for the insertion loss of the pan pots.

Leaving the booster amplifier, the signal is fed to the input of a pan pot (described in Question 5.73). The output of each pan-pot is combined in a resistive network which feeds the input of three recording channels. The building-out resistors R_b, are to achieve an impedance match between the pan-pots.

The mixer network shown may be used for stereophonic rerecording using

Fig. 9-21C. Interior view of Westrex portable stereophonic production mixer.

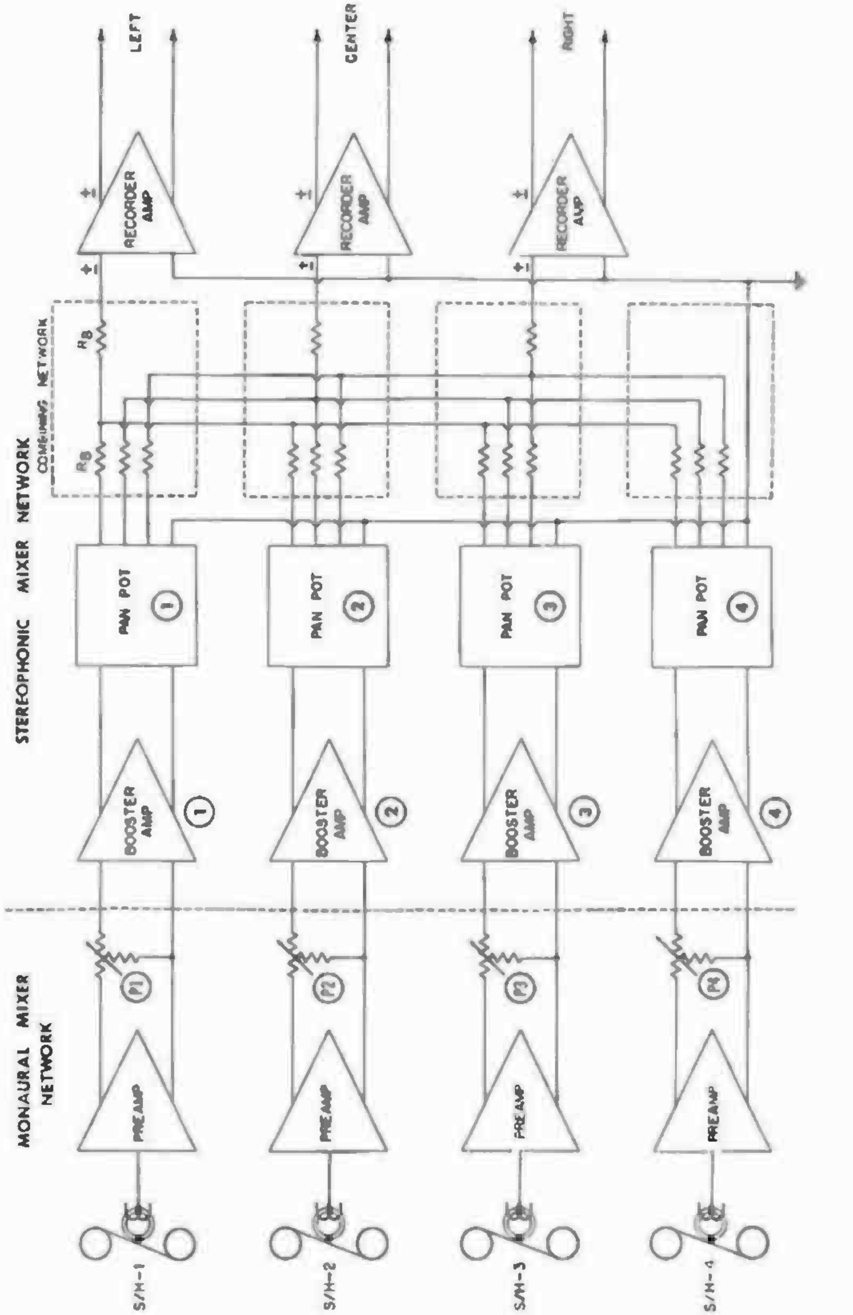


Fig. 9-22A. Simplified diagram for a 4-position pan-pot rerecording mixer network.

original stereophonic sound tracks, or monaural sound tracks. In the latter case, the stereophonic effect is called pseudostereophonic sound and is quite widely used in the motion picture industry.

A stereophonic rerecording mixer console built by RCA for 20th Century Fox Studios is shown in Fig. 9-22B. A similar type built by the sound department of Universal Studios, Universal City, California, is pictured in Figs. 9-22C and D.

The circuitry and controls for this console are typical of that required for rerecording stereophonic sound for motion pictures. Four groups of 8 vertical mixer controls are placed each side of

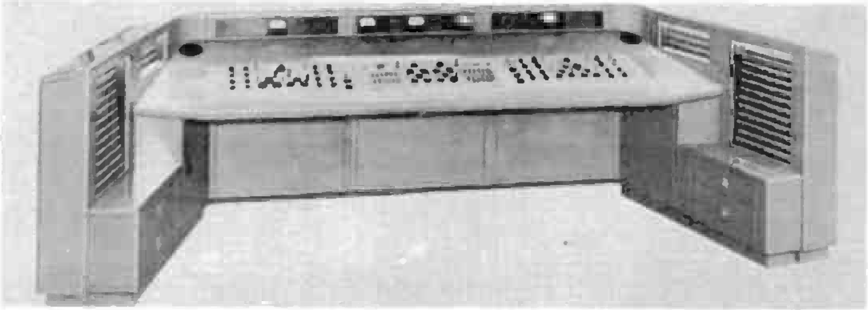


Fig. 9-22B. Stereophonic rerecording mixer console built for 20th Century-Fox Studios by RCA.

Fig. 9-22C. Stereophonic rerecording console at Universal Studios. (Courtesy Universal Studios Sound Department)

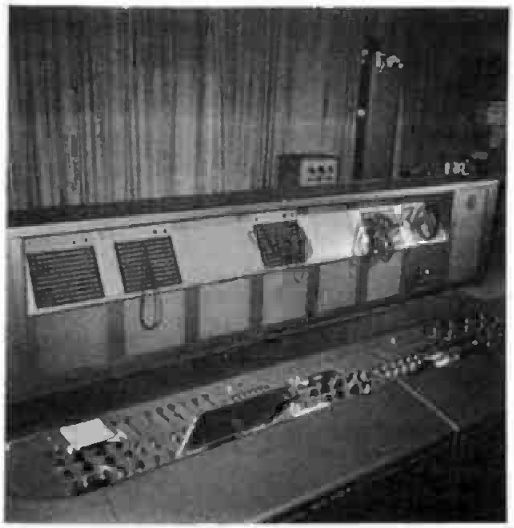


Fig. 9-22D. View taken looking over the console toward the patch bay. Here changes in the method of operation are made by the use of patch cords. (Courtesy, Universal Studios Sound Department) the center. The controls in the center section consist of pan-pots along with other controls for placing the sound in proper perspective. Three of the mixer-control groups have graphic equalizers, with additional equalizers and filters at the other positions. Many of the circuits are operated from push buttons for on-cue effects. No. VU meters are used at the console; instead, three *projection* type, VU meters about 30 inches in diameter, are placed on the floor below the screen.

At the rear of the console is a patch bay, containing 1360 tip, ring and sleeve type jacks. Here, the circuits may be connected by means of patch cords for different modes of operation. The patch bay has a total of 1360 circuits. At the lower right jack bay is a compressor amplifier, for controlling the peak level of dialogue.

9.23 *What is a hybrid-coil mixer?* —A network split into several sections each containing a group of mixer pots. The several sections are combined by means of hybrid coils, as discussed in Question 8.66. Hybrid coils are used in only the most elaborate mixers where unusual recording combinations are required. Typical uses for a hybrid coil mixer are for stereophonic rerecording, and recording large orchestras and choral groups. It is not unusual for such mixers to have up to 20 positions. A block diagram for a 16-position mixer network using hybrid coils appears in Fig. 9-23. Basically, the network consists of four separate, four-position, resistive networks (1 to 4).

Starting at network one at the upper left, the output of this network is connected to one side of the primary of hybrid coil T1. The output of network two is connected to the other side of the primary of the coil. The secondary of this coil feeds a booster amplifier having 30 dB gain. The output of the booster amplifier feeds a submaster control P1 for fading out all of the pots in the four-group mixer network simultaneously. The submaster control terminates in one side of the primary of hybrid coil T3 and from there, to the secondary and a master control P3, and thence to the recording circuits. Normal jacks are connected in all inputs and outputs, for testing and patching. Resistors R1, R2, and R3 are the balancing resistors normally used with hybrid coils. The bottom ends of these resistors are brought to a common ground point. Booster amplifiers 1 and 2 compensate for the insertion loss of the coils and other network components. If the recording circuits are unbalanced, an unbalanced pot may be substituted for the balanced control P3.

The signal levels indicated are based on an assumed signal level, at the input of the four-position mixer group, of a minus 30 dBm which is the output level from an average optical film sound head.

Each hybrid coil induces a loss of approximately 3 dB between the primary and the secondary. An additional loss of 6 dB takes place because of the two-group mixer panels at the primary. It is assumed that no loss will be carried in the sub- and master-gain controls. If the loss occurs, additional gain will be required. (See Question 8.66.)

9.24 *What is a building-out resistor?* —A noninductive resistor used for obtaining an impedance match in a mixer or combining network. The symbol for this resistor is R_{bo}.

Building-out resistors are also used in the output of low-impedance amplifiers to build out the internal output impedance to a given value. This subject is discussed in Question 12.142.

9.25 *What is the equation for calculating the value of a building-out resistor to a parallel-connected mixer network similar to that shown in Fig. 9-19?*

$$R_{bo} = \left(\frac{N-1}{N+1}\right) Z_1$$

where,

N is the number of mixer controls,

Z_1 is the impedance of the mixer control.

9.26 *What causes mixer-network insertion loss?* —Referring to Fig. 9-19, a fixed loss, termed *insertion loss* (IL) is created by the building-out resistors R1 to R5, the impedance-matching transformer T1, and any loss induced by the mixer-control configuration. Insertion loss is measured from the input of one control (control set to zero) to the output of T1. Insertion loss may be overcome by the use of an active mixer-combining network, as described in Question 5.98. (See Question 23.68.) The insertion loss for a mixer network employing ladder-type mixer controls will be 6-dB greater than for one using T-type controls.

9.27 ***How is the insertion loss of the parallel-connected mixer network shown in Fig. 9-19 calculated?***

$$\text{dB loss} = 20 \operatorname{Log}_{10} N$$

where,

N is the number of mixer positions.

9.28 ***How are the terminating resistors in the series-mixer network of Fig. 9-20 calculated?***—The terminating resistors R1, R2, R3, and R4 may be calculated:

$$R_T = \left(\frac{N+1}{N-1}\right) Z_1$$

where,

N is the number of controls,

Z_1 is the mixer-control impedance.

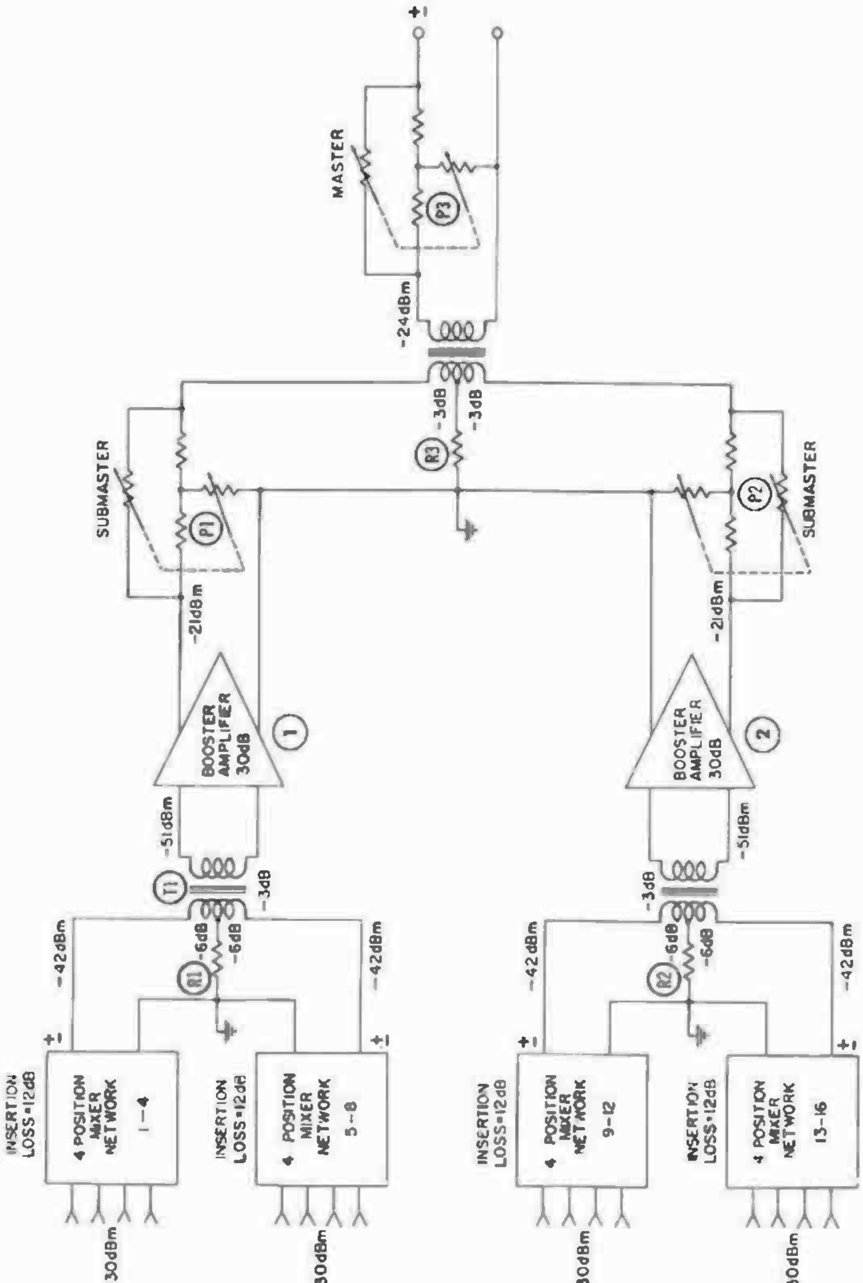


Fig. 9-23. A 16-position hybrid-coil mixer network. Losses are indicated for the coils and control networks.

9.29 What is the impedance at the output of the series-mixer network in Fig. 9-20?—The impedance from the top of the number one control to the bottom of the number two control is:

$$Z_2 = \frac{Z_1 N^2}{2N - 1}$$

where,
N is the number of mixer controls,
Z_1 is the mixer-control impedance,
Z_2 is the output impedance from the top of R1 to the bottom of R4.

Resistor R5 is equal to the output impedance of the four controls. A transformer of the correct impedance ratio may be substituted for the resistor, if desired. If the impedance at the output of the network is of an odd value, a taper pad may be substituted for the coil. The loss of the pad must be included in the insertion loss. Such pads are discussed in Question 5.34.

9.30 How is the insertion loss for a series-connected mixer calculated?

$$10 \text{ Log}_{10} (2N - 1)$$

where,
N is the number of mixer controls.

9.31 How are the balancing resistors for the hybrid-coil mixer network in Fig. 9-31 calculated?

$$Z_3 = \frac{Z_1 (N - 1)}{2}$$

where,
Z_3 is the balancing resistor,
Z_1 is the mixer-control impedance,
N is the total number of mixer controls.

Z_3 is a noninductive resistor. (See Question 8.66.)

9.32 How is the insertion loss of the hybrid-coil mixer network in Fig. 9-31 calculated?

$$\text{dB} = 10 \text{ Log}_{10} N$$

where,
N is the total number of mixer controls.

To this is added the loss of a ladder configuration (if used) and the transfer loss of the hybrid coil (3 dB). (See Question 8.67.)

9.33 How is the impedance ratio of primary to secondary calculated for a mixer-hybrid coil?

$$Z_R = \frac{2 \times Z_1}{Z_4}$$

where,
Z_1 is the impedance of the mixer controls,
Z_4 is the impedance of the hybrid-coil secondary.

9.34 Are plain-T type attenuators used in mixer networks?—Yes. However,

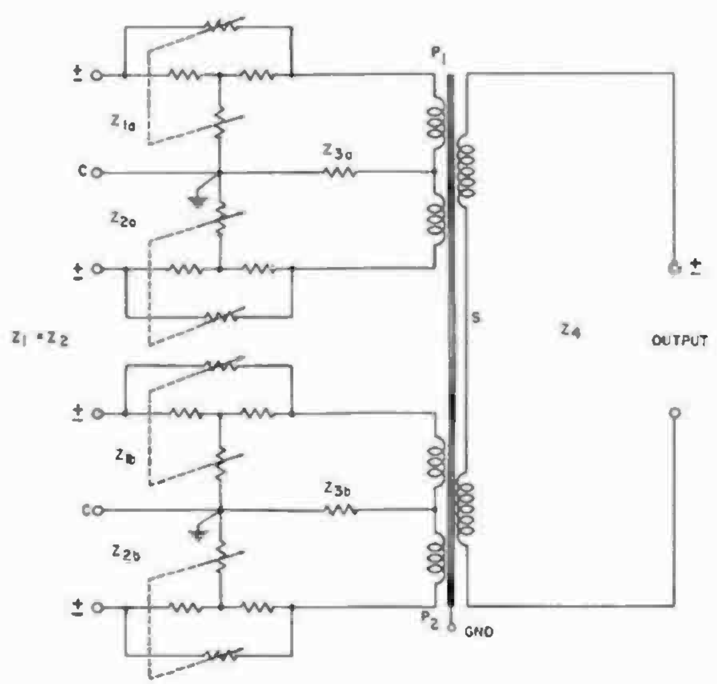


Fig. 9-31. Hybrid-coil mixer network.

as they require three rows of contacts, they have a higher noise level than the bridged-T type and therefore require a greater amount of maintenance. Ladder or bridged-T type mixer controls are recommended for commercial installations.

9.35 ***What is the recommended method of servicing a mixer control?***—If a mixer control becomes noisy while in operation, apply a signal of 40 to 60 Hz to the pot in question, and burnish the contacts (or slide-wire) by a rapid movement of the arm by wrapping a piece of cord around the control knob. The low-frequency signal supplies a signal voltage, yet it is of a low enough frequency that the noise of the contact may be easily heard. A light clear mineral oil such as Nujol may be applied to the contact surfaces to prevent oxidation of the surfaces. *Never use an abrasive on the contact surfaces.*

9.36 ***What is meant by phasing a mixer?***—Phasing a mixer means polarizing the mixer inputs and output so that they are electrically in phase with each other.

The phasing of a mixer network is quite important, particularly if the mixer is to be used for stereophonic or photographic film recording. Optical film-recording systems must be phased from the microphone inpnt to the light modulator unit.

When a pressure wave is applied to a microphone, the diaphragm moves inward. At the same instant the light modulator is deflected and the noise reduction equipment moves toward the maximum cancellation point. Because the human voice is unsymmetrical, the system is phased to prevent the clipping of the modulation peaks by means of the noise reduction equipment. Therefore, the film-recording system must always be in phase for correct operation.

It is good practice with any type recording and reproducing system to observe proper phasing of the various components in the system. This is particularly true for photographic sound track transfer channels. Phasing of multiple-channel mixer networks is discussed in Section 23. Microphone phasing is discussed in Questions 4.84 and 4.85.

9.37 ***What method is recommended for grounding mixer networks?***—The ground side of each mixer control is brought to a common ground point, as shown in Fig. 9-19. Individual twisted shielded pairs are used to connect each mixer control. If balanced-to-ground input circuits are to be used with this network, a repeat coil must be connected between the signal source and the input of the mixer control to isolate the grounding systems.

All interconnecting wiring in the interior of the mixer inclnding the ground wires must be shielded, and the shields returned to a common point. If the shields are covered with cloth braid, the shield is grounded at one end only. If the shield is bare and can make contact with the mixer case, it must be honded together every few inches and securely bonded to the case at short intervals.

9.38 ***When key switches are used in mixer networks, how are clicks prevented when the circuits are broken?***—By the use of a transmission ground system. The switches are of the make-before-break type. If a mixer control is switched out of the circuit, a terminating resistor equal in value to the impedance of the control must be connected in the circuit to maintain the proper impedance relations.

9.39 ***What are low- and high-frequency attenuators?***—Reactive circuits connected ahead of the mixer control for the purpose of attenuating either the low- or the high-frequencies. As a rule, they are set for a characteristic that conforms to the standards used in the motion picture industry. The circuits for such attenuators appear in Fig. 9-39. It will be noted that low-frequency attenuation is secured in the upper control by connecting capacitors in series with the high side of the mixer-control input. Two positions are provided, −8 dB and −12 dB at 100 Hz. When the capacitors are in the circuit, 100 Hz is attenuated the indicated amount with respect to 1000 Hz. High-frequency attenuation is obtained by shunting capacitors across the line. Three positions are provided: −4 dB, −8 dB, and −12 dB at 10,000 Hz, with respect to 1000 Hz.

For purposes of illustration, the high-frequency attenuation in the lower control is obtained by the use of a series resonant circuit. A resistor is shunted around the coil to lower its Q and shape the response curve. High- and low-

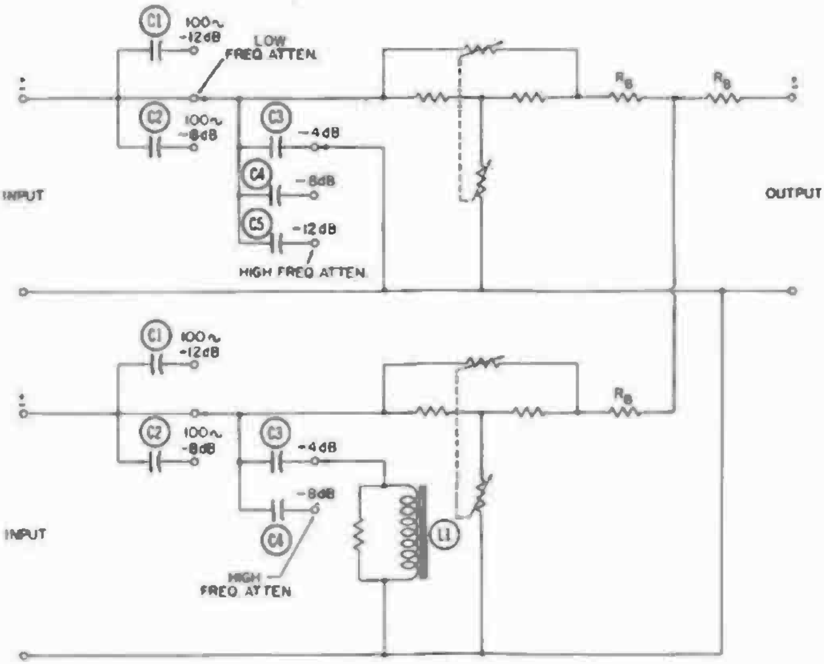


Fig. 9-39. A 2-position mixer.

frequency attenuators are discussed in Questions 6.80 through 6.84.

9.40 *What is the insertion loss of a mixer network when using ladder mixer controls?*—The insertion loss is 6 dB greater than when using plain or bridged-T controls. This loss is due to the configuration of the ladder pot and is independent of the loss setting of the control. Ladder pots are discussed in Question 5.51.

9.41 *How is the insertion loss of a mixer measured?*—This subject is discussed in Question 23.68.

9.42 *Describe a solid-state sound reinforcement and public address mixer-amplifier.*—Sound mixers designed for high-quality sound reinforcement or public-address service generally require a minimum of six positions, and up to 20 positions is not uncommon. A mixer-amplifier of solid-state design, manufactured by the Langevin Co., is shown in Fig. 9-42A. Although this particular mixer has only 6 positions, another 12 positions may be added by the addition of an auxiliary mixer network of Fig. 9-45. The schematic diagram for

Fig. 9-42A. Model AM1A solid-state, 6-position mixer manufactured by the Langevin Co.

the unit of Fig. 9-42A is shown in Fig. 9-42B.

At the left are six microphone inputs, feeding six plug-in, module-type preamplifiers using silicon planar transistors, with individual speech-music equalizer switches in their outputs. From there, the signals are combined into a resistive mixer network and to a booster amplifier, variable equalizer, a second booster amplifier, master gain control and a line amplifier. Provision is

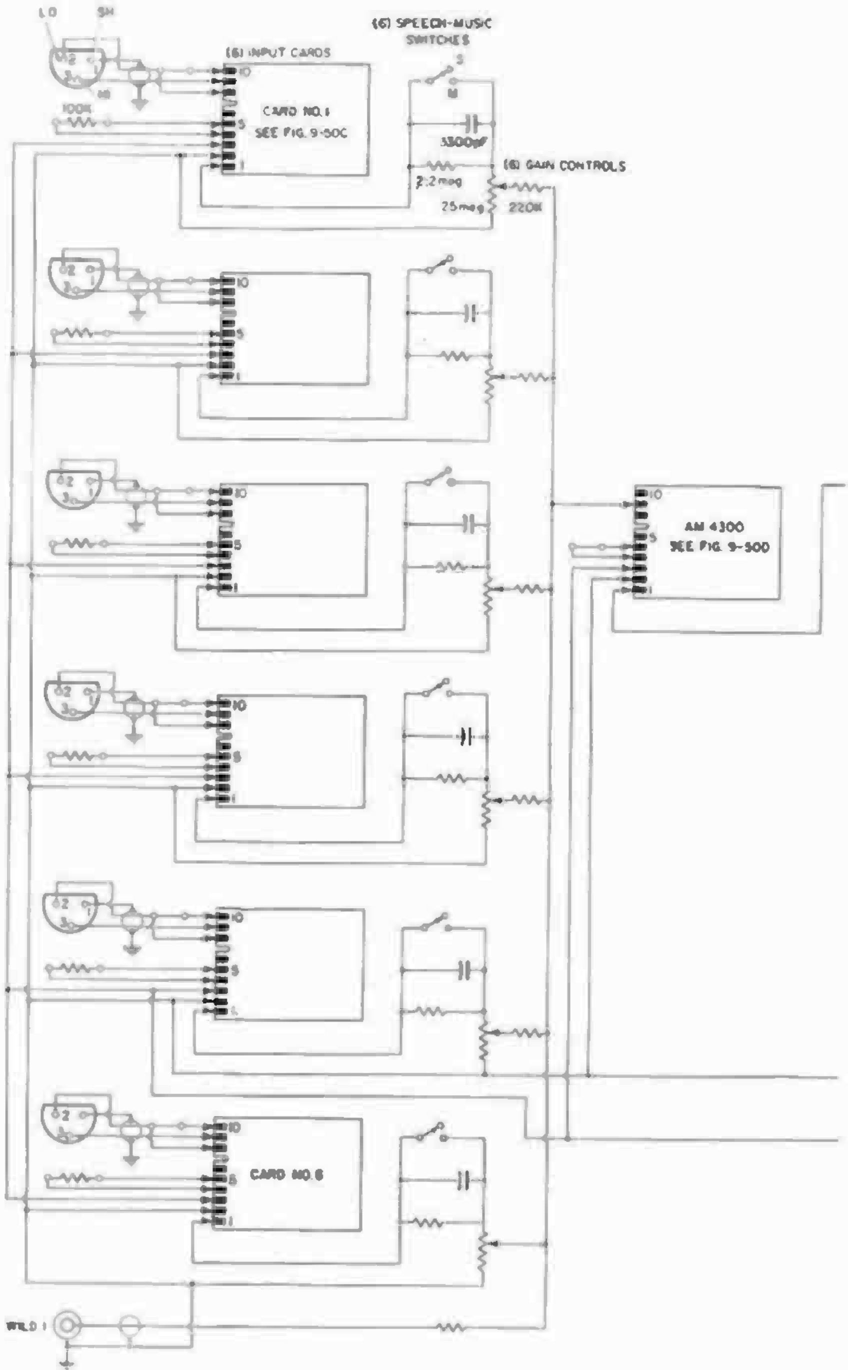
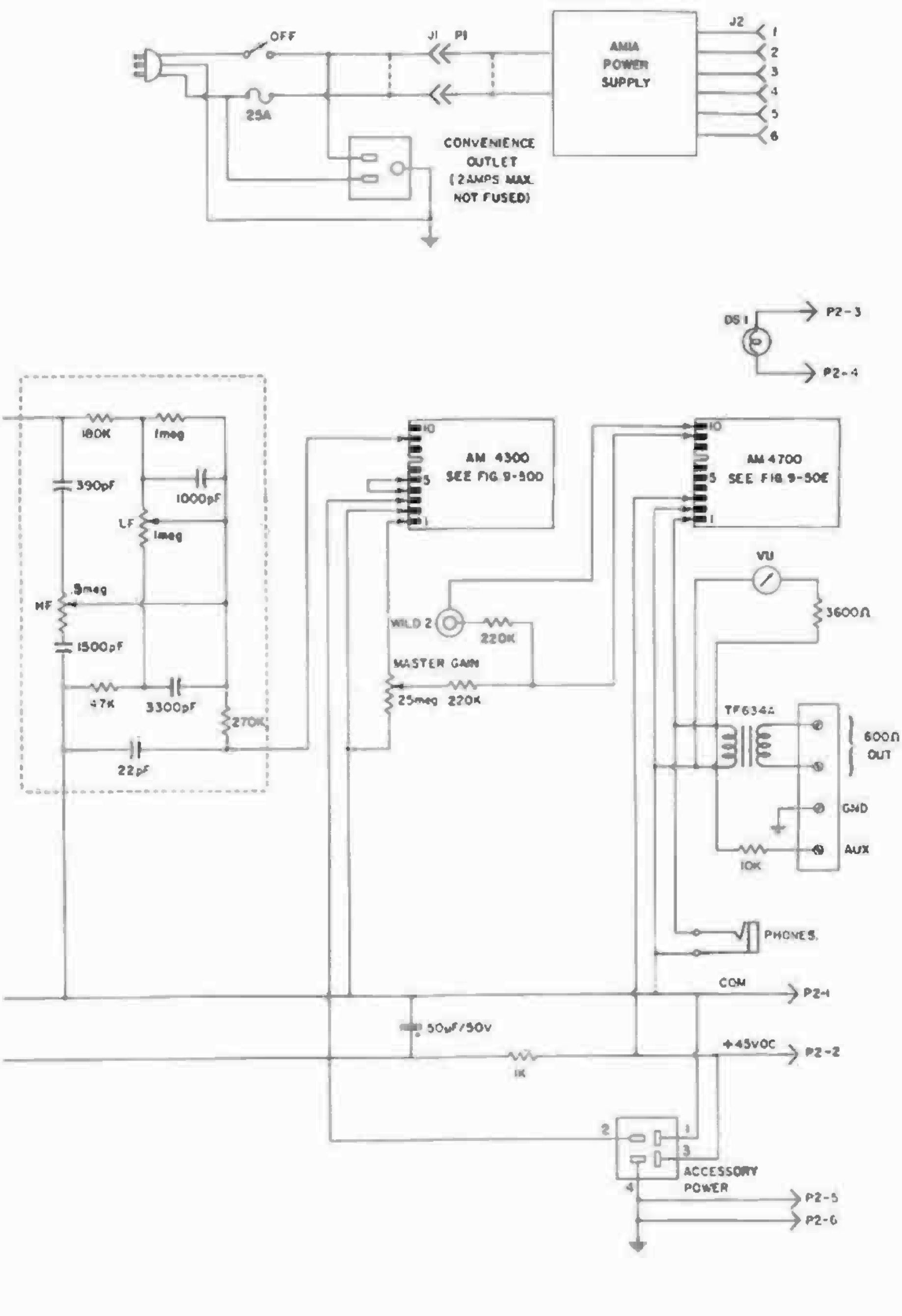


Fig. 9-42B. Schematic diagram for

made for headphone monitoring. A regulated solid-state power supply is also contained in the mixer housing. The frequency response is plus-minus 1 dB, 30 to 15,000 Hz, with less than 0.25 percent total harmonic distortion. The equivalent input noise is minus 112 dBm (60 dB signal-to-noise ratio). Total gain is 88 dB, maximum output power is plus 18 dBm.

The internal circuitry for the Model AM-4100 preamplifier, the Model AM-



Langevin AM1A mixer amplier.

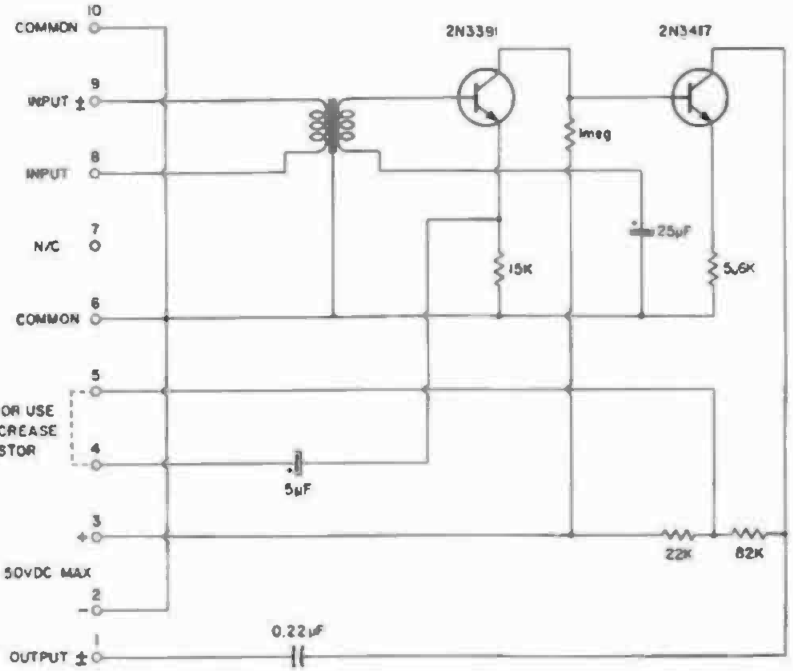


Fig. 9-42C. Langevin Model AM4100B solid-state amplifier for AM1A mixer-amplifier.

4300B booster amplifier, and the Model AM4700 output amplifier, is shown in Figs. 9-42C, D, and E.

9.43 ***What is an electronic mixer?*** —This is a term which is sometimes used to identify a compressor amplifier employed for motion picture recording. Electronic mixers are discussed in Questions 18.84 through 18.102.

9.44 ***What are the values of build-***

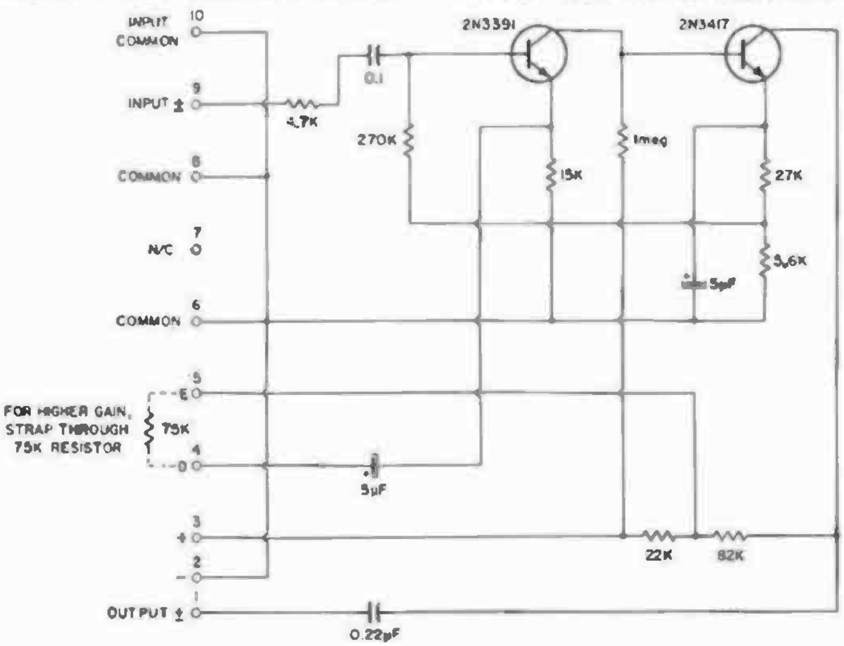


Fig. 9-42D. Langevin Model AM4300B solid-state amplifier for AM1A mixer-amplifier.

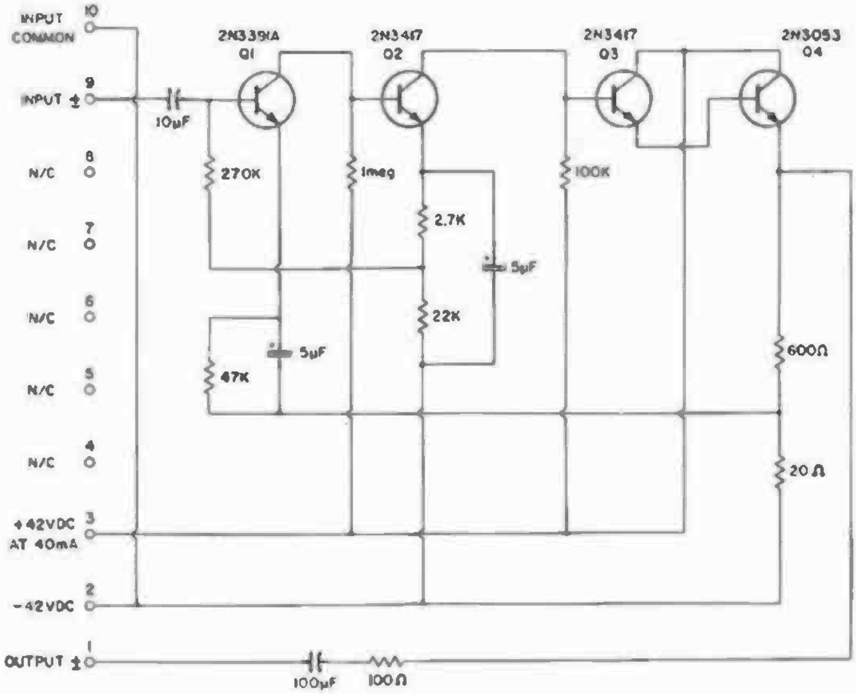


Fig. 9-42E. Langevin solid-state amplifier Model AM4700 used in the Model AM1A mixer-amplifier.

ing-out resistors used for parallel-type mixer networks?—The values of the building-out resistors and the insertion loss for mixers using a parallel mixer network (Fig. 9-19) are given in Fig. 9-44. The number of mixer positions appears at the left and the impedance of the network appears at the top of the chart.

The insertion loss is the loss caused by the building-out resistors, with only a single input at its minimum attenuation. The measurement of insertion loss is discussed in Question 23.68.

9.45 *Describe an auxiliary mixer.*—The auxiliary mixer pictured in Fig. 9-45 is manufactured by Langevin Co., and is designed to be operated as an auxiliary 9-position mixer-amplifier with any existing mixer to provide additional input positions. The circuitry of the unit is quite similar to that of the mixer described in Question 9.42, and is the same up to the booster amplifier. The unit is self-powered and contains the necessary preamplifiers and booster amplifier, which is operated directly into a low-gain position of the master mixer.

9.46 *Describe the mechanical layout and components for a three-section rerecording mixer console.*—Pictured in Fig. 9-46A is a three section rerecording console, designed by the author, and constructed by McCurdy Radio Industries, Canada. This console is designed for use with three-track magnetic film rerecorders or any type single track recorder. Basically, the console consists of three sections, termed dialogue, effects, and music. Each section consists of four mixer controls and a master control, plus two graphic equalizers, high- and low-pass filters, and two low-frequency attenuators. One or both of the graphic equalizers may be connected in any one of the four positions by means of push buttons above the control panel. In addition, key switches are provided for on-cue operation. The low-frequency attenuators provide standard roll-off characteristics as discussed in Questions 6.122 and 6.131. Graphic equalizers are discussed in Questions 6.108 and 6.126. The high-pass filters employ cutoff frequencies of 45, 80, 100, 135, and 150 Hz. The low-pass cutoffs are at 6, 8, and 10 kHz. These may also be cued in as required. The usual practice is to leave in the 45-Hz filter for rerecording, and use the 8-kHz low-pass filter for normal dubbing operation. For original music, the 10-kHz filter is used.

The dialogue position varies in its components, as an additional program equalizer, discussed in Question 6.12, has been added to the upper right, along with three controls for an EMT reverberation unit (installed at a remote point) described in Question 2.130. A remote meter indicates the reverber-

Mixer Positions	Impedance 150Ω			Impedance 200Ω			Impedance 250Ω			Impedance 500Ω			Impedance 600Ω		
	R_n Ohms	Insertion Loss—dB		R_n Ohms	Insertion Loss—dB		R_n Ohms	Insertion Loss—dB		R_n Ohms	Insertion Loss—dB		R_n Ohms	Insertion Loss—dB	
		BT, T	Lad.		BT, T	Lad.		BT, T	Lad.		BT, T	Lad.		BT, T	Lad.
2	50.0	6.00	12.00	66.6	6.00	12.00	83.3	6.00	12.00	166.0	6.00	12.00	200.0	6.00	12.00
3	75.0	9.50	15.50	125.0	9.50	15.50	100.0	9.50	15.50	250.0	9.50	15.50	300.0	9.50	15.50
4	90.0	12.00	18.00	120.0	12.00	18.00	150.0	12.00	18.00	300.0	12.00	18.00	360.0	12.00	18.00
5	100.0	14.00	20.00	133.0	14.00	20.00	166.0	14.00	20.00	333.0	14.00	20.00	400.0	14.00	20.00
6	107.0	15.60	21.60	142.8	15.60	21.60	178.5	15.60	21.60	357.0	15.60	21.60	428.4	15.60	21.60
7	112.5	16.90	22.90	150.0	16.90	22.90	187.5	16.90	22.90	375.0	16.90	22.90	450.0	16.90	22.90
8	116.6	18.10	24.10	155.5	18.10	24.10	194.0	18.10	24.10	388.0	18.10	24.10	466.0	18.10	24.10
9	120.0	19.08	25.08	160.0	19.08	25.08	200.0	19.08	25.08	400.0	19.08	25.08	480.0	19.08	25.08
10	123.0	20.00	26.00	163.8	20.00	26.00	202.2	20.00	26.00	409.0	20.00	26.00	491.0	20.00	26.00
11	124.9	20.82	26.82	166.5	20.82	26.82	208.0	20.82	26.82	416.0	20.82	26.82	499.0	20.82	26.82
12	127.0	21.58	27.58	169.1	21.58	27.58	210.8	21.58	27.58	422.5	21.58	27.58	507.0	21.58	27.58

Fig. 9-44. Values for building-out resistors R_n, and the insertion loss for parallel mixer networks.

Fig. 9-45. A 9-position auxiliary mixer for use with the Langevin AM1A mixer-amplifier.

Fig. 9-46A. Three-section rerecording (dubbing) console, designed by H. M. Tremaine and constructed by McCurdy Radio Industries, Canda.

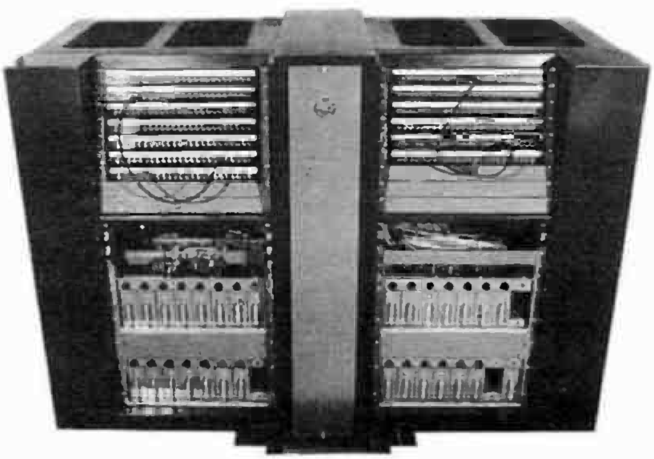
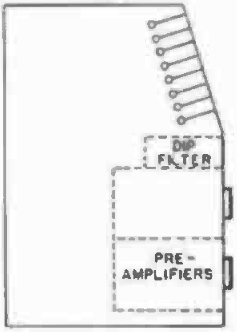


Fig. 9-46B. Patch-bay showing 32 plug-in amplifiers. Background amplifier and dip filter are not shown.



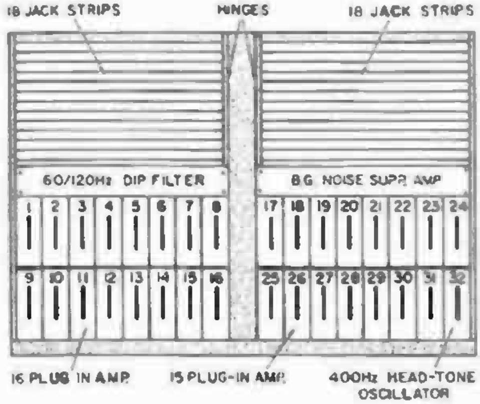


Fig. 9-46C. Patch-bay for console in Fig. 9-46A. All amplifiers, jacks strips, dip-filter, and background-suppression amplifier are housed in this unit. Its the same height as the console. The left and right sides are hinged at the center post to swing forward.

ation time being used, for future reference. At the upper left is a group of on-the-line machine indicators to assure the mixer that all the reproducers and recorders are on the line. The machine indicators light automatically whenever a machine is connected across a distributor bus. A group of intercommunication key switches at the lower left of each section permits the mixer to slate the head end of each take, talk to the machine room, maintenance department, power room, test laboratory, and the music scoring console (in an upstairs monitor room). The intercommunication system is solid-state, installed at a remote position.

It is the practice in the motion picture industry when recording master sound tracks, to record 400-Hz head tone on each roll of magnetic tape or film. This tone is used by the mixer to establish the correct listening level for playback, and later is used by the transfer activity. The output of the 400-Hz oscillator has three separate controls, one at each section. If the manual button is pushed, the tone appears in only the section where the button is depressed. If the button is pushed to automatic in any of the three sections, the tone is automatically recorded on the three sound tracks for a duration of six seconds, starting at the dialogue section. If three tracks were recorded simultaneously, the level at the composite bridging bus would be in error by several dB. The small diagonal panel at the upper left contains a monitor gain control and a 20-dB talk key for reducing the monitoring level while the sound track is running and discus-

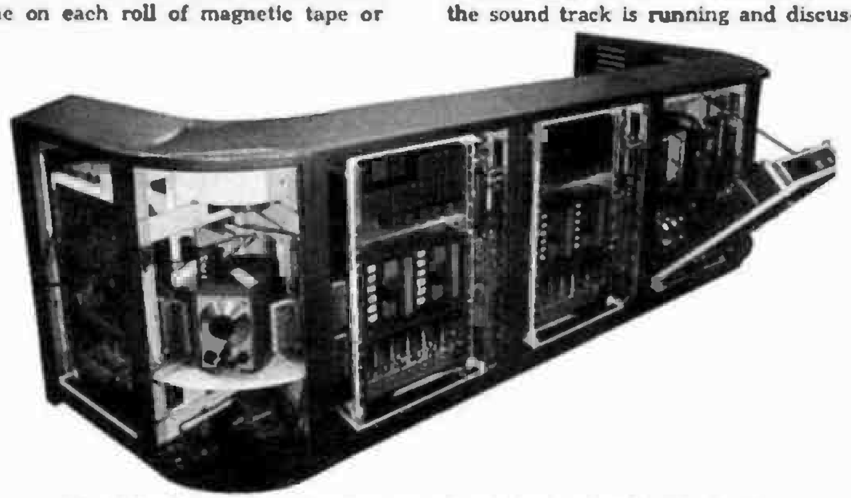


Fig. 9-46D. Rear view of mixer console similar to that in Fig. 9-6A.

sions are in progress. In the panel at the lower left are three controls for a compressor-limiter amplifier, with input and ceiling controls and a de-essing key switch. The playback mode is set by the position of a selector switch, with pilot lights to indicate that the machine is being used. If the recording levels are correct, the playback level will be within plus-minus 0.5 dB, using the head tone as a reference. Two variable sound-effects filters at the bottom of the panel complete the controls in this area. (See Question 18.84.)

The fact that the three sections of the console are termed dialogue, effects, and music, does not mean they are confined to this particular type recording. This terminology has been adopted to identify the sections and is the order in which the three sound tracks appear on a three-track magnetic film recorder. All three sections have the same identical electrical characteristics, except for the compressor amplifier in the dialogue channel.

A remote-control panel at the left side of the console (not shown) contains switches for controlling the house lighting, ventilation, quiet lights, baby spot above the console, dubbing and projection mode, screen curtains and a control permitting the loudspeaker behind the screen to be operated in conjunction with a similar speaker in the music scoring monitor room, or separately. This control is duplicated in the projection booth and at the music-mixer console. Between the second and third VU meters are push buttons for activating a footage counter (mounted at the screen) and resetting to zero. A reverberation time meter is mounted between VU meters three and four, for resetting the reverberation unit, and a push button for measuring the monitor level, using the effects VU meter. All relay control and heater voltages for the amplifiers are supplied from 1 percent regulated dc power supplies located in a remote power room.

The rear of the console houses the compressor amplifier, high- and low-pass filters, relays, pads, and other equipment necessary to its operation. The patch bay is mounted in a steel cabinet directly behind the console, and contains 18 jack strips, with 432 active circuits, using tip, ring, and sleeve type jacks (Fig. 9-46B). It also contains a dip filter for 60 and 120 Hz, and a background suppressor amplifier. The wiring between the console and the patch-bay is carried in a section of 6-cell metal gutter laid in the floor. A similar gutter is described in Question 24.49.

Although the patch bay houses 31 amplifiers, only 22 are directly associated with the mixer circuitry. The others are for microphones, spares, and special setups. The 400-Hz oscillator is also housed in this area, and slides into the amplifier tray at the extreme lower right. When a microphone is required such as for looping, a spare amplifier is patched into the desired input. In this instance, the mixer uses headphones for monitoring. Looping is discussed in Question 17.223. In the drawing of Fig. 9-46C, is shown the mechanical layout of the patch bay. A rear view of a similar console is pictured in Fig. 9-40D. When the rear panels are removed, the equipment falls forward on a frame to facilitate servicing. The placement of a console on a dubbing stage is shown in the drawing of Fig. 2-76D.

9.47 Describe the block diagram for the three-section mixer console of Question 9.46.—Basically, the circuitry (Fig. 9-47A) is divided into three sections, each consisting of four mixer controls, amplifiers, combining networks, equalizers, filters, bridging bus and VU meter. The output from each bridging bus is fed to one track of a three-track magnetic film recorder. For monitoring purposes and for feeding of single-track recorders a composite bridging bus is used, fed from the combined output of the dialogue, effects, and music-bridging busses. A fixed playback system controlled by relays permits the mixer to check the recording levels, in reference to a head-tone recorded at the head end of each roll of magnetic film. To simplify the drawing, only one input is carried through for each group of four mixer controls. Up to the mixer-combining network, each input is identical.

Starting at the extreme upper left are the input circuits for the four controls of the dialogue section. The signal may be from any source having an output level of minus 4 dBm or greater. From the input line, the signal passes through a push-button controlled graphic equalizer, then through a preamplifier (1 to 4), isolation coil (1:1 repeat coil), and to the mixer control. Leaving the control, it passes through a low-frequency attenuator, mixer-combining network, a second isolation coil, booster amplifier (5) where the signal is again amplified and passed to a master mixer control (isolated at its input and output by coils—normal setting zero loss), then to a compressor-limiter amplifier (see Question 18.86). At the input and output of the compressor are two controls, one for setting the correct input level, and a ceiling control for adjusting the amount of compression. Leaving the ceiling control, the signal encounters high- and low-pass filters. At this point, push-button switches are provided for connecting an EMT reverberation unit (described in Question 2.130) into the circuit. Amplifiers (29) and (30) when once set require no further adjustment. The reverberation is available only in the dialogue section; however, it may also be patched in at either the effects or music sections.

At the output of the low-pass filter, the signal is passed to a combining network, where a slating microphone can be keyed into the dialogue channel when required, or the playback signal injected, or the head-tone oscillator signal injected. Normally relays K_1, K_2 and K_3 are in the upper position, supplying a termination for the network. The output of the network feeds the signal to dialogue line amplifier (6), and then to the dialogue bridging bus and VU meter.

From the bus, the signal is amplified by a bridging amplifier (7), which terminates in a three-branch monitor-combining network, and to a composite

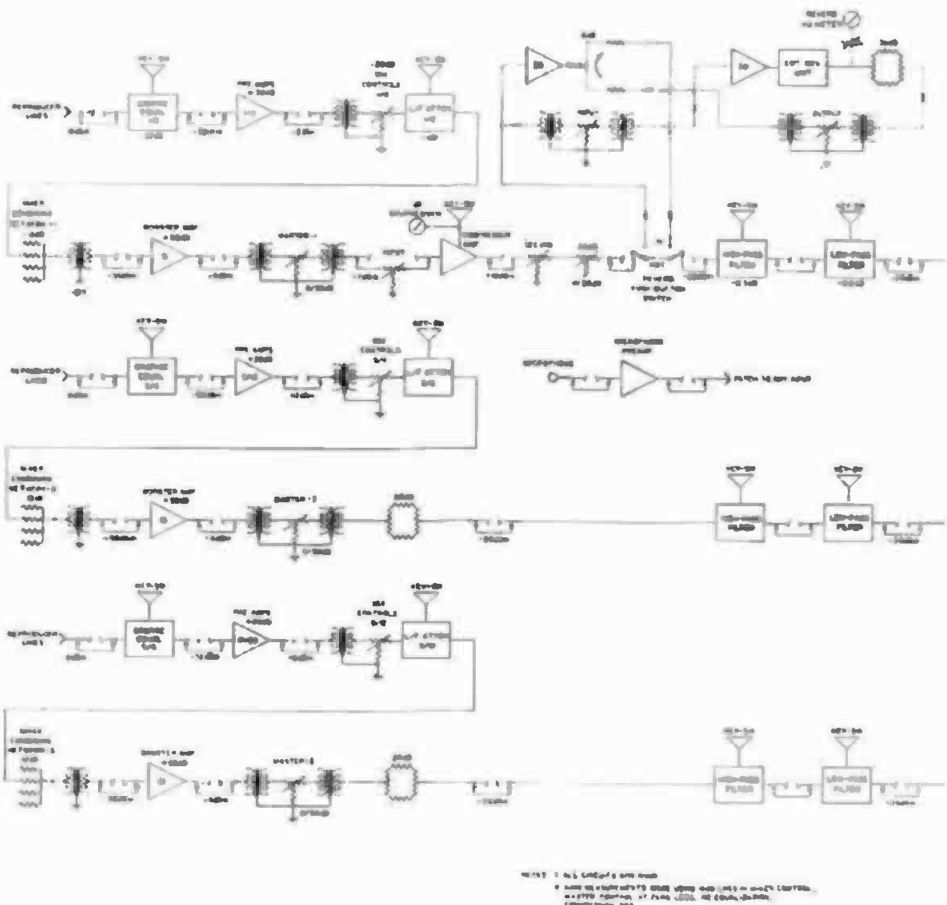
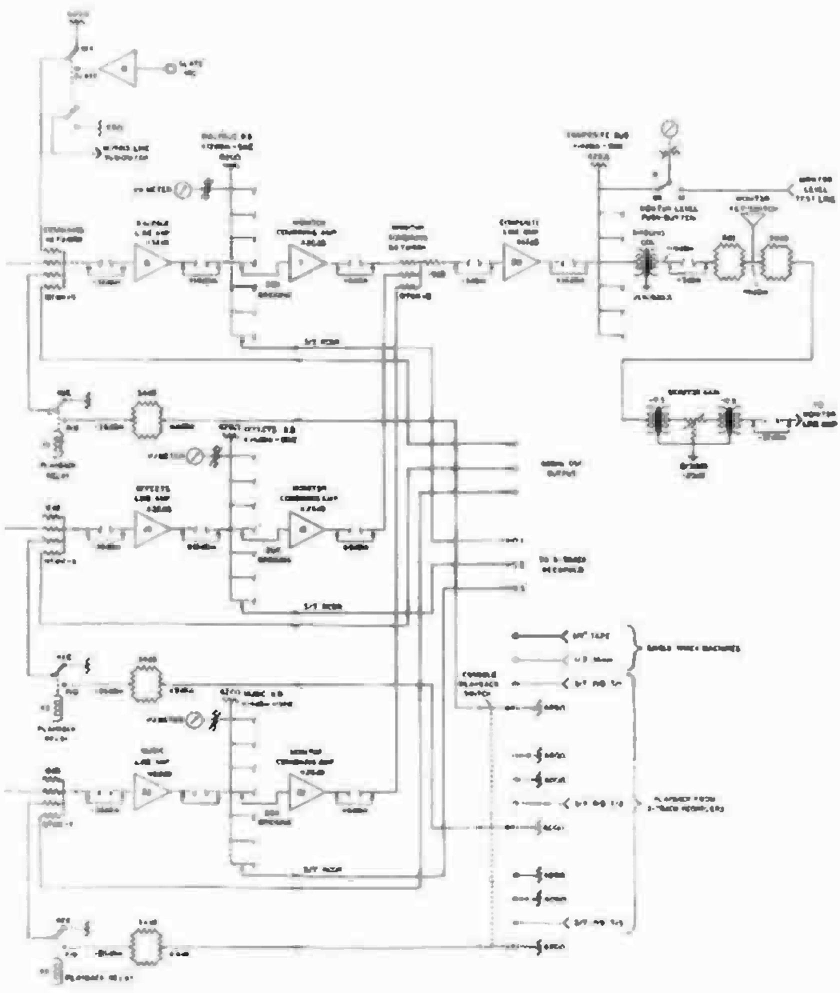


Fig. 9-47. Block diagram for 3-section split-

bridging bus amplifier (28), which drives the composite bridging bus and VU meter. The composite bus feeds a bridging coil which terminates in a 6-dB and 20-dB pad to reduce the monitor level while conversing through a key-switch control. The output of the pad is taken to a monitor gain control and then fed to the monitor system—a 50-watt power amplifier in the projection booth. To complete the monitor circuit, a signal is taken from the effects and music bridging busses and fed to two other branches of the monitor-combining network and amplified by the composite line amplifier (28), then the signal is passed on to the composite bridging bus.

The other two sections—effects and music—are identical to the dialogue section up to the output of the master controls. Here a 20-dB pad is connected in these two sections at the input of the high- and low-pass filters. From then on, the circuitry is the same as for the dialogue section, and the circuits feed their respective tracks for three-track recording. Single-track machines are fed from the composite bridging bus.

The gain through the whole system must be balanced within plus-minus 0.5 dB, particularly at the input to the combining network at amplifier (6). The system is lined up by applying a signal to the input or mixer control (1) in the dialogue channel, with 16 dB of loss in the control and the preamplifier, and the booster and line amplifiers adjusted for plus 14 dBm at the three bridging busses. For recording, the VU meters are set for an 8-dB lead by



channel mixer console in Question 9.46.

setting their attenuators back to plus 6 dBm. This is to take care of the unseen peaks that could overload the system while recording. The adjustment of the amplifier gains will vary somewhat from the indicated levels with different equipment, therefore they cannot be taken as absolute value.

When the playback relays are in the playback position, the playback signal is passed through a 34-dB pad before being applied to the combining network. The loss of this pad may have to be altered, depending on the output level from the playback machine. When the 400-Hz head-tone is recorded, the signal passes through the same combining network and is recorded on each sound track for 6 seconds, in its proper sequence, by a cam-driven relay system. In the off position, the oscillator output is shorted to prevent leakage; this circuitry is not shown.

Although the circuitry is for vacuum-tube type amplifiers, transistor types may be used if the gain and impedance matching is observed (see Question 9.45). The gain of the amplifiers should be divided to keep the signal-to-noise ratio high. The frequency response for this particular console is, plus-minus 1 dB, 40 to 15,000 Hz, with a total harmonic distortion (THD) of less than 0.5 percent at any frequency between 40 and 10,000 Hz. The signal-to-noise ratio is 65 dB, or 79 dB below a bus level of plus 14 dBm. Crosstalk between the three sections is greater than 85 dB at 10,000 Hz. All transmission circuits are symmetrical employing "O" type pads throughout. The several isolation coils permit the use of unbalanced mixer-network circuits and controls, with symmetrical input and output circuits. This also provides a means for grounding the various sections to a common ground point, thus reducing the possibility of circulating ground currents. All input and output circuits are 600-ohms impedance. The monitor system is discussed in Question 9.48.

It is of extreme importance that the three sections of the console be in *phase* with each other, to prevent cancellation between the three sections for certain conditions of operation. Phasing of this and other type mixer networks is discussed in Questions 23.106 and 23.107.

It is the policy of some recording activities to patch an optical film recorder across the composite bridging bus and record an optical negative along with the master dubbing track to save transfer time. The magnetic sound track is then held for protection and playback purposes. If the optical recording is to be variable-area direct-positive, it may be advantageous to employ a *cross-modulation compensator-amplifier* with the film recorder, a *film loss equalizer*, and if possible a *limiter-amplifier* to prevent overmodulating the light modulator in the optical film recorder. Cross-modulation compensating amplifiers are discussed in Question 18.246.

9.48 Describe the monitor system and associated equipment for the console in Question 9.46.—Leaving the composite bridging bus of Fig. 9-47, and following the monitor line to the upper left of Fig. 9-48, the monitor signal is amplified by a monitor line amplifier, and fed to a projection changeover relay panel, used for selecting the mode of operation—projection or dubbing. With the relay in the dubbing position, the signal is applied to the input of a monitor compensator panel, consisting of a 10-dB pad and a low-pass filter. The characteristic of this filter is representative of the average theater. Components in the panel provide adjustment for the characteristics of the dubbing stage acoustics, if required. At the output of the compensator panel is a three-branch combining network for driving the input of two 50-watt power amplifiers, one for driving the dubbing loudspeaker behind the screen, and the second for driving a duplicate speaker system in the music scoring monitor room. The output of the amplifiers feeds a speaker selector relay panel for playing the two speakers simultaneously or individually. The selector panel may be operated either from the projection booth or at the console.

From the selector panel, the signal is fed to a loudspeaker crossover network installed at the speaker system, using autotransformers to match the impedance of the network to the voice-coil impedance. It will be observed that the impedance of the monitor line from the amplifiers is 250 ohms. This permits the signal to be transmitted over the line at high voltage, low current, thus reducing

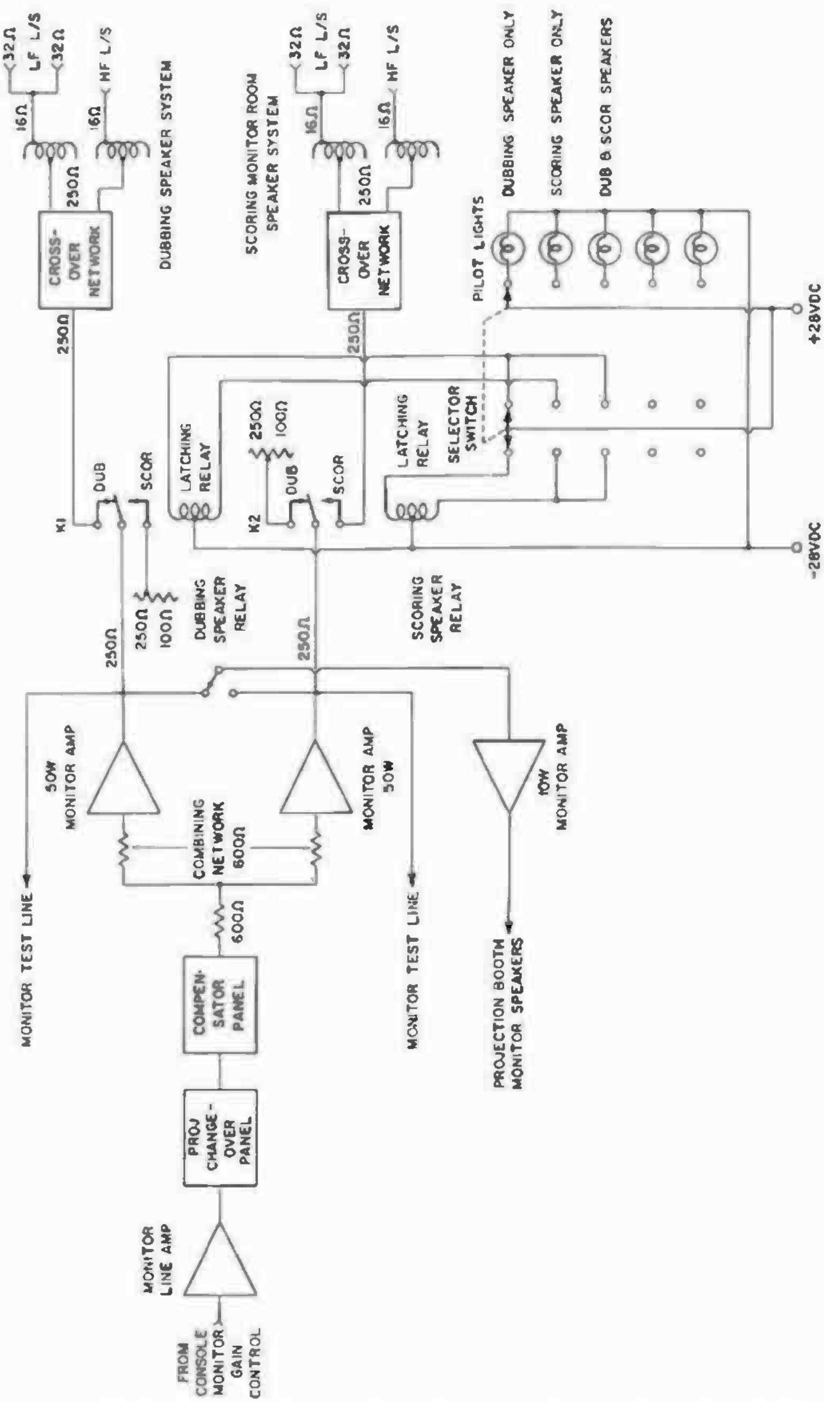


Fig. 9-48. Monitor speaker-selection relay panel for console in Fig. 9-46.

the line losses between the screen and the amplifier system.

Connected at the output of the power amplifiers is a 10-watt monitor amplifier for driving the projection booth monitor speakers. A total of four are employed, one at each projector and one at the operator's rewind bench. Separate talkback speakers are used at each projector control station.

9.49 Describe the operation of the reverberation unit used with the console

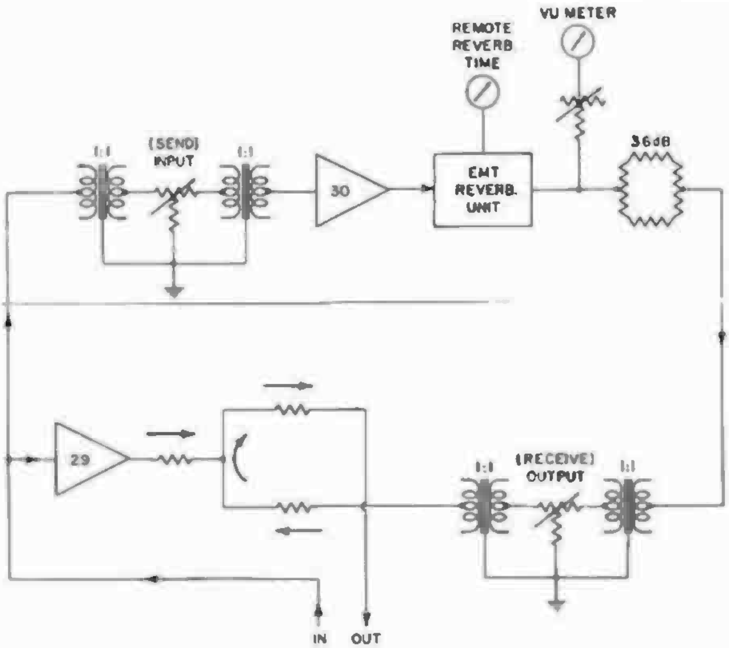


Fig. 9-49. Block diagram for EMT reverberation unit when used with the console of Fig. 9-46. All circuits are 600 ohms.

of Fig. 9-46.—The construction and technical details of the EMT reverberation unit are discussed in Question 2.130. Referring to Fig. 9-49, when the push buttons associated with the reverberation unit are depressed, the direct-signal from the dialogue channel is applied to the input of a two-branch network through amplifier (29). Passing through the upper branch of the network, the direct signal is returned to the recording channel as indicated by the arrows. At the same time, the direct signal is also applied through an input gain control and isolation coils to the input of amplifier (30), and then to the input of the EMT unit. At the output of the reverberation unit is a 36-dB pad and an output gain control which feeds the reverberated signal back to the lower branch of the network indicated by the arrows. As the signal cannot go back through the amplifier, it takes the path of least resistance and flows through the upper branch of the network, returning to the recording channel, in accordance with the arrows. In this manner, the direct and reverberated signals are mixed. After the gains of amplifiers (29) and (30) are determined, they require no further adjustment. The VU meter indicates the reverberated sound level for future reference. A remote meter at the console indicates the reverberation time in seconds.

9.50 ***Describe a solid-state portable mixer-amplifier designed for public address, sound reinforcement, and recording.***—The mixer-amplifier illustrated in Fig. 9-50A, manufactured by Altec-Lansing, is completely solid-state and may be used for public address, recording, and sound reinforcement. The schematic diagram appears in Fig. 9-50B. Starting at the upper left of the diagram, five low-level inputs are available, each connected into a standard octal socket, for use of plug-in microphone and phonograph preamplifiers or line transformers. Tracing the signal from the output of the upper preamplifier socket, the signal passes through an individual speech-music switch (S1). In the music position, the frequency response is flat, and in the speech position the response is rolled off starting at 500 Hz, and continuing downward to 6 dB at 100 Hz, and 16 dB at 30 Hz. The signal is then applied to a 750-ohm mixer control (P1), then to a mixer-combining network consisting of resistors R1 to R5, and to the input of a booster amplifier. At the output of the booster amplifier is a master gain control, which feeds a compressor amplifier having an

Fig. 9-50A. Solid-state portable mixer-amplifier Model 352A manufactured by Altec-Lansing.

attack time of 10 milliseconds, and a fixed compression ratio greater than 5:1. The compressor may be switched in as required. At the output of the booster amplifier is also a connection to feed a magnetic tape recorder. The signal voltage at this point is 0.10 volt, or about minus 55 dB for an input impedance of 10,000 ohms. By taking off the input signal at this point, it is not affected by the master gain control, but only by the mixer controls.

The compressor amplifier is followed by a high- and low-frequency equalizer circuit, then passed on to two voltage amplifier stages (Q2 and Q3), then a phase-inverter (Q4) which drives a compound amplifier (Q5 to Q8). The final amplifier stage consists of four power transistors (Q9 through Q12) operating class-B, and capable of developing 40 watts of output power, with less than 1-percent distortion over a range of 35 to 5000 Hz, and 45 watts with less than 5-percent distortion from 30 to 10,000 Hz. The output transformer is designed to supply output impedances of 4, 8, and 16 ohms, balanced to ground (see Question 8.25). A 70-volt winding is included for use with a loudspeaker distribution system.

The power for the amplifier may be taken from an internal power supply or from external batteries, such as a 12-volt car battery. Provision is also made for a second output to a tape recorder, with a voltage of 0.12 volt. Since this output is taken at the emitter of Q4, it

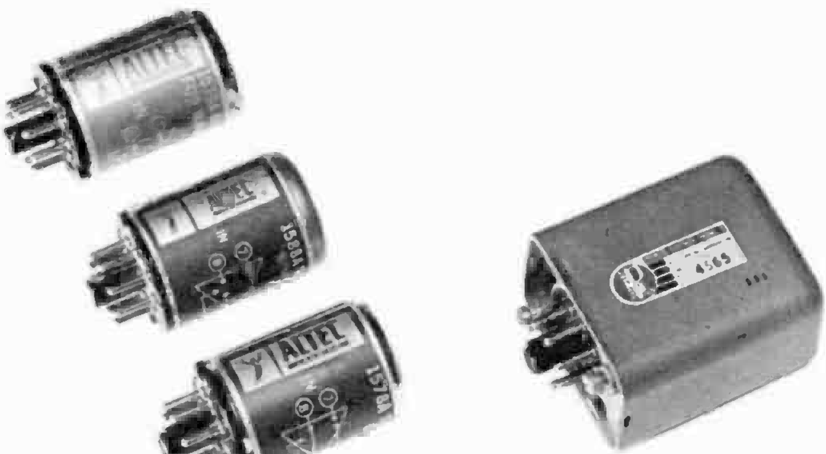


Fig. 9-50C. Plug-in units for the Altec-Lansing Model 352A.

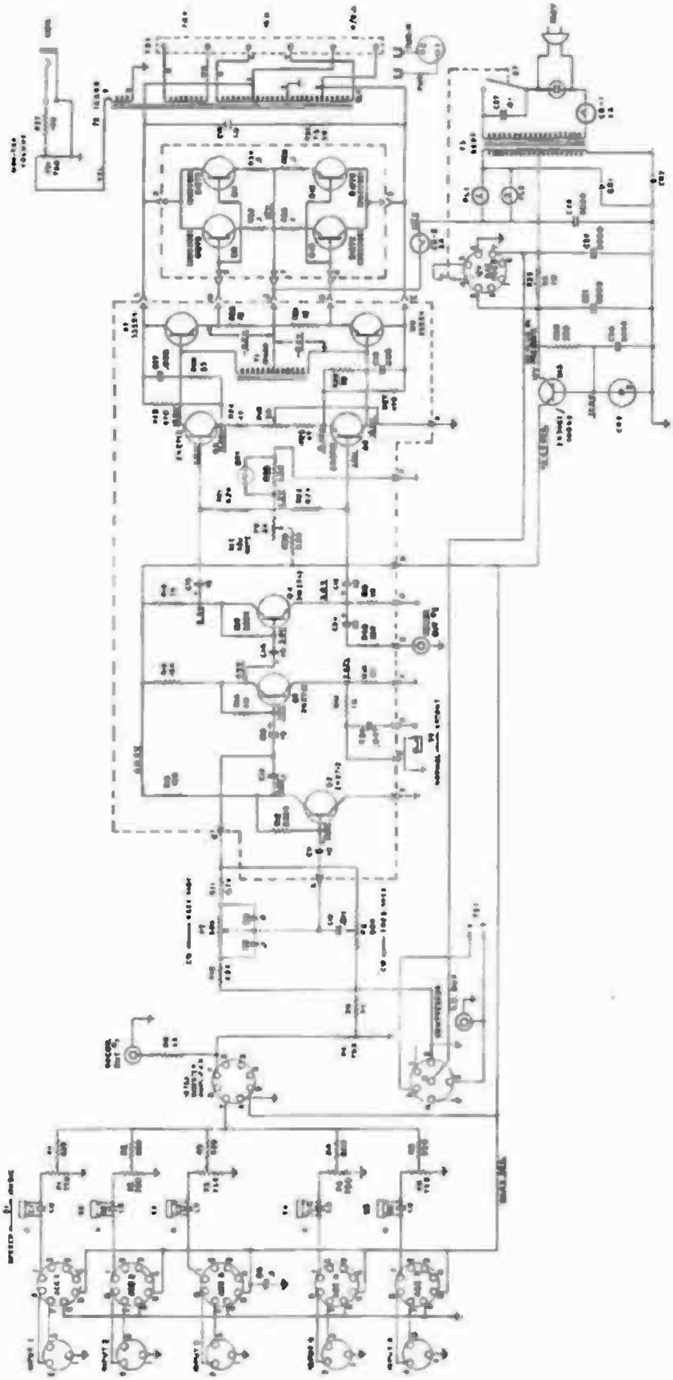


Fig. 9-50B. Schematic diagram for Altec-Lansing Model 352A.

is subject to the adjustments of both the mixer control and the master gain control. A separate winding on the output transformer supplies a means of monitoring the output signal with either headphones or loudspeaker. The VU meter is connected between one side of the 16-ohm output winding and ground, which is also connected to the input of the compressor amplifier. Fig. 9-50C shows the plug-in units. Transistor amplifiers are discussed in Section 12.

9.51 ***How is the output signal from a microphone prevented from overload-***

ing the input of a microphone preamplifier?—When recording program material from a high level source, a 10- to 20-dB network is connected between the output of the microphone and the input of the microphone preamplifier (Fig. 9-51). Many mixing consoles have this feature designed into the circuitry.

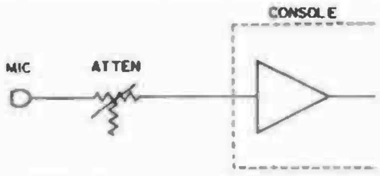


Fig. 9-51. Attenuator network connected in the output of a microphone to prevent overloading the preamplifier.

In a second manner by which preamplifier overloading is prevented, an attenuator network is included in the microphone housing, connected after the transducer element. This reduces the signal level in the microphone itself before it reaches its own preamplifier (if one is used).

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Section 10

VU and Volume Indicator Meters

To properly operate a sound recording or reproducing system, some method for determining the signal levels in different parts of the system to avoid overloading and distortion is required. This is the purpose of the VU or VI meter.

VU meters have been standardized and are widely used in various devices. However, when they are improperly used, difficulties may be encountered for different impedance levels.

This Section discusses the various ramifications of such meters, and the pitfalls that are to be avoided. Simple tables have been included for compensating, if the meter is used at other than the designed impedance. A peak-indicating instrument, the Neon VI meter, widely used before the VU meter made its appearance in the motion picture industry, is described as the experimenter might find it most useful. Ballistics, reference terminology, peak-indicating devices and their calibration, and the insertion of VU meter lead in a recording channel are discussed.

10.1 What is a volume indicator meter?—A meter used to measure power levels of audio-frequency signals. The term volume indicator, abbreviated VI, is generally associated with meters calibrated in decibels. Meters designed for program monitoring purposes have special characteristics, as explaned in Question 10.3.

10.2 What is a VU meter?—A meter used for monitoring broadcast and recording circuits. Such meters employ special ballistic characteristics to properly indicate program material levels which are of complex waveforms and vary simultaneously in both amplitude and frequency. The meter consists of a 200-microampere dc, D'Arsonval movement fed from a full-wave, copper-oxide rectifier unit mounted within the meter case. VU meters are calibrated in reference to one milliwatt of power in a 600-ohm circuit. A typical VU meter panel is shown in Fig. 10-2.

10.3 What are the characteristics of a VU meter?—Because of the inaccuracies inherent in the early copper-oxide rectifier power-level meter and because of the fact it was not satisfactory for program monitoring, the development of an entirely new meter was jointly undertaken by the Bell Telephone Laboratories, Columbia Broadcasting System, and the National Broadcasting Company. The results of this research were the development not only of a new type volume-indicator meter but also the standardizing of a new reference level of one milliwatt, a unit which was adopted by the elec-

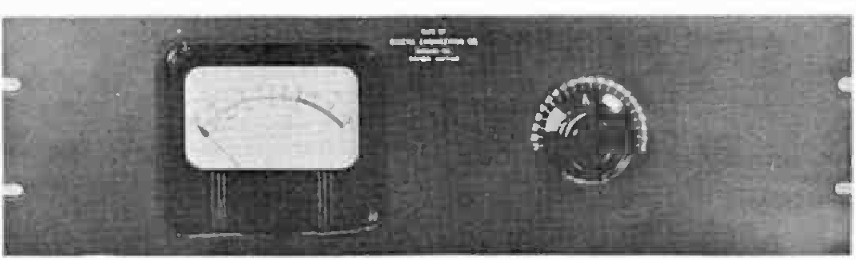


Fig. 10-2. A VU meter panel manufactured by Cinema Engineering Co.

tronics industry in May 1939. The current Standard is USASI (ASA) C16.5-1961.

The characteristics of this meter are as follows:

GENERAL—The meter consists of a dc meter movement with a noncorrosive, full-wave, copper-oxide rectifier unit (mounted in the instrument case) and responds approximately to the root mean square (RMS) value of the impressed voltage. This will vary somewhat depending on the waveforms and the percent of harmonics present in the signal.

INSTRUMENT SCALE—The face of the instrument may have either of the two scale cards shown in Fig. 10-19. Each card has two scales, a VU scale ranging from −20 to +3 VU and a percent modulation scale ranging from zero to 100 percent, with the 100 point coinciding with the zero point on the VU scale. (See Question 10.19.) The normal point for reading volume levels is at zero VU or 100 scale point which is located to the right of the center at about 71 percent of the full scale arc.

DYNAMIC CHARACTERISTICS—With the instrument connected across a 600-ohm external resistance, the sudden application of a sine-wave voltage, sufficient to give a steady-state deflection at the zero VU, or 100 scale point, shall cause the pointer to overshoot not less than 1 percent nor more than 1.5 percent (0.15 dB). The pointer shall reach 99 on the percent scale in 0.3 second.

RESPONSE VS. FREQUENCY—The instrument sensitivity shall not depart from that at 1000 Hz by more than 0.2 dB, between 35 and 10,000 Hz, nor more than 0.5 dB between 25 and 16,000 Hz.

IMPEDANCE—For bridging across a line, the volume indicator, including the instrument and proper series resistor (3600 ohms), shall have an impedance of 7500 ohms when measured with a sinusoidal voltage sufficient to deflect the meter to zero VU or 100 scale point.

SENSITIVITY—The application of a sinusoidal potential of 1.228 volts (4 dB above 1 milliwatt in a 600-ohm line) to the instrument in series with the proper resistance (3600 ohms) will cause a deflection to the zero VU or 100 percent point.

HARMONIC DISTORTION—The harmonic distortion introduced in a 600-ohm circuit, caused by bridging the volume indicator across it, is less than 0.3 percent, under the worst possible condition (no loss in the variable attenuator).

OVERLOAD—The instrument must be capable of withstanding, without injury or effect on the calibration, peaks of 10 times the voltage equivalent to a reading of zero VU or 100 for 0.50 second and a continuous overload of five times that voltage.

10.4 What is a power-level meter? —A VI meter calibrated in decibels. As a rule, this type meter is confined to test equipment for steady-state measurements and is not used for monitoring program material.

10.5 What is a rectifier-type meter? —One which employs a copper-oxide or selenium rectifier unit to rectify the applied ac voltage to dc for operating the instrument movement. This is a very common method employed in meters designed for audio frequency measurements, such as a VU meter, or ac voltmeters for servicing. Such meters have a high input resistance to prevent loading the circuit under measurement.

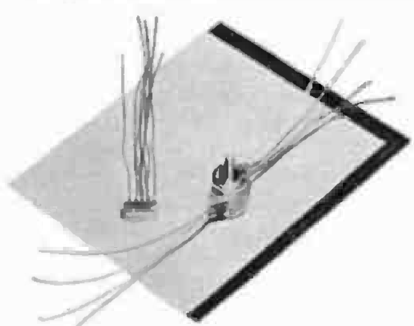


Fig. 10-5. Selenium instrument rectifiers manufactured by Conant Electrical Laboratories.

Two rectifier units manufactured by the Conant Electrical Laboratories are shown in Fig. 10-5. A complete explanation of bridge rectifiers will be found in Questions 21.95 through 21.99.

10.6 Describe a wide range electronic VU meter for program monitoring. —Standard VU meters measure only the upper 23 dB of the signal level. From a practical standpoint this limits the display to about 20 dB below the reference level of the zero indication.

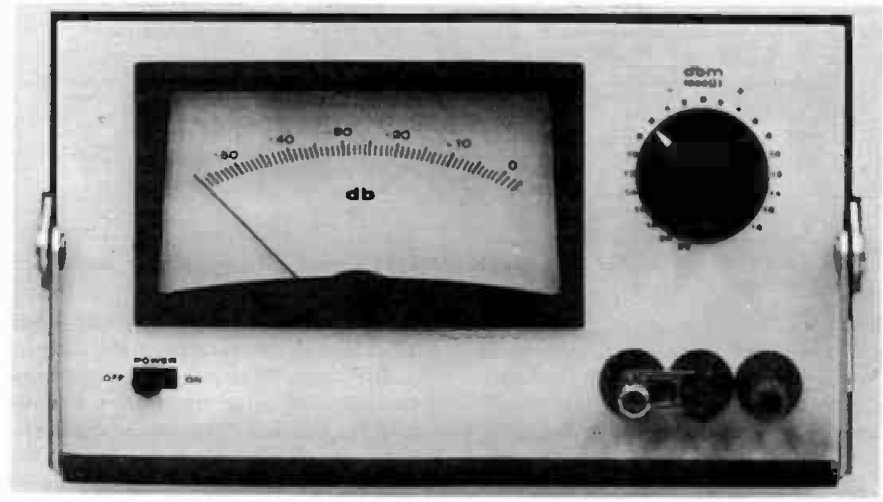


Fig. 10-6A. CBS Laboratories Model 600 wide-range program-monitor meter.

This short range of operation limits its usefulness, particularly when it is connected across a bridging bus for monitoring program information. A wide-range program-monitor meter, Model 600, manufactured by CBS Laboratories, which displays the program information over a 60-dB meter scale, spread from minus 57 dB to plus 3 dB, is shown in Fig. 10-6A. The large spread of program material permits the very low level signals to be observed, and also the noise between program pauses.

This instrument was not designed to replace the conventional VU meter; however, its characteristics are compatible with the present VU meter. In addition, a dc output is provided for connection of a linear tape recorder for logging program levels over a range of 60 dB. The zero-dB indication may be set to represent a reference level from minus 22 dBm to plus 18 dBm.

Referring to the block diagram in Fig. 10-6B, the basic component is a logarithmic solid-state amplifier, with a nonlinear feedback circuit (see Question 12.196), a preamplifier, a 15,000-ohm bridging input transformer, a reference-level selector switch, and a sensitive indicating meter movement. The total range of measurement is 79 dB to plus 21 dB, with a frequency response of plus or minus 1 dB from 50 to 15,000 Hz.

10.7 What is a neon VI meter?— In the early days of motion picture sound recording, and long before the use of magnetic film, sound was recorded directly on photographic film (optical recording). Because the VI meters at that time were not standardized and had radical ballistics, peaks of 8 to 10 dB were not indicated by the meter, and the light modulator could be easily overloaded, causing serious

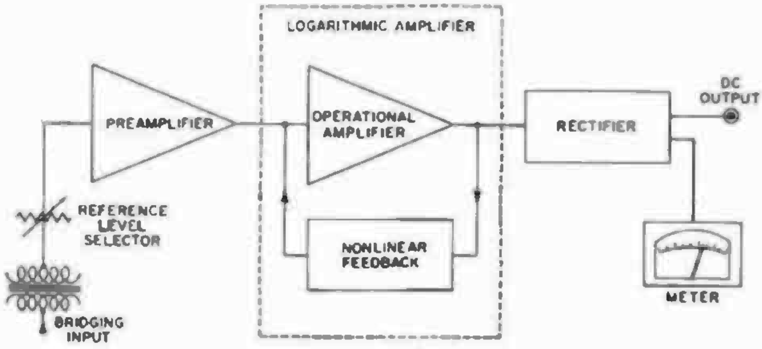


Fig. 10-6B. Block diagram for CBS Laboratories Model 600 wide-range program-monitor meter.

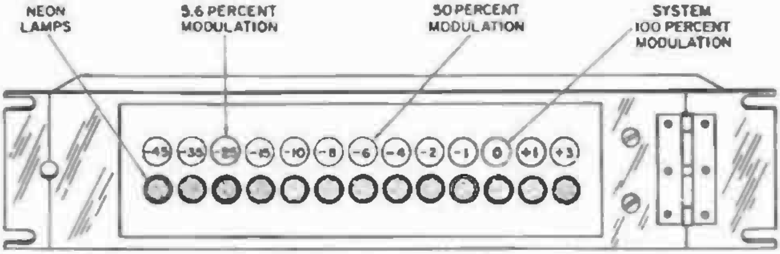


Fig. 10-7A. Front view of RCA Model MI-3176A neon volume indicator meter.

distortion. To provide a meter that would read the peaks and at the same time permit the mixer to see the indications, RCA developed an instrument known as a neon volume indicator meter (no longer manufactured).

The meter consisted of 20 small vacuum tubes and 16 neon lamps, arranged in a circuit to indicate the recording levels from minus 45 dB to plus 3 dB, with reference to 100 percent modulation of the light modulator. The reason for the use of such a device was its sensitivity and ability to read both the negative and positive peak excursions of the light modulator.

As this device is of interest to present day recording and is in some instances still in use, a description of its principles are in order. The mechanical design was such that the neon lights were mounted horizontally in a straight line (Fig. 10-7A). In each hole in the lower line is a neon lamp, which is actuated by the audio signal and made to glow. At the extreme left is a minus 45-dB lamp and at the extreme right the plus 3-dB lamp. Pilot lamps glowing continuously indicate the minus 25-, 6-, and 0-dB levels. The basic circuit for such an instrument is shown in Fig. 10-7B. The revision of this circuit to transistors would be of considerable value to present day recording, as the industry is still faced with the use of monitoring meters that do not indicate the actual peak values of the audio signal.

The block diagram for a neon VI meter is shown in Fig. 10-7B. The signal is applied to the input transformer T1, then amplified by a two-stage amplifier. At the output of the amplifier, the signal is applied to two voltage dividers, P1 and P2. From there the signal passes through several other amplifiers, each with a neon lamp at its output. The gain of each amplifier stage is adjusted by P1 and P2 to fire the neon tube at a given level.

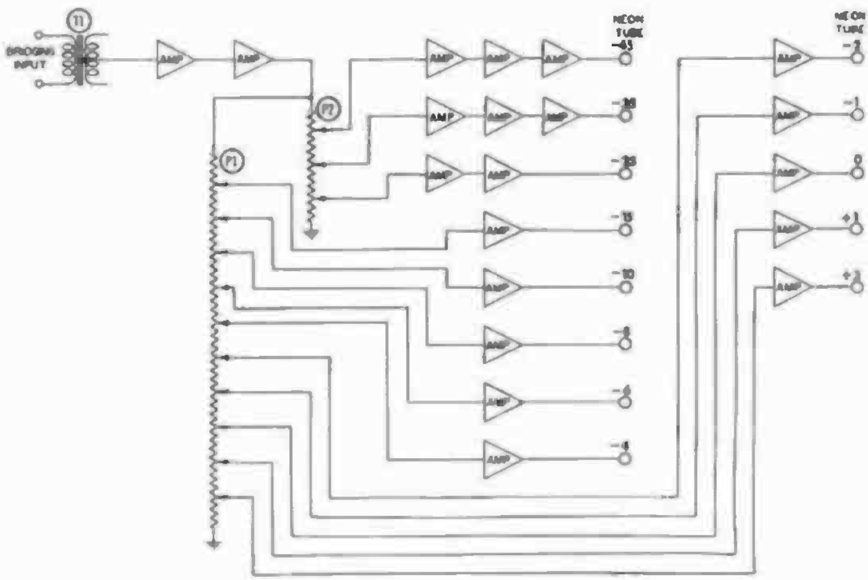


Fig. 10-7B. Basic circuit for neon volume indicator.

10.8 What is the standard reference level used for recording and broadcasting?—Both the recording and broadcasting industries as well as most other electronic industries use the 1-milliwatt reference level. This level was standardized in May of 1939. In a few instances for special purpose devices, the 6-milliwatt reference level is used.

10.9 What does the term dBm mean?—It indicates the stated level is with reference to 1 milliwatt of power in a 600-ohm line.

10.10 What is the reference level when it is stated only in dB?—It is with reference to 6 milliwatts in a 500-ohm line.

10.11 What is the voltage across a 600-ohm line for 1 milliwatt of power?—The voltage equals 0.773 volt.

10.12 What is the voltage across a 500-ohm line for 6 milliwatts of power?—The voltage is equal to 1.73 volts.

10.13 Define the terms volume unit (VU) and decibels (dB).—The volume unit (VU) is a unit of measurement used with volume-indicator meters to specify a change of 1 dB in volume for a complex waveform. Such units are used with the VU meter described in Question 10.3. Complex waveform changes can only be measured in volume units. Meters designed for such use have special characteristics. For complex waveforms, such as speech, a VU meter reads between the average and the peak values of a complex wave. No simple relationship exists between volume measured in VU and the power of a complex waveform. The indicated reading will depend on the particular wave shape at the moment. For sine-wave measurements, a change of one VU is numerically equal to a change of 1 dB.

VU meters are designed to have a dynamic characteristic that approximates the response of the human ear. When a speech waveform is applied to a VU meter the movement will indicate peaks and valleys in the signal. The average of the three highest peaks in 10 seconds (disregarding occasional extremes) is taken to be the indication of the meter movement.

From the above discussion it appears the VU should not be used to interpret steady-state measurements, and is only used with complex waveforms which are constantly changing in amplitude and frequency, and depend on human interpretation of a constantly changing condition. The decibel is used to interpret steady-state conditions. It should be pointed out that many meters indicated as VU meters are not actually such meters, as they are normal sensitive movements without the special characteristics of the standard VU meter. Therefore, they do not indicate a true reading in volume units.

10.14 Why are VU meters calibrated with reference to 1 milliwatt of power in a 600-ohm line?—This reference level was selected as a level which would conform to the Telephone Company's standards of limiting the signal level on a transmission line to a value that would produce a minimum of cross talk and still provide a satisfactory signal-to-noise ratio. Also, the 1 milliwatt reference level is a unit quantity. Hence, it is readily applicable to the decimal system, being related to the watt by the factor 10^{-3} which results in positive values for the majority of measurements.

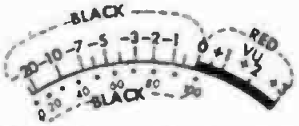
A further advantage is that all meters of this type are exactly alike in construction and characteristics and, when several are connected across the same circuit, may be tested and their operation checked by the application of a 1000-Hz signal.

10.15 What is the maximum signal level that may be applied to a telephone line?—Plus 8 dBm.

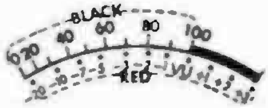
10.16 What does the term zero level mean?—It is a reference power level. Example; for 1 milliwatt of power in a 600-ohm line, zero equals 0.773 volt. For a 500-ohm line, zero equals 1.73 volts (7.78 dB higher than 1 milliwatt).

10.17 Where is 1 milliwatt indicated on a VU meter scale?—For a VU meter of 7500-ohms input impedance connected across a 600-ohm line in which 1 milliwatt of power is flowing, 1 milliwatt will be indicated at the minus 4 VU (or dB) calibration point, for the reason that a VU meter input attenuator has no zero position, but starts its calibration at plus 4 VU. Therefore, 1 milliwatt will be indicated at the minus 4 VU (or dB) calibration point. This will be discussed further in Question 10.24.

10.18 What reference levels have been used other than 1 milliwatt?—In the early days of broadcasting and re-



(a) *Recording and test equipment.*



(b) *Broadcast monitoring.*

Fig. 10-19. VI meter scales.

cording, both 10 and 12.5 milliwatts in a 500-ohm line were used as a reference level. However, later this was changed to 6 milliwatts. In May, 1939 the present standard of 1 milliwatt in a 600-ohm line was adopted. Fifty milliwatts was used at one time by the radio industry for rating the output level of a radio receiver for a given input at the antenna in microvolts. This too has been replaced with other standards.

10.19 Explain the difference between an A and a B scale used on VU meters.—VU meters may be obtained with the upper portion of the scale arc calibrated in either percent modulation or decibels. For recording purposes the A scale is preferred because the levels are read in decibels. For broadcast work, the percent-modulation scale is preferred as it indicates the percent modulation of the radio transmitter. However, both scales are widely used. The two scales are pictured in Fig. 10-19.

10.20 Where is the point of greatest accuracy on the scale of a VU or VI meter?—The accuracy for the full-scale deflection is 2 percent; however, the minus 1 and the plus 1 dB calibration points above and below the zero calibration are generally used when reading levels as they are quite accurately indicated.

10.21 What are the ballistics of a VU meter?—Ballistics are the mechanical and electrical characteristics built into the meter movement. A given characteristic may be obtained by shaping the pole pieces and counterweighting the pointer mechanism. Shunts are sometimes used across the meter terminals, but this will reduce the sensitivity of the movement. (See Question 10.3.)

The ballistics characteristics of a typical old style VI meter and a standard VU meter, when a 1000-Hz signal is applied for a period of one second, are shown in Fig. 10-21. It will be noted the VU meter comes to a steady state at the end of 0.30 second, while the VI meter continues to oscillate showing peaks and valleys over a period of one second.

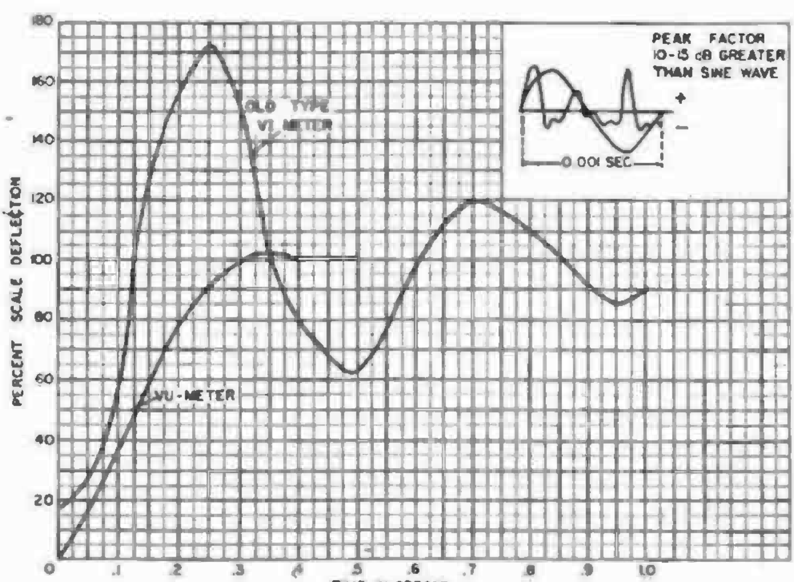


Fig. 10-21. Comparison of the original VI meter and the present VU meter ballistics for a sudden application of a 1000-Hz signal for a period of 1 second. (Courtesy, Proceedings of IRE, now IEEE.)

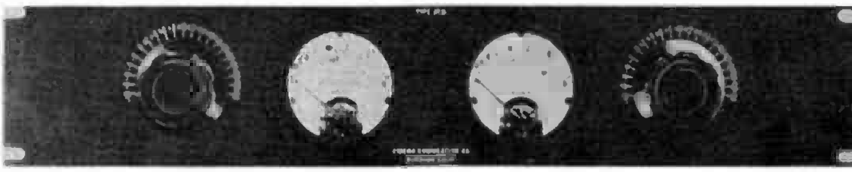


Fig. 10-22. An early type dual volume indicator panel manufactured in 1939 and now obsolete. (Courtesy, Cinema Engineering Co.)

The VI meter movement overshoots about 65 percent in the first 0.25 second, while the VU meter movement has almost reached a steady state. This clearly indicates why the ballistics of the VU meter are more desirable than those of the older-type VI meter for monitoring program material containing complex waveforms.

10.22 What type meters were used before the adoption of the VU meter?—Copper-oxide rectifier meters of the type shown in Fig. 10-22. The characteristics of these meters were not standardized, although they were used extensively in both the recording and broadcasting industries. This meter possessed characteristics which were not suitable for monitoring purposes, although they were quite suitable for sine wave power measurements.

The older-type meters used a reference level of 6 milliwatts in a 500-ohm line as the zero level calibration point (1.73 volts). This appeared on the meter scale slightly to the left of center scale.

Two sensitivities in this meter were used; they were zero equals 1.73 volts and zero equals 0.548 volt. In the latter meter, zero was equal to a minus 10-dB signal in a 500-ohm line. Three movements were available; fast, medium, and slow. The range of measurement was extended by the use of an attenuator as used with the present VU meter.

10.23 How may the ballistics of the older type VI meters be identified?—By the letters appearing on the meter scale. HS indicates high speed, GP is for general purpose, and SS indicates slow speed. The high-speed movement was used for monitoring purposes, the general-purpose movement for test equipment, and the slow speed for acoustic measurements or where an extremely slow response was required.

10.24 What are the recommended connections for a VU meter?—Referring to Fig. 10-24 it will be seen the instrument consists of an indicator movement, a variable attenuator, and a series resistor of 3600 ohms. The meter movement is a 200-microampere D'Arsonval movement with an internal resistance of 3900 ohms. A full-wave, copper-oxide or selenium rectifier is contained within the meter case. The attenuator is variable in steps of 2 dB and presents a constant resistance of 3900 ohms to the meter movement. This design prevents the ballistics of the meter from being affected when the attenuator setting is changed.

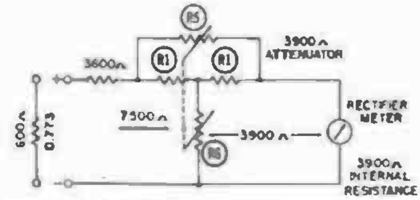


Fig. 10-24. Schematic diagram for a 7500-ohm VU meter, calibrated for one milliwatt reference level or, 0.773 volt across 600 ohms.

Standard VU meters are designed to read zero VU, or 100 percent, with 1.228 volts (+4 dBm) applied to the instrument with the 3600-ohm series resistance connected ahead of the attenuator. If the meter is used with the attenuator but without the 3600-ohm series resistor and is connected across a 600-ohm source in which 1 milliwatt of power is flowing, the movement will be deflected to the 100 percent calibration point.

This method of use is not recommended because the impedance looking back into the meter is only 3900 ohms and is too low an impedance to bridge a 600-ohm circuit. It is the usual practice to keep the impedance of bridging devices at a ratio of 10:1 or greater.

To increase the input impedance of the VU meter from 3900 ohms to 7500 ohms, a 3600-ohm resistor is connected in series ahead of the attenuator. However, in so doing a 4-dB loss is incurred across the 3600-ohm resistor. If a signal of 1 milliwatt (0.773 volt) is impressed across the input terminals of the circuit in Fig. 10-24, it will not deflect the meter to the 100 percent calibration but only to the minus 4 VU (or dB) mark, or approximately 65 percent. This means that if the meter is to be deflected to the 100 percent point, the input signal must be increased to a plus 4 dBm. This is the reason why 1 milliwatt of power will be indicated at the minus 4 calibration mark.

Attenuators used with VU meters have no 0-dBm calibration step but start at a plus 4 dBm. The bridging loss of a 7500-ohm VU meter is quite small (in the order of 0.34 db) when connected across a line of 600 ohms impedance.

10.25 How does the attenuator of a VU meter function?—VU meters are, in reality, voltmeters calibrated in decibels to read with respect to a reference voltage. The attenuator inserts loss ahead of the meter in a manner similar to a voltmeter multiplier, except that in the case of the VU meter, the loss is inserted in steps of one or more dB while the impedance remains constant.

10.26 How are the readings of the input attenuator and the meter scale added?—Algebraically. For example, if the attenuator is set to a plus 10 dBm and the meter indicates a plus 3 dB, the level is plus 13 dBm. On the other hand, if the attenuator is set to a plus 30 dBm and the meter reading is a minus 3 dB, the true level is plus 27 dBm.

10.27 Does a VU meter indicate the true level of a complex waveform?—No. It indicates somewhere between the average and the peak values. Program material is of a complex and transient nature; therefore, the VU meter indicates considerably under the instantaneous peak program level. This means that many peaks present in the program material are not indicated by the meter because the meter movement cannot follow small instantaneous peaks which are varying both in amplitude and in frequency simultaneously. Therefore, the meter must either be set or caused to indicate in a manner that will not overload the system in which it is operating.

Peak voltages of 8 to 14 dB may be occurring above the meter indication but, because of the ballistics of the meter, are not indicated. Even if they could be seen it would be too late to reduce the level. Therefore, a VI or VU meter must be set with at least an 8-dB lead, which is described in Question 10.28.

10.28 What is the procedure for inserting lead into a VU meter circuit to prevent overloading a recording system?—Since VU meters do not indicate the true peak values of program material (complex waveforms) it is quite easy to overload a recording system. To protect against these unseen peaks, a lead or margin of safety is inserted in the VU meter circuit.

To illustrate the procedure for inserting a lead into a VU meter circuit, assume a recording channel employing both photographic film and magnetic recorders is to be adjusted. Further assume that the VU meter is connected across a bridging bus with a sine-wave level of plus 14 dBm. With the VU meter set to plus 14 dBm, a 400- or 1000-Hz signal is sent into the input of the recording console. The mixer control is set to its normal operating range and the signal level adjusted to bring the bus level to plus 14 dBm (the VU meter reads 100 percent or zero dBm). Adjust the recording amplifier for the photographic recorder to deflect the light modulator to *exactly* 100-percent modulation. Assuming the operating level for the magnetic recorder has been determined as set forth in Question 17.163, the recording amplifier is adjusted for 100-percent modulation.

Remove the input signal and return the VU meter attenuator to its plus 6-dBm position. This inserts an 8-dB lead or margin of safety in the VU meter by making it 8 dB more sensitive, thus protecting the system against unseen peaks up to 8 dB. The program material is now mixed in the usual manner. Some recording activities, because of the heavy peaks and overloads encountered in some types of music, use a 10- to 12-dB lead in the VU meter.

Radio transmitters are adjusted in a similar manner, only in this instance the percentage modulation indicated by the VU meter indicates the percent modulation of the radio transmitter.

10.29 What is the input impedance of the older type VI meters described in Question 10.22?—Generally, 5000 ohms for those designed to indicate 1.73 volts across a 500-ohm line. The circuit diagram for the older type meter is shown in Fig. 10-29. This type meter panel uses a 5000-ohm L-type attenuator ahead of the meter movement. Although the L-pad does not present a constant impedance to the meter movement, it does present a constant 5000 ohms to the circuit being bridged. The two arms of the attenuator are ganged together mechanically and are varied inversely to each other. The attenuator is designed to increase the level of the meter readings in steps of 2 dB each, starting at zero level (6 milliwatts).

One of the disadvantages of this type meter, when used for monitoring program material, is the ballistics of the meter which are changed with changes in the setting of the input attenuator. This is due to the shunt arm of the attenuator being reduced in value and acting as a shunt across the meter movement, as the attenuator is set for a higher range The ballistics of this meter are not suitable for program monitoring, and have been obsolete for some years. The ballistics and frequency characteristics also vary with different manufacturers.

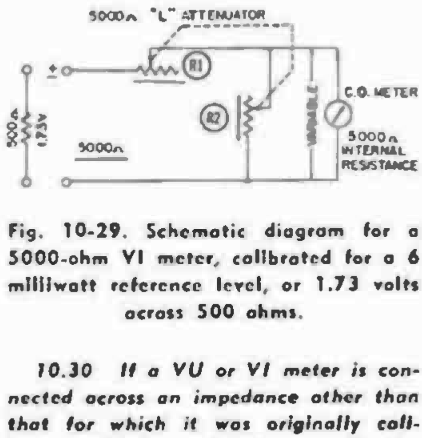
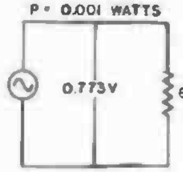


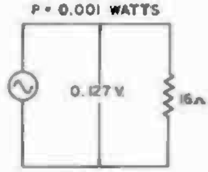
Fig. 10-29. Schematic diagram for a 5000-ohm VI meter, calibrated for a 6 milliwatt reference level, or 1.73 volts across 500 ohms.

10.30 If a VU or VI meter is connected across an impedance other than that for which it was originally calibrated, will the reading be correct?—No. A correction factor must be applied to the indicated reading to correct for the difference in the impedance.

10.31 Why must a correction factor be applied to a VU meter when connected across an impedance different from that which it was originally calibrated?—VU and VI meters are in reality voltmeters calibrated in decibels with respect to a reference power level. If connected across a line impedance different from that for which it was originally calibrated, the voltage supplied to the meter will either be lower or higher than the original calibration. Therefore, the meter indicates incorrectly. Two circuits are shown in Fig. 10-31, one a 600-ohm circuit and the other a 16-ohm circuit. Both are dissipating the same amount of power, yet the voltage across each circuit is quite different. For the 600-ohm circuit, the voltage is 0.773 volt, and for the 16-ohm circuit it is 0.127 volt. It may readily be seen that if a VU meter is connected across the 16-ohm circuit, it will not deflect the same amount as for the 600-ohm circuit, although the same amount of power is flowing in each circuit. To arrive at the correct power level in the 16-ohm circuit a correction factor must be applied to the meter indication.



(a) *600-ohm line.*



(b) *16-ohm line.*

Fig. 10-31. Voltages across lines of different impedance but with same power in milliwatts.

10.32 Give the equation for VU and VI meter impedance level correction.

$$dB = 10 \operatorname{Log}_{10} \frac{Z_2}{Z_1} \text{ or } \frac{Z_1}{Z_2}$$

where,

Z_1 is the impedance of the circuit bridged,

Z_2 is the impedance for which the meter is calibrated.

When the line impedance is greater, the equation is inverted and the correction subtracted from the indicated level. When the line impedance is low, the correction factor is added. A typical example of applying a correction factor would be as follows: A VU meter calibrated for a line impedance of 600 ohms is bridged across a 16-ohm line as shown in Fig. 10-31. If the meter indicates a level of plus 1 dBm, what is the true level?

$$dB = 10 \,Log_{10} \frac{600}{16}$$
$$= 10 \times 1.574$$
$$= 15.74 \,dB.$$

The correction factor of 15.74 dB is added to the meter reading of plus 1 dBm, which results in a true level reading of plus 16.74 dBm. Typical correction factors are given in Fig. 10-32A.

Shown in Fig. 10-32B is an impedance-level correction graph for meters calibrated with reference to 1 milliwatt, 600 ohms. The new impedance to be bridged is entered at the lower margin and followed upward to the diagonal line. The correction factor is read at the left margin.

Line Impedance (ohms)	Meter Cal. 500 Ohms (dB)		Meter Cal. 600 Ohms (dB)	
600	−0.791	subtract	0.000	
500	0.000		+0.791	
250	+3.010		+3.800	
200	+3.970		+4.770	
150	+5.230		+6.020	
125	+6.020		+6.810	
100	+6.990		+7.780	
50	+10.000	add	+10.790	add
30	+12.220		+13.010	
16	+14.940		+15.740	
15	+15.220		+16.020	
8	+17.960		+18.750	
4	+20.970		+21.760	

Fig. 10-32A. VU and VI meter impedance correction factors.

10.33 Define the term bridging loss when applied to a volume indicator meter.—It is the reduction in signal level experienced when a VU meter is bridged across a circuit. The drop in signal level is caused by the absorption of power by the meter circuit. As a rule, the power absorbed is quite small and may for most purposes, be ignored. However, at high powers, it may become important. Bridging loss may be calculated by the equation:

$$dB = 20 \,Log_{10} \frac{2B_R + R}{2B_R}$$

where,

B_R is the VU meter input impedance,
R is the line impedance.

A table of bridging losses for the more commonly used impedances and meters is given in Fig. 10-33.

10.34 How may a standard VU meter designed for 600-ohm use be employed in a line having an impedance of 150 ohms?—By the use of an input transformer as shown in Fig. 10-34, recommended by the Weston Instru-

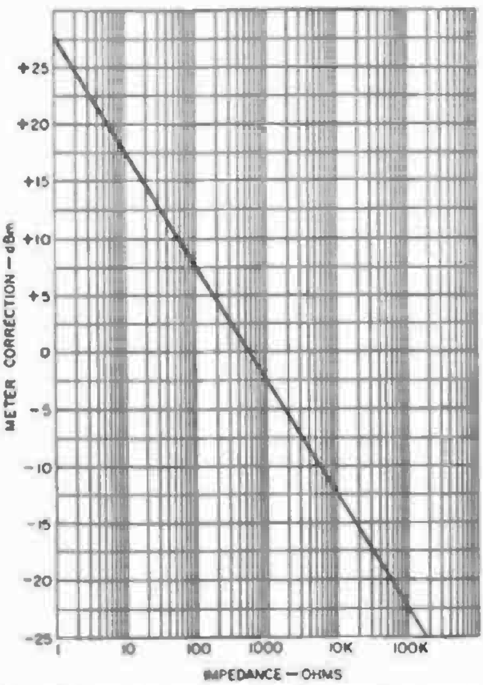


Fig. 10-32B. Impedance-level correction graph.

ment Corporation. The transformer has a turns ratio of 1:2 which results in a gain of 6 dB, and may be used with the switches S1 and S2 set to "T," up to any level for which the transformer is suitable. Switch S3 may be open for ungrounded circuits or omitted entirely. The circuit is closed if a grounded system is necessary.

Meter Impedance (ohms)	*Line Impedance (ohms)*	*Bridging Loss (dB)*
5000	500	0.47
5000	600	0.55
7500	500	0.30
7500	600	0.38

Fig. 10-33. VU and VI meter bridging losses for circuits of 500 and 600 ohms.

However, since any transformer has power limitations and since a 12-dB pad can be arranged to serve as a match between the line and the load, the diagram shows an arrangement whereby the transformer is switched out of the circuit and a fixed pad switched in at the 12-dB or 16-dB point. The two line switches S1 and S2 are transferred to "P" and switch S3 must be closed, if previously left open.

It is convenient to associate these transfer switches with the 12-dB position on the attenuator. The transformer is used for levels up to plus 12 dBm. Above plus 12 dBm, the pad is switched into the circuit.

10.35 What is a power-output meter?—A meter similar to the one shown in Fig. 10-35A which is used for measuring the power output of audio amplifiers and other devices. It may also be used to determine the characteristic and internal output impedance, the effect of load-impedance variation, and other applications involving the measurement of output power and impedance with respect to frequency.

The circuit of the meter (Fig. 10-35B) consists of a load impedance adjusting network composed of resistor R1, impedance-selecting switch S1, multiplier switch S2, frequency-correcting networks, a calibrating pot, and indicating meter M1.

The impedance-adjusting network provides for the selection of 40 different load impedances between 2.5 and 20,000 ohms. The meter scale is calibrated from 1 to 50 milliwatts, and from 0 to 17 dB. The meter multiplier switch extends the power readings from 0.10 to 5000 milliwatts (5 watts), and the dB reading from minus 10 to plus 30 dB in steps of 2 dB. The location of the controls may be seen in Fig. 10-35A.

Fig. 10-35A. A power-output meter manufactured by the Daven Company.

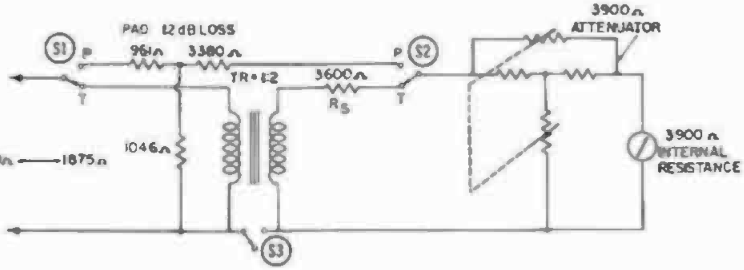


Fig. 10-34. Circuit for a standard VU meter panel, with input transformer and pad, for operation on a 150-ohm transmission line.

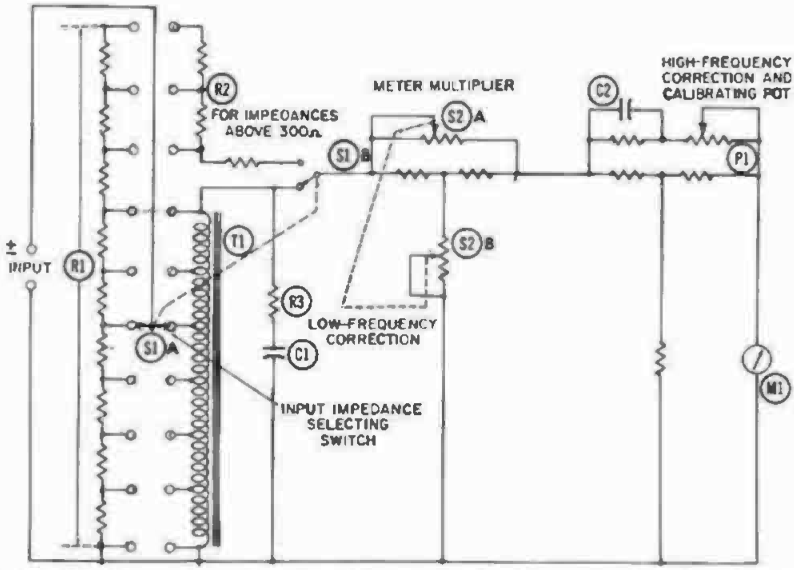


Fig. 10-35B. Elementary schematic diagram of the Daven Model OP-961 power-output meter.

Referring to Fig. 10-35B, the input terminals are shunted by a 40 section noninductive resistance network R1, its value controlled by switches S1A and B. For impedance values up to 300 ohms, resistor R1 is bridged by a high impedance autotransformer T1, the appropriate tap being selected by switch S1A. As the resistance of R1 is increased or decreased, the ratio of transformation of the autotransformer coil is increased or decreased in the proper relation, so that the power dissipated in resistor R1 is directly proportional to the voltage appearing across the meter M1.

To ensure that the input impedance remains essentially noninductive, the bridging impedance presented by the autotransformer to resistor R1 is quite high. For impedances above 300 ohms, switch S1B disconnects the coil and substitutes a noninductive resistor, R2. This resistor then controls the voltage appearing across the meter. The selection of the taps on resistor R1 as well as switching from the autotransformer T1 to resistor R2 is controlled by the positions of switches S1A and B.

The actual value of impedance presented by the meter is the value of R1 in parallel with the autotransformer T1. (The design of such autotransformers is discussed in Question 22.124.) Above 300 ohms, resistor R2 is substituted for the transformer and is shunted by the remainder of the circuit. A frequency-correcting network consisting of resistor R3 and capacitor C1 is connected across the output side of coil T1 to compensate for losses in the coil at the lower frequencies.

The meter-multiplier switch S2A and B consists of a bridged-T network calibrated in power ratio and decibels. A second bridged-T network serves as a combination high-frequency correction network and basic calibration control. The indicating meter is a conventional rectifier meter.

The input impedance and power accuracies are plus or minus 2 percent over a range of 30 to 10,000 Hz. The reference power is 1 milliwatt in a 600-ohm circuit (zero dBm). The device may be used for many different types of measurements; among them are:

(A) *MEASUREMENT OF POWER OUTPUT AT A GIVEN IMPEDANCE AND FREQUENCY.* Adjust the meter for a reading near its center scale. Maximum power output is the meter indication multiplied by the meter multiplier (S2). For variable-impedance devices, set the meter impedance to the required load impedance and note the power output. This same procedure is used for all impedance values of interest.

(B) *MEASUREMENT OF INTERNAL OUTPUT IMPEDANCE.* Vary the meter impedance until a maxi-

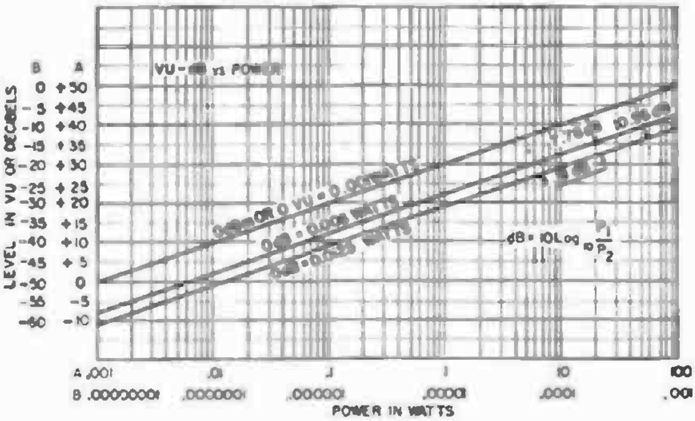


Fig. 10-40. Chart showing the difference in decibels between 1, 6, and 12.5 milliwatt reference power levels.

mum power-output indication is obtained. The internal output impedance of the device under test is equal to the meter impedance.

(C) *MEASUREMENT OF INTERNAL OUTPUT IMPEDANCE WITH RESPECT TO FREQUENCY.* Hold the input signal to the device under test at a constant value. Measure the internal output impedance for each frequency of interest, by noting the maximum power output.

Care must be exercised in assuming that the impedance which develops the maximum power output is the correct load impedance for the device under test. Many devices (such as negative-feedback amplifiers) have an internal output impedance. If known, the correct load impedance should always be used when making measurements involving the use of power output meters.

10.36 What is the reference-level terminology used in the sound and electronic industries?

dB	6 milliwatts, 1.73 volts, 500 ohms.
dBa	Noise measurements (dB adjusted).
dBj	1000-microvolts reference.
dBk	1-kilowatt reference.
dBm	1 milliwatt, 0.0775 volt, 600 ohms.
dBμ	1-microvolt reference.
dBV	1-volt reference.
dBW	1-watt reference.
dBx	Crosstalk measurements.
dBrap	Decibels above the reference acoustical power, 10^{-16} watts.
dBrn	Relationship of noise to a reference level.
dBrnc	Crosstalk measurements.
dBVg	Decibels of voltage gain.
VU	1-milliwatt; complex waveforms varying in both amplitude and frequency.

10.37 What is a transmission unit?—A now obsolete term formerly used for expressing gain or loss. It has been replaced by the term decibel.

10.38 What does the term mile-of-loss mean?—This is an obsolete term formerly used in the telephone industry to express gain or loss in terms of loss

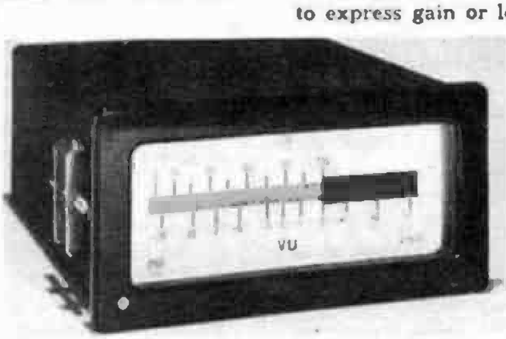


Fig. 10-41. Moving beam of light VU meter. (Courtesy, Gotham Audio Corp.)

for one standard mile of telephone cable or wire. It was replaced by the transmission unit (now also obsolete) and subsequently, by the present term decibel. 1 Mile of Standard cable × 0.947 = 1 decibel.

10.39 ***Define the neper.***—A term used to express gain or loss. Mathematically, the neper is defined:

$$\text{Nepers} = \tfrac{1}{2}\ \text{Log}_\epsilon \frac{P_1}{P_2}$$

where,

P_1 and P_2 are two powers,
ϵ equals 2.718 the base of the Naperian system of logarithms.

The relationship between decibels, nepers, and miles of standard cable may be found as follows:

Multiply	By	To find
decibels	0.1151	nepers
decibels	1.056	miles of standard cable
miles of standard cable	0.947	decibels
miles of standard cable	0.109	nepers
nepers	8.686	decibels
nepers	9.175	miles of standard cable

(Nepers are still used in some parts of Europe.)

10.40 ***Show the relationship in decibels, between a 1, 6, and 12.5 milliwatt power-reference level.***—The difference between these three reference powers may be seen in the graph of Fig. 10-40. The difference between 1 milliwatt and 6 milliwatts is 7.78 dB, and for 12.5 milliwatts 10.96 decibels.

Decibel conversion graphs for 1 and 6 milliwatts sine-wave power are given in Fig. 25-105A and B.

10.41 ***What is a moving light beam VU meter?***—A VU meter employing a straight-line scale, consisting of a d'Arsonval movement and a rectifier. An optical system projects a beam of light through a slot onto a mirror attached to the movement; in turn it casts a narrow beam of light onto a ground glass scale, calibrated in decibels (Fig. 10-41). The frequency response is 20 to 20,000 Hz, 3900-ohms input impedance—normally used with 3600-ohm resistor

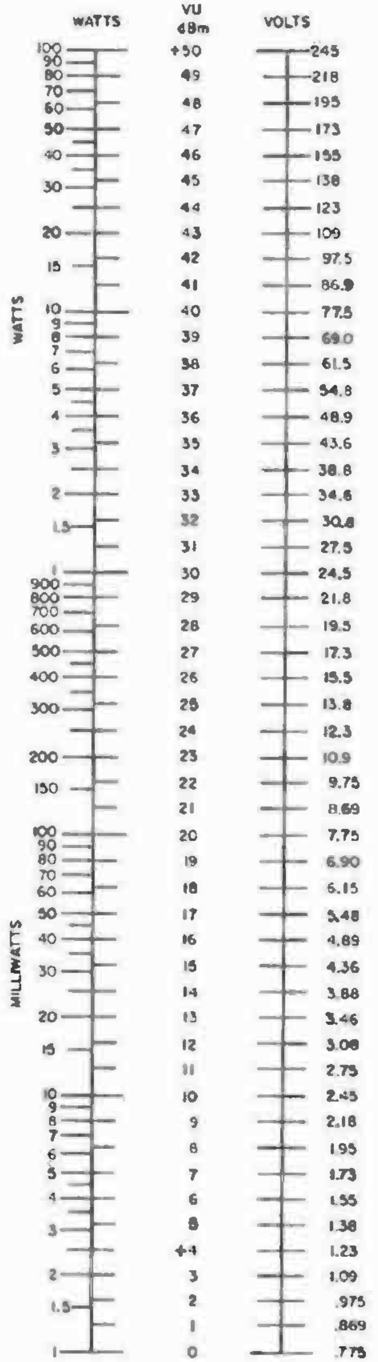


Fig. 10-42. Relationship of VU and dBm to power in watts and voltage in a 600-ohm line.

(see Question 10.24). The ballistics are in accordance with USASI (ASA) Standard C-16.5, 1961. The sensitivity is the same as for a standard VU meter. When the level deflects over zero VU, the light doubles in height and turns

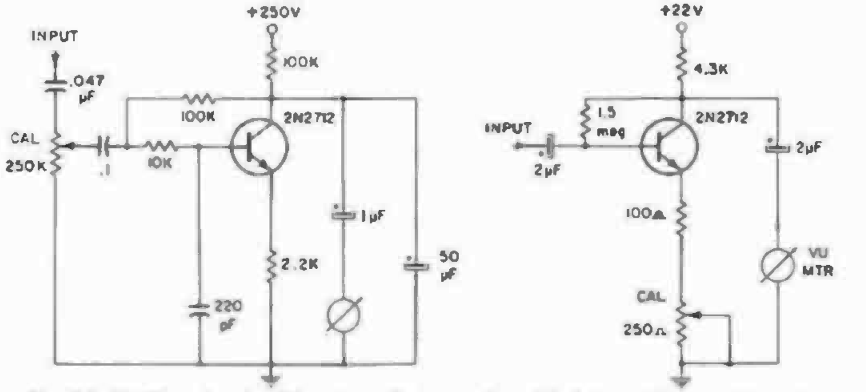


Fig. 10-43. Two simple VU meters that may be added to existing equipment.

red, thus warning the operator that the system is in overload. The scale reads in both percentage modulation and decibels.

10.42 Knowing the level in decibels and line impedance, how is the voltage determined for other line impedances?—If the line voltage for a given level at 600 ohms is known, voltages for other line impedances may be calculated:

$$E_x = V_{600}\sqrt{\frac{Z}{600}}$$

where,

E_x is the unknown voltage,
V_{600} is the voltage for 600-ohms.

As an example, assume voltage E_x is required for a line impedance of 150 ohms at a level of plus 4 dBm. Referring to Fig. 10-42, the voltage for a level of plus 4 dBm is 1.23 volts. The new voltage may now be calculated:

$$E_x = 1.23\sqrt{\frac{150}{600}} = 0.615 \text{ volt.}$$

Voltages for a line impedance of 600 ohms for levels between minus 20 dBm to plus 50 dBm may be taken from Fig. 10-42. Additional information is given in Question 23.167, and in Question 25.115.

10.43 Show a simple circuit for adding a VU meter to existing equipment.—Two such circuits each employing a

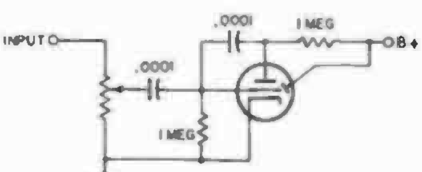


Fig. 10-44. A 6E5 or Magic Eye Tube connected as a volume indicator meter.

2N2712 transistor are shown in Fig. 10-43. Either of the circuits may be added to existing equipment. The meter movement can be the conventional VU meter with its internal rectifier or an inexpensive 200 to 500 microampere movement using a small rectifier similar to those discussed in Question 10.5. The completed meter is calibrated by applying a constant frequency of 1000 Hz to the input, and calibrating the meter movement in decibels. This subject is discussed in Section 23.

When the meter is put to use, the 100-percent modulation of the system is established, and the sensitivity of the meter circuit is increased by 10 dB to provide a 10-dB lead as explained in Question 10.28. It should be understood that unless the meter movement is an actual VU meter, the peak indications will not be compatible with the conventional VU meter. However, with a 10-dB lead, the system should be protected from serious overload.

10.44 How may a magic eye tube be connected for use as a volume indicator?—In the manner shown in Fig. 10-44. The terminal indicated input should connect to a point ahead of the recording device and be adjusted to indicate the peak recording level. (See Question 10.7.)

10.45 How is an oscilloscope used as a peak reading VI meter?—As discussed in Question 10.28, the standard VU meter does not indicate the true peaks of a complex waveform. To protect a recording channel from overload a lead of 8 dB or more is inserted in the VU meter circuit to protect the system from unseen peaks. A peak-reading VI meter may be had by connecting an oscilloscope in parallel with the VU

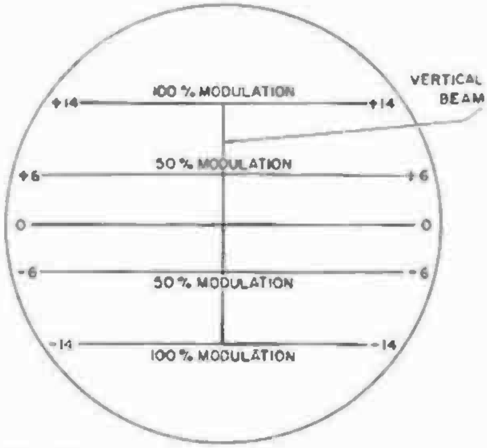


Fig. 10-45A. Oscilloscope screen calibrated for use as a peak-indicating VI meter. The horizontal sweep control is closed-off to present only a vertical line.

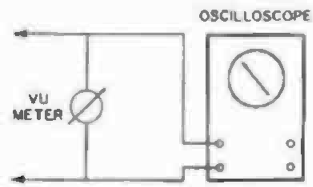


Fig. 10-45B. Oscilloscope connected in parallel with VU meter.

meter (Fig. 10-45B). To align the oscilloscope with the VU meter, two measurements are required.

A 1000-Hz signal is sent into the channel and the bridging bus level set, for example, to plus 14 dBm, and the deflection of the oscilloscope (sweep off) is adjusted for a convenient deflection (100%) on the graticule (Fig. 10-45A). The VU meter is now set for an 8-dB lead by turning its attenuator to plus 6 dBm. This level is also marked on the oscilloscope graticule. Now under normal recording conditions the oscilloscope will indicate the peak modulations. Thus, the readings of the VU meter for a complex waveform may be compared to the actual peak excursions not indicated by the VU meter. This method of monitoring recording levels is often used where the control of recording levels is critical.

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Section 11

Vacuum Tubes, Transistors, and Diodes

In 1883, Edison discovered that electrons flowed in an evacuated lamp bulb, from a heated filament to a separate electrode (the Edison effect). Fleming, making use of this principle, invented the "Fleming Valve" in 1905, but when de Forest, in 1907, inserted the grid, he opened the door to electronic amplification with the "Audion." The millions of vacuum tubes are an outgrowth of the principles set forth by these men.

The subject of vacuum tubes is complex, and only design considerations essential to proper use in audio circuitry are entered into. Semiconductors have resulted in the obsolescence of many type tubes; however, tubes will continue to play an important role in electronic circuitry for years to come. Vacuum tubes and transistors may be complementary, and many hybrid devices have been designed utilizing a combination of the two. While entirely different in concept, they may both perform the same function, and with modification of circuitry may be used interchangeably. However, the design engineer must use a different approach when considering devices employing either transistors or vacuum tubes.

Characteristics and variations in vacuum tubes, transistors, diodes, photoconductors, and other devices of the semiconductor family are discussed in this section.

11.1 What is the Edison effect?—The emission of electrons from a heated body. This effect was discovered by Thomas A. Edison in 1883 during his experiments with the electric light.

11.2 What are elements?—The inner electrodes of a vacuum tube.

11.3 What is a filament?—The element in a directly heated vacuum tube which emits electrons.

11.4 What is a heater?—A coiled element used to heat the cathode element in an indirectly heated vacuum tube.

11.5 What is thorium?—A rare mineral used in the manufacture of vacuum-tube filaments. Thorium when mixed with tungsten in the form of a filament is a profuse emitter of electrons, but gradually evaporates during its use.

11.6 What is barium oxide?—A rare earth used in a manner similar to thorium for coating the surface of a cathode or filament of a vacuum tube.

11.7 What is an indirectly heated tube?—One which employs a heater inside the cathode sleeve. The electrons are emitted from the surface of the cathode and not the heater. (See Fig. 11-7.)

11.8 What is a directly heated tube?—One employing a filament which serves the dual purpose of heater and cathode. The operating temperature of such tubes ranges from −900 to 2500 degrees Centigrade.

11.9 What is a getter?—A barium or tantalum disc enclosed in a vacuum

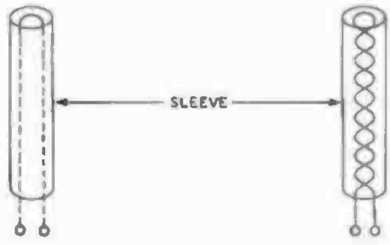


Fig. 11-7. Indirectly heated tube.

tube for the purpose of absorbing gasses released from the elements. The disc is flashed from the exterior during the exhausting process. This generally results in a silver or reddish discoloration on the interior of the glass envelope.

11.10 What is a cathode?—A sleeve surrounding the heater in a vacuum tube. The surface of the cathode is coated with barium oxide or thoriated tungsten to increase the emission of electrons. (See Fig. 11-7.)

11.11 What is a control grid?—A spiral wire element placed between the plate and cathode elements of a vacuum tube to which the input signal is generally applied. This element controls the flow of electrons between the cathode and the plate elements.

11.12 What is a suppressor grid?—A gridlike element situated between the plate and screen elements in a vacuum tube to prevent secondary electrons emitted by the plate from striking the screen grid. The suppressor is generally connected to the ground or cathode circuit.

11.13 What is a screen grid?—An element in a pentode-type vacuum tube which is situated between the control grid and the plate elements. This screen grid is maintained at a positive potential to reduce the capacitance existing between the plate and control-grid elements. It thus acts as an electrostatic shield and prevents self-oscillation and feedback within the tube.

11.14 What is a plate?—The positive element in a vacuum tube. The element from which the output signal is nsually taken. It is also called an anode.

11.15 What is a diode?—A two-element vacuum tube consisting of a plate and a cathode. It is also known as a Fleming valve, after its inventor Dr. J. A. Fleming, an English scientist.

11.16 What is triode?—A three-element vacuum tube.

11.17—What is a dual triode?—A single envelope containing two sets of triode elements.

11.18 What is a tetrode?—A four-element vacuum tube containing a cathode, a control grid, a screen grid, and a plate. It is frequently referred to as a screen-grid tube.

11.19 What is a hexode?—A six-element vacuum tube consisting of a cathode, control grid, suppressor grid, screen grid, injector grid, and a plate.

11.21 What is a heptode?—A seven-element vacuum tube containing a cathode, control grid, four grids, and a plate.

11.22 What is a pentagrid?—A seven-element vacuum tube consisting of a cathode, five grids, and a plate.

11.23 What is an octode?—An eight-element vacuum tube consisting of a cathode, six grids, and a plate.

11.24 Where are multigrid tubes used in audio circuits?—In compressors, expanders, and special applications.

11.25 What is a beam-power tube?—A power-output tube having the advantage of both the tetrode and pentode tubes. Beam-power tubes are capable of handling relatively high levels of output power for application in the output stage of an audio amplifier. The power-handling capabilities stem from the concentration of the plate-current electrons into beams of moving electrons. In the conventional tnbe the electrons flow from the cathode to the plate, but are not confined to a beam. In a beam-power tube the internal elements consist of a cathode, control grid, screen grid, and two beam-forming elements which are tied internally to the cathode element. The cathode is indirectly heated as in the conventional tube.

The internal construction of an RCA 6L6 beam-power tube is shown in part (a) in Fig. 11-25. One of the most important points of the construction of this tube is the method used to form the electron beams. This is accomplished by making the pitch of the winding of the control- and screen-grid elements the same, and in a physical plane relative to the paths of the plate-current electrons. Because of the control-grid winding shadow being in the same plane as the screen-grid winding, fewer electrons strike the windings of the screen-grid element. Thus the screen-grid current is less, permitting a greater number of electrons to reach the plate element, which results in a greater plate current for a given signal voltage at the control grid. The plate current being high results in the plate resistance being relatively low, thus increasing the power-handling capabilities. The beam-forming elements are

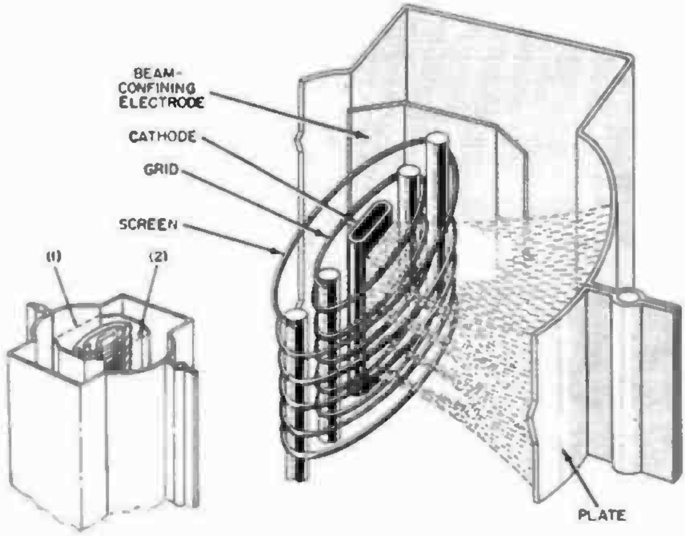
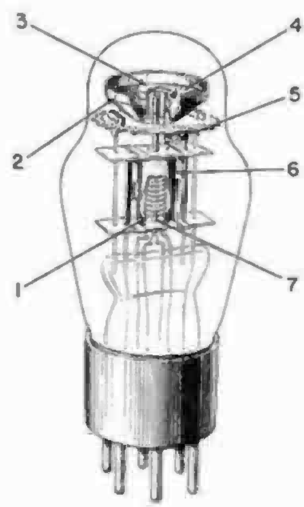


Fig. 11-25. View of the interior of a 6L6 beam-power tube. (Courtesy, Radio Corporation of America.)

connected internally to the cathode element and influence the movement of the electrons in their passage to the plate element. Because the beam-forming elements are at cathode potential, an equivalent space-charge effect is caused between the screen grid and the plate; this charge has an effect similar to that of a surface existing between the ends of one beam-forming plate and the other. This is shown by the dashed lines (points 1 and 2) in part (b) of Fig. 11-25. This effect is referred to as a virtual cathode. This invisible wall also repels secondary electrons from the plate and prevents them from striking the screen-grid element.

11.26 What is an electron-ray tube?—A miniature cathode-ray tube used as an amplitude indicator in recording and test equipment and also as a tuning indicator in radio receivers. Cathode-ray tubes are discussed in Question 11.91. The interior of an RCA 6E5 electron-ray tube is shown in Fig. 11-26.

11.27—Describe the construction of a nuvistor tube.—Nuvistor tubes (6CW4 and 7586) are used quite frequently in sound equipment, particularly where a high signal-to-noise ratio is a necessity. They are often used in the microphone preamplifier circuits with transistorized recording channels. When used as such, the circuitry is termed a *hybrid* circuit. Nuvistors have desirable characteristics for portable equipment, as they only require 0.135 amp at 6.3 volts for the heater, with a plate current of 10 milliamperes at 75 volts. The transconductance is 11,500 micromhos, with an amplification factor of 35. They also have the additional feature of being non-microphonic. An interior view showing the construction of such a tube is given in Fig. 11-27.



1—TRIODE GRID
2—TARGET
3—CATHODE LIGHT SHIELD
4—FLUORESCENT COATING
5—RAY-CONTROL ELECTRODE
6—TRIODE PLATE
7—CATHODE

Fig. 11-26. Interior view of a 6E5 electron-ray (magic eye) tube. (Courtesy, Radio Corporation of America.)

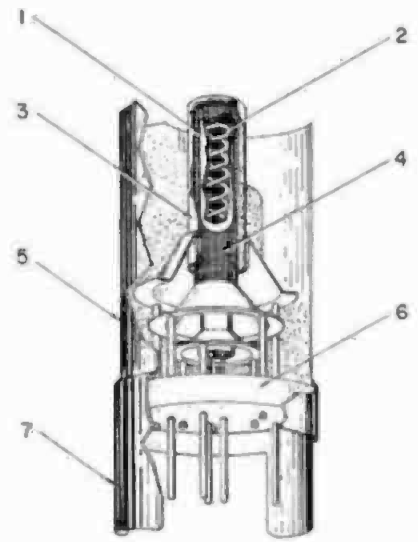


Fig. 11-27. Interior view of a 6CW4 or 7586 Nuvistor tube. (Courtesy, Radio Corporation of America)

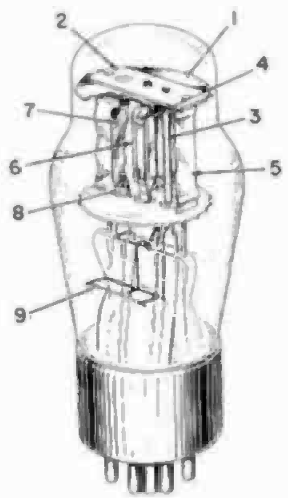
11.28 What is an acorn tube?—A small vacuum tube designed for very-high-frequency use and having a low internal electrode capacitance. The element connections are brought out at the sides of the glass envelope. This tube requires a special socket.

11.29 What is a thyratron?—A gas-discharge tube which may be used similarly to a relay. A positive charge is applied to a control grid which starts the control cycle. (See Fig. 11-29.)

11:30 What is a voltage-regulator tube?—A vacuum tube containing two elements, a cold cathode, and a plate and filled with a rare gas such as neon or argon. This tube requires no heater and is used as a voltage stabilizer in power supplies.

There are several different type voltage-regulator tubes (VRT). The two most useful to the audio engineer are the gas-filled and the glow-discharged types. Of these types, the voltage regulator and the voltage-reference tubes are the ones most commonly used. The voltage-reference tube is a special VRT, with a sharply defined voltage drop, which may be used for a source of calibration within limits. Voltage-regulator types are generally used in *shunt* with the circuit to be regulated, and act as a *variable* load, countering the effects of load variation. When there is little or no current drawn from the supply, the VRT conducts heavily. As the load current requirement increases, the VRT takes *less* current and so maintains a constant load on the supply. This effective flatting of the load current provides the voltage stabilization of which a VRT is capable. The average VRT draws from 5 to 30 milliamperes, and over this range of current will maintain the output voltage within 2 to 3 percent of the operating point.

Voltage-regulator tubes will also maintain a constant output voltage, even with variation in power-line voltages supplying the power supply. A rise in line voltage causes the VRT to draw more current, thus lowering the output voltage. Voltage-reference tubes are used where a higher order of control is required, as in recording equipment power supplies. These devices use high gain dc sensing amplifiers, which



1—SHIELDING MICA
2—INSULATING MICA
3—CATHODE
4—CONTROL GRID
5—SHIELD GRID
6—SHIELD GRID APERTURE
7—ANODE
8—GLASS SLEEVE
9—GETTER

Fig. 11-29. Interior view of a gas tetrode (thyratron) tube. (Courtesy, Radio Corporation of America.)

control the load current by means of heavy-current regulating tubes that accurately regulate the load current. The accuracy of the regulation is the function of the reference voltage tube. Voltage-reference tubes are similar to voltage-regulator tubes, except they contain certain modifications which restrict them to this form of service. The principal difference is that voltage-reference tubes are limited to a rather narrow range of current operation. However, as the reference tubes are designed to operate at a fixed current, this may be taken advantage of to adjust it to its most stable operating point. A second difference is that reference tubes operate at lower voltages than voltage-regulator tubes.

Voltage-regulator tubes are constructed with an anode consisting of a thin rod mounted in the center of a cylindrical cathode, which is the reverse of the conventional vacuum tube. The cathode in a VRT is cold since there is no heater element. The envelope is filled with an inert gas, such as argon or neon. When a rising voltage is applied across the tube, nothing happens until a critical voltage, called the firing potential or starting voltage, is reached. At this point the tube conducts. At this voltage the inside surface of the cathode becomes partially illuminated with either a purple or reddish glow, depending on the type of gas used. An increase of voltage across the tube causes an increase in current, increasing the area of the glow, which varies with the current and distinguishes it from the reference voltage tube which usually glows over the entire cathode surface.

Voltage-regulator tubes suffer from several inherent problems, the first of which is starting instability. It will be noted on the restarting of such tubes that they do not regulate at the same voltage for each start. This can be as much as plus or minus 2 or 3 volts in 148 volts. Another inherent fault is the flickering of the glow around the cathode as the current is increased from its minimum value to its maximum value. These sudden variations in voltage across the tube are the result of portions of the glow area skipping from one point on the cathode to another. A third form of instability is the operating current accidentally placed at one of the cathode skipping points. Voltage-regulator tubes also have a habit of oscillating at high frequencies or motorboating, depending on the circuit constants.

Although the above discussion does not present a very good picture of such tubes, they are still a valuable tool in the field of regulation. It is good practice to design the supply voltage to be high enough that the tubes will always fire when the initial voltage is applied; this causes a heavy surge of current and will probably regulate at the same voltage. Since a bypass capacitor is required around this tube, it should not be larger than 0.10 μF, since larger values tend to cause the tube to motorboat. Voltage-regulator tubes may also be operated in series or parallel. In series, they may be used as voltage dividers, or where the voltage to be regulated is higher than the specified VRT rating. The use of the voltage-regulator tube in regulated power supplies is discussed in Section 21.

It has been found that both voltage-regulator and voltage-reference tubes when placed in a light-tight box require a higher firing voltage than they do in daylight. In some instances they will not fire at all. For certain types of equipment, a small pilot light is placed near the tube to cause it to fire instantally when the voltage is applied. Typical voltage-regulator and voltage-reference tubes are the OD3/150, 6626, and 6627. Voltage-regulator tubes are also discussed in Section 21.

11.31 What is a phototube? — A two-element vacuum tube containing a cathode and a plate. Light falling on the cathode, which is coated with a light-sensitive substance, causes the tube to conduct. It is used for light control circuits and in the reproduction of sound tracks from motion picture film. The sensitivity of a phototube depends on the frequency or color of the light falling on its elements. Phototubes may be obtained that are sensitive only to certain bands of light. Others respond to the colors common to the human eye. Practical uses of the phototube are discussed in Question 19.93. The name *phototube* is used when referring to a vacuum-type photosensitive tube. The term *photocell* is used for solid-state devices and voltage-generating devices.

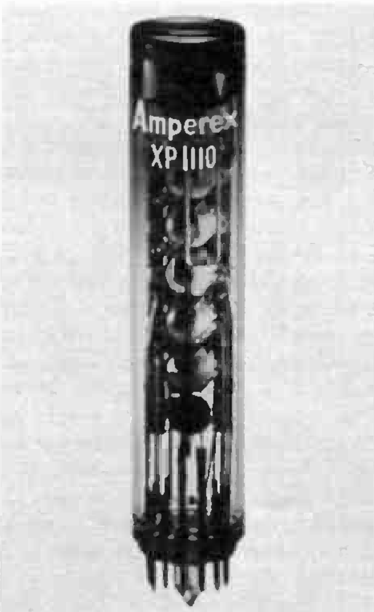


Fig. 11-32A. Amperex Electronic Corp. XP 1110 10-stage, photocathode photomultiplier tube.

11.32 What is a secondary-emission multiplier?—A phototube of extreme sensitivity. The initial electrons emitted from the cathode are directed to strike against a group of plates called dynodes.

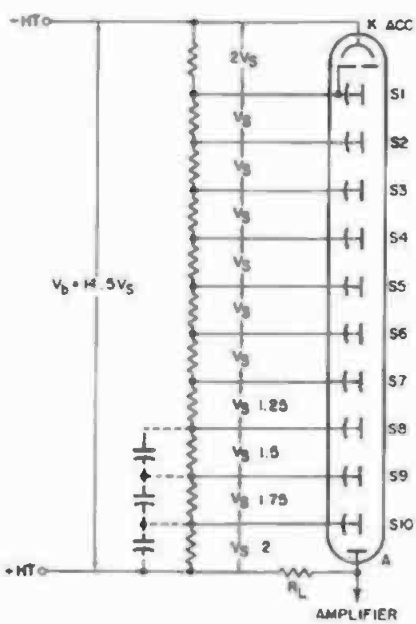


Fig. 11-32B. Voltage divider for Amperex XP 1110, 10-stage photocathode photomultiplier tube.

Each dynode is treated so that it will produce a high secondary-electron emission. When the first dynode is struck, several secondary electrons are released. These, in turn, strike the second dynode, releasing additional electrons. Photomultiplier tubes use 4- to 14-stage dynodes, achieving amplifications up to 100 million, with transit times of 28×10^{-9} to 70×10^{-9}, depending on the number of stages. The total voltage supply will range from 1800 to 7500 volts, again depending on the number of stages. A photocathode-type photomultiplier tube manufactured by Amperex, is shown in Fig. 11-32A, with its internal elements and connections shown in Fig. 11-32B.

11.33 What is the meaning of the term mutual conductance?—It is a term originated by Hazeltine in 1919 to express the conductance of a vacuum tube. This term has now been replaced by the term transconductance, discussed in Question 11.35.

11.34 Give the letter symbols used for identifying the electrical characteristics of vacuum tubes and basing diagrams.—The following symbols are in general usage.

C	Coupling capacitor between stages
C_{g2}	Screen grid bypass capacitor
C_k	Cathode bypass capacitor
E_{bb}	Supply voltage
E_{ff}	Plate efficiency
E_p	Actual voltage at plate
E_{sg}	Actual voltage at screen grid
E_o	Output voltage
E_{sig}	Signal voltage at input
E_g	Voltage at control grid
E_f	Filament or heater voltage
I_f	Filament or heater current
I_p	Plate current
I_k	Cathode current
I_{sg}	Screen grid current
I_{pa}	Average plate current
I_{ka}	Average cathode current
I_{sga}	Average screen grid current
G_m	Transconductance (mutual conductance)
mu	Amplification factor
P_{sg}	Power at screen grid
P_p	Power at plate
P-P	Plate load, plate-to-plate, or push-pull amplifier
R_g	Grid resistor
R_k	Cathode resistor
R_l	Plate load impedance or resistance
R_p	Plate load resistor

R_{sg} Screen dropping resistor
R_d Decoupling resistor
r_p Internal plate resistance
V_g Voltage gain

The above are just a few of the more commonly used symbols. It will be found although such symbols are fairly well standardized, considerable discrepancies exist in their usage. Basing diagrams for the more commonly used tubes are given in Fig. 11-34.

11.35 Define the term transconductance.—It is the change in the value of plate current expressed in microamperes divided by the signal voltage at the control grid of a vacuum tube and expresses its conductance. Conductance is the opposite of resistance and the name mho (ohm spelled backward) was adopted for this unit of measurement.

The basic unit, mho, is too large for practical usage; therefore, the term micromho is used. One micromho is equal to one-millionth of a mho, or 1000 micromhos equal 0.001 mho.

When a vacuum tube is measured in a dynamic transconductance tube tester, all constants of the tube are considered. An ac signal voltage is applied to the control grid and the tube is measured under simulated operating conditions and the actual conductance is measured. Emission-type tube testers do not measure the dynamic characteristics, only the emission of the emitting surfaces of the filament or cathode elements.

The transconductance (g_m) of a tube may be equated:

$$g_m = \frac{\Delta I_p}{\Delta E_{sig}} \ (E_{bb} \text{ held constant})$$

where,

ΔI_p is the change of plate current,
ΔE_{sig} is the change of control-grid signal voltage,
E_{bb} is the plate supply voltage.

A change of one milliampere of plate current for a change of one volt at the control grid is eqnal to one thousandth of a mho or one thousand micromhos.

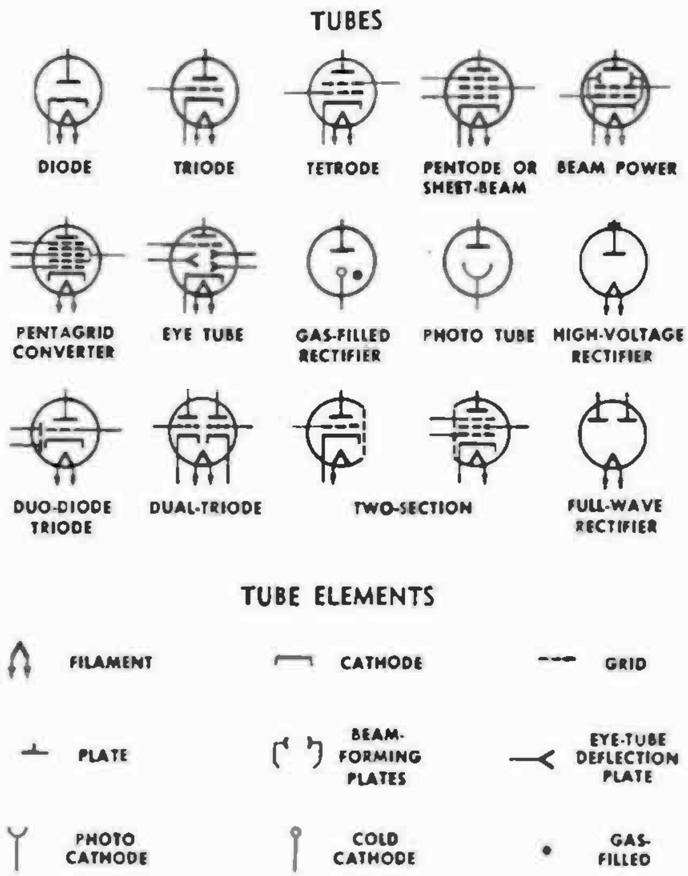


Fig. 11-34. Basing diagrams for vacuum tubes.

To convert mho's to micromhos multiply by 1,000,000. Thus, a tube having a change of two milliamperes plate current for a change of one volt at the control grid would have a transconductance of 2000 micromhos.

11.36 What is the amplification factor or voltage gain of a vacuum tube?—It is the amount the signal at the control grid is increased in amplitude after passing through the tube. This is also referred to as the mu, or voltage gain (V_g) of the tube.

Tube voltage gain may be computed:

$$V_g = \frac{g_m \times r_p \times R_p}{10^6 \times (r_p \times R_p)}$$

or

$$V_g = \frac{mu \times R_p}{r_p \times R_p}$$

where,

g_m is the transconductance in micromhos,
r_p is the plate resistance in ohms,
R_p is the load resistance in ohms,
mu is the voltage gain.

11.37 What does the term gain-per-stage mean?—It is the amount the signal is amplified after passing through a stage of amplification. If the amplifier consists of several stages, the amount of amplification is multiplied by each stage. The gain of an amplifier stage varies with the type tube and the interstage coupling used. The general equation for voltage gain is:

$$V_{gt} = V_{g1} \times V_{g2} \times V_{g3} \ldots$$

where,

V_g is the voltage gain of the individual stages.

11.38 What is the plate resistance of a vacuum tube?—The plate resistance (r_p) of a vacuum tube is a constant and denotes the internal resistance of the tube or the opposition offered to the passage of electrons from the cathode to the plate. Plate resistance may be expressed in two ways: the dc resistance and the ac resistance. The first is the internal opposition to the current flow when steady values of voltage are applied to the tube elements and may be determined simply by Ohm's law:

$$\text{dc } r_p = \frac{E_p}{I_p}$$

where,

E_p is the dc plate voltage,
I_p is the steady value of plate current,
r_p is the plate resistance.

The ac resistance is not so simple to calculate but requires a family of plate-current curves from which the information may be extracted. As a rule, this information is included with the tube characteristics and is used when calculating or selecting components for an amplifier. The equation for calculating ac plate resistance is:

$$\text{ac } r_p = \frac{\Delta E_p}{\Delta I_p}$$

where,

E_p is the voltage at the plate,
I_p is the plate current,
Δ (Delta) is the change in E_p and I_p with the control grid signal voltage (E_{sig}) held constant.

The values of E_p and I_p are those taken from the family of curves supplied by the manufacturer for the particular tube under consideration.

11.39 How is the plate resistance affected by a change of element voltage?—Increasing the plate voltage or decreasing the grid-bias voltage decreases the plate resistance.

11.40 What are steady-state or static characteristics?—The characteristics of a vacuum tube obtained when all normal voltages are applied to the elements with no signal at the input. This is tested when using an emission-type tester.

11.41 What are dynamic characteristics?—Applied to a vacuum tube or transistor, dynamic characteristics are a plot of the steady-state characteristics to which have been added the effect of load impedance. The resultant plate-current and grid-voltage curves are termed *dynamic-transfer characteristics*. Dynamic-transfer curves may be used to determine the linearity and nonlinearity of the input signal compared to the input signal for a specific operating point for a given load resistance or impedance.

11.42 Define the term perveance.—Perveance is a figure of merit for a diode rectifier tube. High-perveance tubes have a lower internal voltage drop for a given fixed current level. This information is generally available in the manufacturer's tube manual.

11.43 What is the cutoff point of a vacuum tube?—A condition when the negative potential applied to the control grid completely stops flow of electrons between the cathode and plate elements.

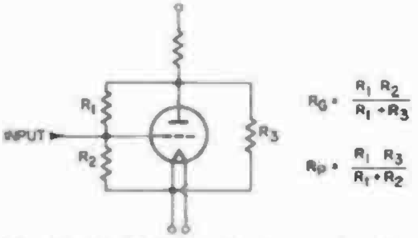


Fig. 11-44. Showing the internal resistance of a vacuum tube which can cause the flow of grid current.

11.44 What is grid current and its cause?—If the control grid is permitted to become positive, with respect to the cathode, it results in a flow of current between the control grid and the cathode through the external circuits. This condition is unavoidable because the wires of the control grid, having a positive charge, attract electrons passing from the cathode to the plate.

Grid-current flow in a vacuum tube is generally thought of as being caused by only driving the control grid into the positive region, and causing the flow of grid current. This is quite true for the operating conditions of an amplifier, but there are other causes of grid current not directly associated with the operation of the tubes. The frequent lack of recognition of grid current hazards is the simplified description of a vacuum tube as a device which operates with a negative grid bias and draws no current from the signal source. This is not entirely true.

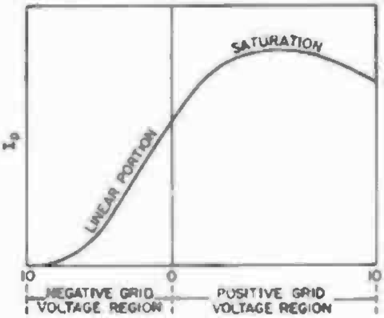


Fig. 11-48. Saturation curve for a typical triode tube.

There are at least eight distinct forms of grid-emission currents, which may be classified: interelectrode resistance, gas-ionization current, grid-emission current, positive grid-electron current, secondary grid-emission current, and negative grid-electron current.

During certain phases of the manufacturing process, such as exhausting and aging, the cathodes are operated at a high temperature which, in turn, produces minute evaporation of the nickel-cathode material and other conductive materials from its coating. The material migrates to a relatively cool surface of the mica supports where it builds up a layer of conductive material and thus shunts a resistance across certain electrodes. In addition, the *getter* material, which is usually a thin coating of barium or magnesium, is deposited on a portion of the glass envelope, or inadvertently deposited on a micro-support area. This resistance is not linear and at higher temperatures and higher voltages, conductance is increased. The leakage paths caused by the foregoing are shown in Fig. 11-44. Grid current is highly undesirable, and can be the cause of considerable distortion.

11.45 How are triode tubes classified for amplifiers?—By their amplification factor (mu or μ). A low-mu tube is one having an amplification factor less than 10. Medium-mu tubes have an amplification factor of from 10 to 50, with a plate resistance of 5 to 15,000 ohms. High-mu tubes have an amplification factor of 50 to 100, with a plate resistance of 50,000 to 100,000 ohms.

11.46 What is anode current?—This is another name for plate current.

11.47 What is secondary emission?—A condition which occurs when electrons are released from a body that is being bombarded by electrons.

11.48 What is meant when it is said a vacuum tube is driven to saturation?—When the maximum emission current has been reached, as shown by the graph in Fig. 11-48.

11.49 What does the term maximum plate dissipation mean?—The maximum power that can be dissipated by the plate element before damage occurs. Plate dissipation can be calculated:

$$\text{Watts dissipation} = E_p \times I_p$$

where,

E_p is the voltage at the plate,
I_p is the plate current.

11.50 What is meant by the expression "a soft vacuum tube"?—A tube in which a small amount of gas remains after evacuation. The retention of gas will often affect the characteristics and cause a blue glow between the plate and other elements due to ionization. This glow should not be confused with the glow seen in mercury rectifiers, gas regulator tubes, and certain type pentode power tubes, as these types of tubes glow when operating normally.

11.51 What is a hard tube?—A vacuum tube having an almost perfect vacuum.

11.52 What are grid-voltage, plate-current characteristic curves for a vacuum tube?—A group of curves, as shown in Fig. 11-52, plotted to show the change in plate current for a change in negative grid bias on the control grid. The curves shown are for a medium-mu triode.

The curves indicate that for a given plate current the plate and grid bias may be determined. For example: the manufacturer states that for a plate voltage of 250 volts and a negative grid bias of 8 volts, the plate current will be 9 milliamperes. This is indicated at Point A on the 250-volt curve. If it is desired to operate this tube with a plate voltage of 150 volts and still maintain a plate current of 9 milliamperes, the grid bias will have to be changed to a negative 3 volts.

11.53 What is ionization?—When an electron in a vacuum tube collides with a gas molecule, the energy imparted by the impact causes the molecule to release an electron. This molecule is called an ion. A gas containing no ions is almost a perfect insulator. Under such conditions no current will flow between elements. However, gases do have some residual ionization due to the action of light falling on the elements and to cosmic rays. If a potential is applied between elements in such a gas, ions migrate between the elements giving the effect of current flow. This current flow is called space, or dark current, because it cannot be observed. This current flow is very small, generally about one to two microamperes in the average tube. Once ionization takes place, the current increases to such proportions that serious damage to the tube can result. Ions are both positive and negative.

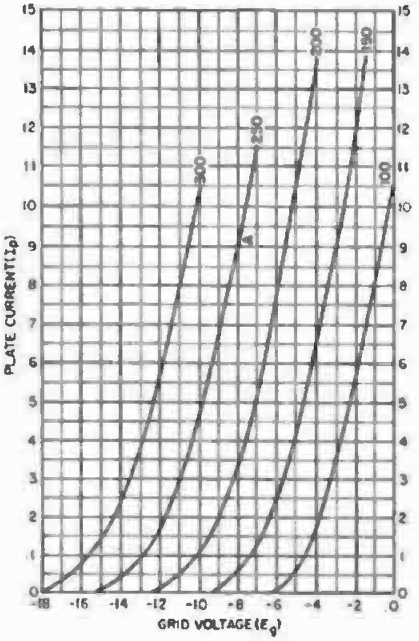


Fig. 11-52. Grid-voltage, plate-current curves for a triode tube.

11.54 What is ionization and deionization potential?—Ionization potential is the potential at which ionization takes place in a gas-filled tube, such as a voltage regulator or voltage-reference tube. Deionization potential is the potential at which ionization of the gas ceases and conductance stops.

11.55 What is the purpose of preheating vacuum tubes?—It is the custom in most plants employing a large number of vacuum tubes in critical places, such as recording and reproducing equipment, to preheat tubes and cook them for at least 100 hours before putting them into service. This is accomplished by connecting the plate, screen grid, suppressor grid, and control grid to one side of the filament or heater circuit (Fig. 11-55). The heater circuit is supplied from a source of voltage regulated to within 1 percent of the rated heater voltage. No plate voltage is used during the cooking period.

After removal from the preheating process, the tubes are individually

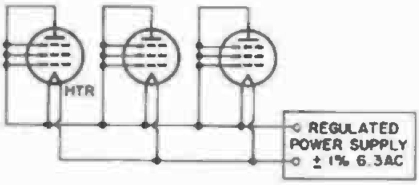


Fig. 11-55. Circuit for cooking vacuum tubes to stabilize their characteristics before putting them into operation.

tested for internal shorts, noise, and leakage, then selected in pairs for devices such as compressors, noise-reduction amplifier, and other equipment requiring matched pairs of tubes.

It has been found from experience when equipment is operated 24 hours a day, tube life is prolonged, internal noise reduced, and the general characteristics are stabilized. (See Ques. 11.71.)

11.56 What is the space charge in a vacuum tube?—In the flow of electrons from cathode to plate, not all the electrons are collected by the plate element. The uncollected electrons congregate in the space between the cathode and plate, and form a negative cloud. Other electrons leaving the cathode have to penetrate this cloud. The clond being negative, uegative charges leaving the cathode on their way to the plate are repelled back toward the cathode. It is these electrons which form the space charge. The space charge around the cathode controls the number of electrons that reach the plate, and therefore, controls the amount of plate current flow. The electron flow can only be increased by increasing the voltage at the plate of the tube.

11.57 Define inverse-peak voltage (IPV).—It is the highest instantaneous plate voltage which a tube can withstand in the *opposite* direction of its normal current flow. In a rectifier tube, it is the maximum value of plate voltage that may be applied, without internal arc-back between the elements, for a given operating temperature.

11.58 What is a microphonic tube?—A tube which, when tapped or vibrated, produces a ringing sound. As a rule, microphonics in a vacuum tube are caused by loose elements within the tube, or these elements may not be sufficiently rugged for the purpose for which the tube is being used.

Special tubes which have no microphonic tendencies are available for use in the low-level stages of an amplifier.

11.59 What causes self-rectification in a vacuum tube?—An increase or decrease in the dc plate current caused by an unsymmetrical waveform. (See Question 12.44.)

11.60 What are the phase reversals that take place between the elements of a vacuum tube?—The phase reversal in electrical degrees between the elements of a self-biased pentode, for a given sine at the control grid is shown in Fig. 11-60A. It will be noted that, for an instantaneous positive voltage at the control grid, the voltage is reversed between the grid and plate 180 degrees and will remain so for all normal operating conditions. The control grid and

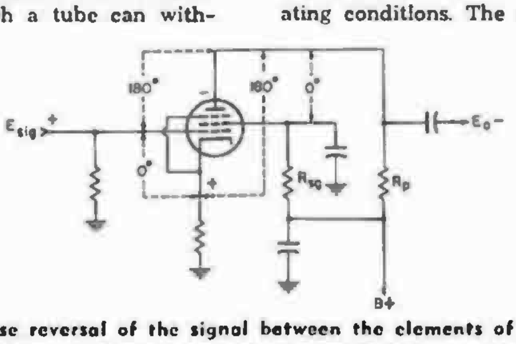


Fig. 11-60A. Phase reversal of the signal between the elements of a pentode vacuum tube. The reversals are the same in a triode for a given element.

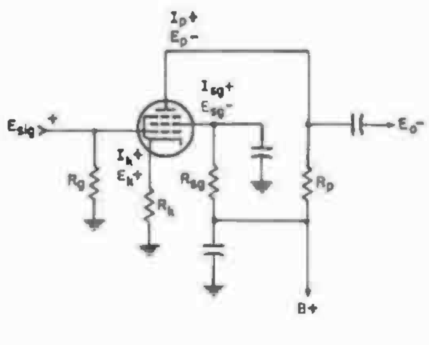
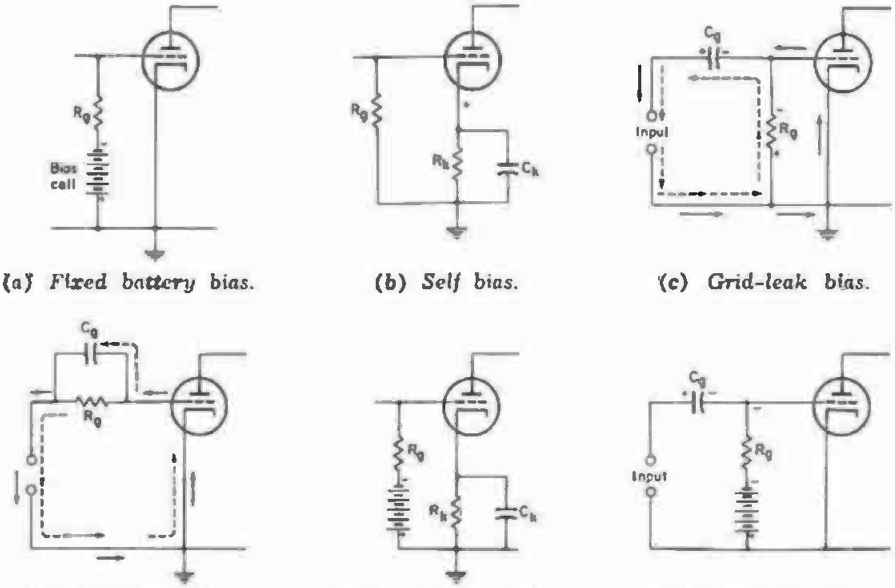


Fig. 11-60B. Phase reversal of the current and voltage in a pentode vacuum tube. The reversals are the same in a triode for a given element.



(a) *Fixed battery bias.* (b) *Self bias.* (c) *Grid-leak bias.*

(d) *Grid-leak bias.* (e) *Combination bias.* (f) *Combination bias.*

Fig. 11-62. Various methods of obtaining grid bias.

cathode are in phase. The plate and screen-grid elements are in phase with each other. The cathode is 180 degrees out of phase with the plate and screen-grid elements.

The phase reversal of the voltage and current for each element is shown in Fig. 11-60B. For an instantaneous positive sine at the control grid, the voltage at the plate and screen grid are negative and the current positive. The voltage and current are both positive in the cathode resistor and are in phase with the voltage at the control grid. The plate reversals are the same in a triode for a given element.

11.61 What is the difference between the terms plate voltage and plate-supply voltage?—Plate voltage is the voltage measured between the plate and the cathode elements. Plate-supply voltage is the voltage supplied to the lower end of the plate load. The first voltage is designated by the symbol E_p, the second by the symbol E_{bb}. In resistance-coupled amplifiers, the voltage considered is the plate-supply voltage E_{bb} at the lower end of the plate-load resistance. In a transformer-coupled amplifier, the voltage considered is the plate voltage E_p.

11.62 What are the different methods used for obtaining grid-bias voltage?—The six methods most commonly used are illustrated in Fig. 11-62. At part (a), a bias cell is connected in series with the control grid. At part (b), the tube is self-biased by the use of a resistor connected in the cathode circuit. In part (c), the circuit is also a form of self-bias; however, the bias voltage is obtained by the use of a grid capacitor and grid-leak resistor connected between the control grid and ground. At part (d), the bias voltage is developed by a grid-leak resistor and capacitor in parallel, connected in series with the control grid. The method illustrated at part (e) is called combination bias and consists of self-bias and battery bias. The resultant bias voltage is the negative voltage of the battery and the bias created by the self-bias resistor in the cathode circuit. Another combination bias circuit is shown at part (f). The bias battery is connected in series with the grid-leak resistor in this case. The bias voltage at the control grid is that developed by the battery and the self-bias created by the combination of the grid resistor and capacitor.

11.63 Show how and where the internal capacitance of a vacuum tube is created.—The internal capacitance of a vacuum tube is created by the close proximity of the internal elements as shown in Fig. 11-63. Unless otherwise stated by the manufacturer of the tube, the internal capacitance of a glass tube is measured using a close-fitting metal

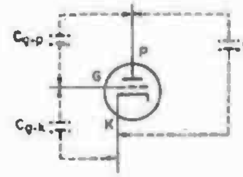


Fig. 11-63. Interelectrode capacitance of a triode.

tube shield around the glass envelope connected to the cathode terminal. Generally, the capacitance is measured with the heater or filament cold with no voltage applied to any of the other elements.

In measuring the capacitance, all metal parts, except the input and output elements, are connected to the cathode. These metal parts include internal and external shields, base sleeves, and unused pins. In testing a multisection tube (dual), elements not common to the section being measured are connected to ground. Input capacitance is measured from the control grid to all other elements, except the plate which is connected to ground. Output capacitance is measured from the plate to all other elements, except the control grid which is connected to ground.

Grid-to-plate capacitance is measured from the control grid to the plate, with all other elements connected to ground.

11.64 What is the terminology for batteries used in connection with vacuum tubes?—"A" battery for filament or heater, "B" battery for plate supply, "C" battery for bias voltages, and "D" battery for screen-grid voltages.

11.65 What is a variable-mu tube?—A pentode vacuum tube constructed with a special grid. In the conventional tube, the mu is constant. The construction of the variable-mu tube is such that the mu may be varied or changed by increasing the negative bias on the control grid. This type of tube is used in compressor amplifiers and expanders. A typical example is the 6SK7 tube.

11.66 What is transit time?—The time required for an electron to travel from the cathode to the plate circuit in a vacuum tube.

11.67 Define contact potential.—In a vacuum tube employing a unipotential cathode, positive grid current begins to flow when the control grid becomes slightly negative, and increases rapidly as the control grid is made more positive (Fig. 11-67). The value of grid voltage at which positive grid current starts to flow, is termed contact potential. Contact potential is caused by the initial velocity of the emission of electrons from the cathode and an electrothermic effect, because of the differences in temperature and the material composition of the control grid and the cathode element.

The value of the contact voltage may reach 1.5 volts. If the operating bias voltage is less than the contact potential, two effects will be present; direct-current flow in the control-grid circuit, and the dynamic input resistance of the tube drops to a relatively low value. The tube should be operated with a value of grid-bias voltage sufficiently high, so that the tube is not operating in the contact potential region. When a tube must be operated to within the contact potential region, care must be taken to avoid undesirable effects in the control-grid circuit, due to grid-current flow, and the lower input resistance. (See Question 11.44.)

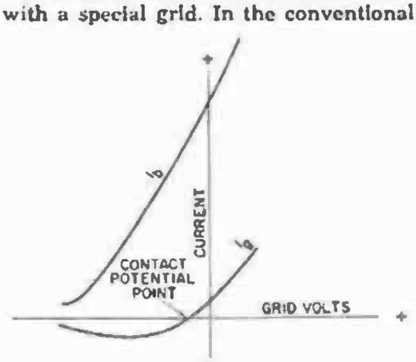


Fig. 11-67. Positive grid-current flow in a tube employing a unipotential cathode.

11.68. What is the direction of current in a vacuum tube?—From the cathode or filament to the anode or plate. This is termed *electron flow*, and is opposite to *conventional current flow*. (See Question 11.109.)

11.69 What is a barretter and a ballast tube?—A barretter is a voltage-regulator tube consisting of an iron-wire filament enclosed in a hydrogen-filled envelope. The filament is connected in series with the circuit to be regulated. For a given voltage variation, the current through the filament is held constant to a given value. Such tubes are used in series with the primary of a transformer and in series-connected filament circuits. Barretters must not be operated uear a strong magnetic field as the filament will vibrate and reduce the life of the filament.

A ballast tube is self-regulating, with its elements enclosed in a glass envelope, without gas, and fitted with an octal tube base. When used, it is connected in series with the circuit it is to regulate. Such regulators as the barretter and ballast tube are rather slow in their reaction to changes in current, compared to other devices. They are used principally in heavy commercial control equipment.

11.70 What is bias voltage?—A dc voltage held constant between the control grid and the cathode of a vacuum tube. (See Question 11.62.)

11.71 How may vacuum-tube characteristics be stabilized?—Tube manufacturers' data sheets generally contain a warning that the heater voltage should be maintained within plus or minus 10 percent of the rated voltage. As a rule, this warning is taken lightly, and little attention is paid to heater voltage variations, which have a pronounced effect on the tube characteristics, internal noise being the greatest offender. Because of heater-voltage variation, emission life is shortened, electrical leakage between elements is increased, heater-to-cathode leakage is increased, and grid currents caused to flow (see Question 11.44). Thus, the life of the tube is decreased with an increase of internal noise. A plot of the loss of tube life compared to percent heater-voltage variation appears in Fig. 11-71.

To reduce the effect of heater-voltage variation, heaters must be operated from a regulated source of voltage, either ac or dc. In large recording plants, it is customary to employ power supplies that furnish a regulated dc voltage, plus or minus 1 percent of the rated heater voltage. Tests have proven that tubes operated for two years or more, using such supplies, have lower internal noise levels than when the tubes were new. (See Question 11.55.)

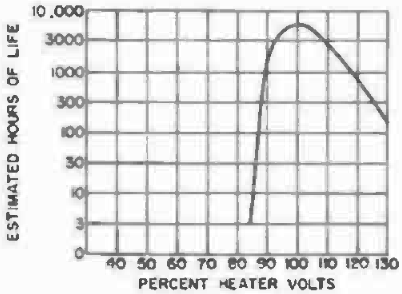


Fig. 11-71. A plot of tube life versus heater-voltage variation in percentage of rated voltage.

11.72 How are vacuum tubes connected in parallel?—As shown in Fig. 11-72. If two tubes of similar characteristics are connected in parallel, the plate current is doubled, the load resistance and bias resistor values halved. The value of the plate voltage remains unchanged. For more than two tubes, the plate current will increase proportionately; the load resistance will decrease in proportion to the number of tubes connected in parallel. This is also true for the bias resistor. As a rule, when vacuum tubes are connected in parallel, resistors of 50 ohms are connected in series with each plate to assist in balancing the plate currents. Also, a resistance of 1000 ohms is connected in series with each control grid (at the socket) to prevent parasitic oscillations. This is particularly important when an odd number of tubes are connected in parallel. The parallel operation of tubes in amplifiers is discussed in Questions 12.60 and 12.111.

11.73 What does the term quiescent mean when applied to a vacuum tube?—A condition that prevails in a vacuum tube when there is no signal voltage at the control grid. The plate current is a steady value.

11.74 Show the interior construction of a typical metal-envelope vacuum tube.—The interior of a metal-envelope vacuum tube with the various elements and their construction marked for identification is shown in Fig. 11-74.

11.75 Show the interior of a typical glass-envelope vacuum tube.—The in-

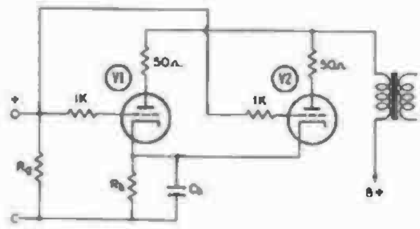
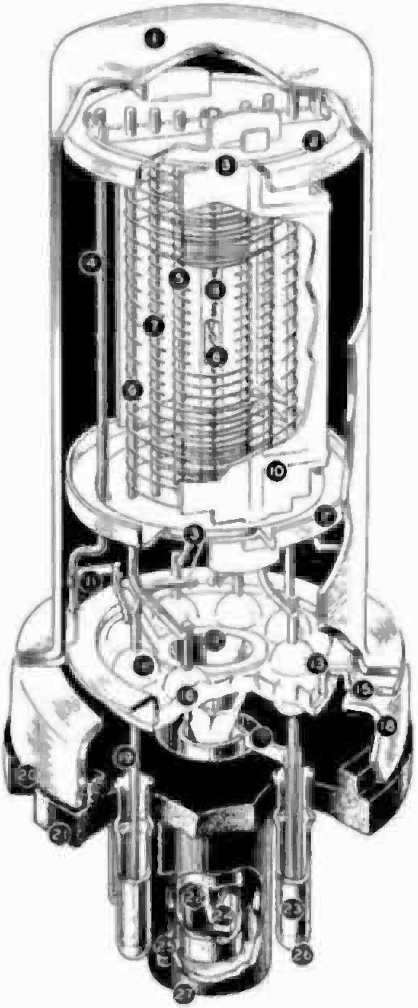


Fig. 11-72. A parallel-connected amplifier-output stage.



1—METAL ENVELOPE
2—SPACER SHIELD
3—INSULATING SPACER
4—MOUNT SUPPORT
5—CONTROL GRID
6—COATED CATHODE
7—SCREEN
8—HEATER
9—SUPPRESSOR
10—PLATE
11—BATALUM GETTER
12—CONICAL STEM SHIELD
13—HEATER INSERT
14—GLASS SEAL
15—HEADER
16—GLASS-BUTTON STEM SEAL
17—CYLINDER BASE SHIELD
18—HEADER SKIRT
19—LEAD WIRE
20—CRIMPED LOCK
21—OCTAL BASE
22—EXHAUST TUBE
23—BASE PIN
24—EXHAUST TIP
25—ALIGNING KEY
26—SOLDER
27—ALIGNING PLUG

Fig. 11-74. View of the interior of a single-ended envelope vacuum tube. (Courtesy, Radio Corporation of America.)

terior of a typical RCA glass-envelope vacuum tube is shown in Fig. 11-75.

11.76 Show the construction of a miniature tube.—The interior construction of an RCA miniature tube is shown in Fig. 11-76.

11.77 If it is desired to operate a vacuum tube with a plate voltage different from the published data, how are the new bias and screen voltages, plate-load resistance, and other characteristics determined?—By the use of conversion factors F_1, F_2, F_3, F_4, and F_5. Assume the following conditions are specified for a single 6V6 beam-power tube:

Plate voltage	250 volts
Screen voltage	250 volts
Grid voltage	−12.5 volts
Plate current	45 μA
Screen current	4.5 μA
Plate resistance	52,000 ohms
Plate load	5,000 ohms
Transconductance	4,100μmhos
Power output	4.5 watts

The new plate voltage is to be 180 volts. The conversion factor F_1 for this voltage is obtained by dividing the new plate voltage by the published plate voltage:

$$F_1 = \frac{180}{250}$$
$$= 0.72.$$

The screen and grid voltage will be proportional to the plate voltage:

$$E_g = 0.72 \times 12.5$$
$$= 9.0 \text{ volts}$$
$$E_{sg} = 0.72 \times 250$$
$$= 180 \text{ volts.}$$

For calculating the plate and screen currents, factor F_2 is used:

$$F_2 = F_1 \times \sqrt{F_1}$$
$$= 0.72 \times 0.848$$
$$= 0.610,$$
$$I_p = 0.61 \times 45\ \mu A$$
$$= 27.4\ \mu A$$
$$I_{c2} = 0.61 \times 4.5\ \mu A$$
$$= 2.74\ \mu A.$$

The plate load and plate resistance may be calculated by use of factor F_3:

$$F_3 = \frac{F_1}{F_2}$$
$$= \frac{0.720}{0.610}$$
$$= 1.18,$$
$$r_p = 52{,}000 \times 1.18$$
$$= 61{,}360 \text{ ohms},$$
$$R_L = 5000 \times 1.18$$
$$= 5900 \text{ ohms}.$$

The power output is found by the use of the factor F_4:

$$F_4 = F_1 \times F_2$$
$$= 0.72 \times 0.610$$
$$= 0.438,$$
$$\text{Power Output} = 0.438 \times 4.5$$
$$= 1.97 \text{ watts}.$$

The transconductance is determined by the aid of factor F_5:

$$F_5 = \frac{1}{F_3}$$
$$= \frac{1}{1.18}$$
$$= 0.847,$$
$$\text{transconductance} = 0.847 \times 4100$$
$$= 3472 \text{ micromhos}.$$

The foregoing method of converting for voltages other than those originally specified may be used for triodes, tetrodes, pentodes, and beam-power tubes, provided the plate, grid 1, and grid 2 voltages are changed simultaneously by the same factor. This will apply to any class of tube operating such as class A, AB_1, AB_2, B, or C. Although the method of conversion outlined in the foregoing is quite satisfactory in most instances, it should be borne in mind that the error will be increased as the conversion factor departs from unity. The most satisfactory region of operation will be between 0.7 and 2.0. When the factor falls outside this region the accuracy of operation is reduced.

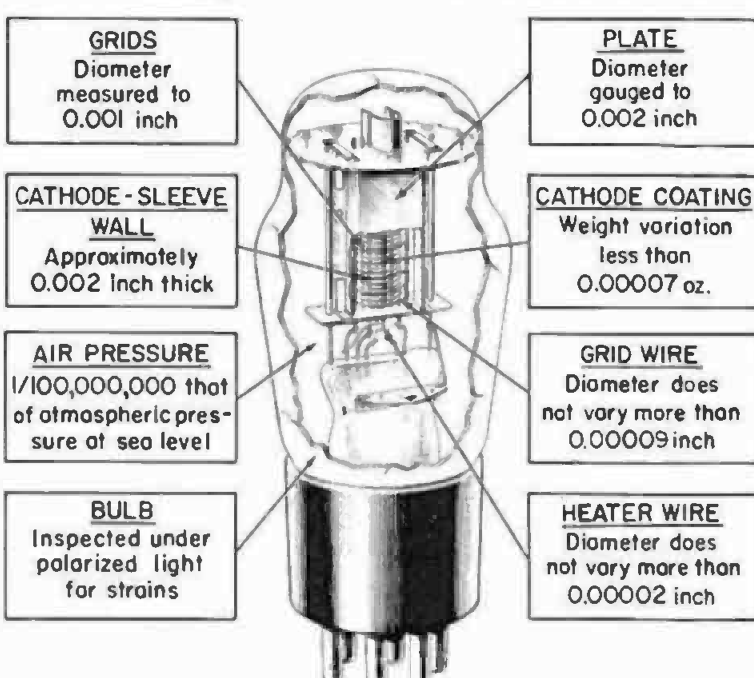


Fig. 11-75. Materials used in the construction of a vacuum tube. (Courtesy, Radio Corporation of America)

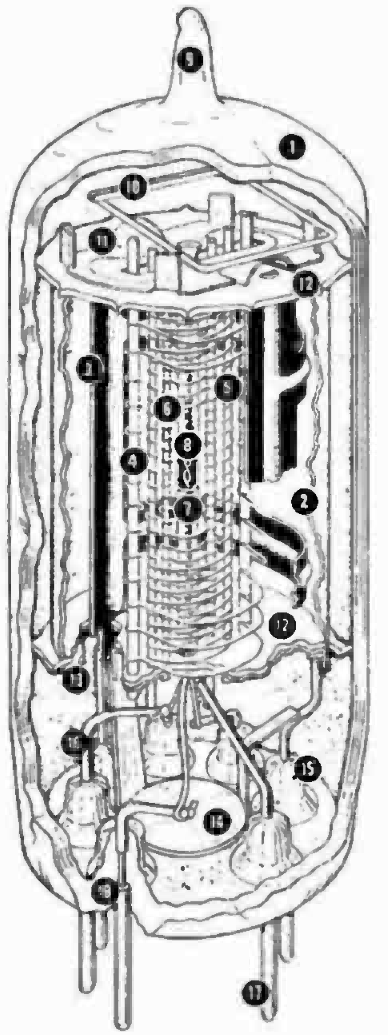


Fig. 11-76. View of the interior of a miniature vacuum tube. (Courtesy, Radio Corporation of America)

The above method of conversion does not take into consideration contact potential or secondary emission effects which become quite noticeable at low operational voltages. When secondary emission effects are noted, the plate voltage must be operated at a higher potential than grid number 2.

The use of low plate voltage and current in a beam-power tube is not recommended. Under normal conditions, the harmonic distortion should be essentially the same as for the originally specified operational voltages.

11.78 How can the heater or filament of a vacuum tube be tested for life reserve?—With a normal indication on the meter of a tube tester, reduce the filament or heater voltage to the values shown in the table below. If the tube with the lowered heater voltage drops and then holds steady within prescribed limits for the tube, it still has a considerable life reserve.

Normal Heater Voltage	*Reduce Heater Voltage to*
1.5	1.1
2.0	1.5
2.5	2.0
3.0	2.5
5.0	4.3
6.3	5.0
7.5	6.3
10.0	7.5
12.6	10.0
25.0	20.0
35.0	25.0
50.0	35.0

11.79 What is heater warmup time and how is it defined?—It is the time required for the voltage across the heater to reach 80 percent of the normal voltage, with four times the normal voltage applied to the heater circuit with a series resistor three times the normal heater operating resistance in series.

11.80 Describe the cause and effect of cathode leakage.—Cathode leakage is caused by internal leakage between the tube elements to the filament or heater circuit. This results in *hum modulation* of the signal voltage, as the heater of a cathode-heater type tube may have one terminal at cathode potential. The opposite heater terminal of a 6.3-volt tube will have a peak negative voltage of approximately 10 volts, at the 60-Hz heater supply. One section of the heater wire, just at the edge or outside on one end of the cathode and electrically near the ungrounded end, may have contaminants deposited during manufacture, which produce leakage and grid emission (see Question 11.44). Even without the contaminants, the 1200° C temperature of the heater wire produces an emission. If such an area is exposed to one of the grid-support side rods outside the mica support, a negative grid bias of less than 10 volts will permit the control grid to pick up the emission at the heater supply frequency.

The effects of cathode leakage can be overcome by the use of a bias voltage on the heater circuit, as discussed iu Question 12.195. A new medium power tube will show a normal leakage between the cathode and heater of about 1 to 2.5 megohms; small receiving tubes show 2 to 3 megohms and higher.

11.81 Why is it necessary to warm up the heater of a mercury vapor rectifier tube before applying the high voltage?—To prevent the active material on the filaments or heater from excessive drain while operating under a subnormal heater voltage.

11.82 When should a tube be considered unsatisfactory and rejected?—When the transconductance (g_m) has fallen to 75 or 80 percent of normal. For some types of tubes and services, they may be permitted to fall to 60 percent of normal.

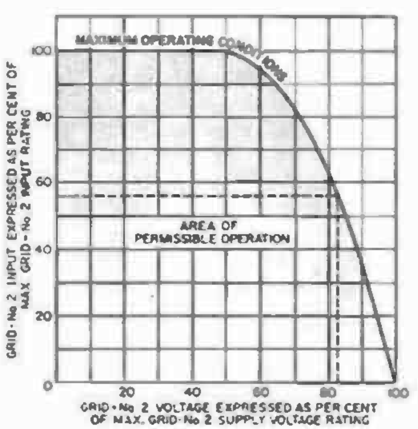


Fig. 11-86. Graph for determining the maximum power that may be dissipated by the screen grid. (Courtesy, Radio Corporation of America.)

11.83 Define the power sensitivity of a tube.—Power sensitivity is the ratio of the power output to the square of the input voltage, expressed in mhos:

$$\text{Power sensitivity (mhos)} = \frac{P_o \text{ watts}}{(E_{in}\text{rms})^2}$$

where,

P_o is the power output of the tnbe,
E_{in} is the rms signal voltage at the input.

11.84 How is plate efficiency calculated?—The plate efficiency (E_{ff}) for any tube may be calculated:

$$E_{ff} = \frac{\text{Watts}}{E_{pa} \times I_{pa}} \times 100$$

where,

Watts is the power output,
E_{pa} is the average voltage at the plate,
I_{pa} is the average plate current.

The measurement is made with a load resistance in the plate circuit equal in value to the plate resistance stated by the manufacturer.

11.85 When measuring the transconductance of a vacuum tube, what is the value of the standard signal voltage applied to the control grid?—Ten millivolts at the control grid. The output cnrrent is measured in ac milliamperes.

11.86 How is a screen-grid series-dropping resistor calculated?—The first step is to refer to the manufacturer's data sheet and find the maximum voltage that may be applied and the maximum power that may be dissipated by the screen grid. These limitations are generally shown graphically as in Fig. 11-86. The value of the resistor may be calculated:

$$R_{sg} = \frac{E_{sg} \times (E_{BB} - E_{s1g})}{P_{sg}}$$

where,

R_{sg} is the minimum value for the screen-grid voltage dropping resistor in ohms,
E_{sg} is the selected value of screen-grid voltage,
E_{BB} is the screen-grid supply voltage,
P_{sg} is the screen-grid input in watts corresponding to the selected value of E_{sg}.

A typical example follows:

Assume the data for a given tube states that a maximum of 300 volts can be applied to the screen grid and that the power dissipation cannot exceed 1 watt. It is desired to operate the screen grid with a voltage of 250 volts between cathode and screen, this voltage to be

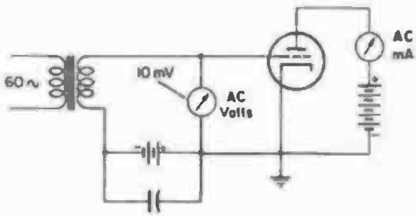


Fig. 11-87. Circuit for measuring the transconductance (G_m) of a typical vacuum tube.

obtained by the use of a series resistance between the power supply and the screen grid.

Because 250 volts is 83 percent of 300 volts, the maximum screen-grid voltage must be limited, as shown on the graph in Fig. 11-86, to 56 percent of the maximum screen-grid input, or 0.56 watt. The minimum value of the series-dropping resistor will be:

$$R_{sg} = \frac{250 \times (300 - 250)}{0.56} = 22{,}320 \text{ ohms.}$$

11.87 How is the transconductance of a vacuum tube measured?—By the use of a circuit such as that shown in Fig. 11-87. Normal grid and plate voltages are applied as indicated. The signal voltage is adjusted at the control grid of the tube. An ac milliammeter, calibrated in transconductance, is connected in the plate circuit.

$$g_m = I_{pac} \times 1000.$$

11.88 Show a circuit for measuring output power.—A circuit suitable for the measurement of power output is shown in Fig. 11-88. The power equals:

$$I_p^2 \times R_L$$

where,

I_p is the plate current in ac units,
R_L is the load impedance in ohms.

11.89 Show a circuit suitable for measuring amplification factor.—A simple test for measuring the ac amplification factor is shown in Fig. 11-89. The ratio of the output voltage to the input voltage is a measure of the amplification factor. This test is used for both triodes and pentodes.

11.90 Show a circuit suitable for measuring only cathode-emission current.—The basic circuit for the measurement of cathode emission is shown in Fig. 11-90. This is the method used in emission-type tube testers. The emission current is measured with all the elements except the cathode and heater tied together.

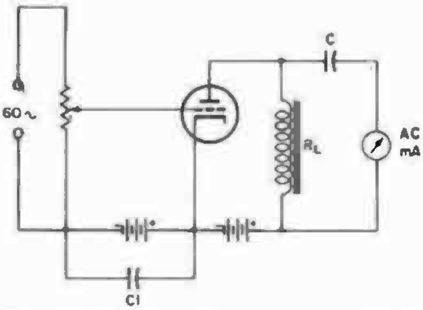


Fig. 11-88. Circuit for measurement of power output.

11.91 Describe a cathode-ray tube.—It is a large vacuum tube used in cathode-ray oscilloscopes, and in television receivers for both black and white and color reception. The cathode-ray tube was developed to its present state of perfection by many different researchers. In 1897 Karl F. Braun perfected the Crookes tube invented by Sir William Crookes for the study of cathode rays. Later, deflection plates were added by J. Thompson, an English physicist, and in 1899 E. Vichart, a German researcher, concentrated the electron beam with coil windings external to the axis of the tube. In 1902 A. A. Petrovski, a Russian, suggested the use of two external coils at right angles to each other to be used for deflecting the beam in the vertical and horizontal planes. Later Boris L. Rosing, also a Russian, experimented with the tube for the transmission of images and letters. Many other researchers have shared in the development of the cath-

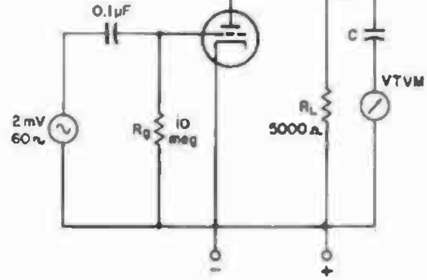


Fig. 11-89. The EIA method of measuring the ac amplification factor of triodes and pentodes. The internal impedance of the signal source should not exceed 2500 ohms.

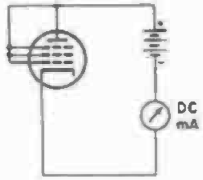


Fig. 11-90. Circuit for measurement of cathode emission current.

ode-ray tube, notably V. K. Zworykiu of the Uuited States for his work in the invention of the iconoscope and iunage orthicon tubes, and Allen B DuMont in his work of developing commercial oscilloscopes.

The cathode-ray tube is the heart of a cathode-ray oscilloscope and is operated in conjunction with quite complicated circuitry for the deflection of the beam and amplification of the applied signal voltages (see Question 22.69.) The tube employs an assembly, termed an *electron gun*, for projecting the electron stream, and a phosphor screen at the other end for displaying the waveform image (Fig. 11-92). The gun consists of a thermionic cathode, with various accelerating electrodes for directing and focusiug on the display screen the emitted electrons from the gun. The resulting narrow beam of electrons strike the screen in the form of a small round spot, with enough energy to cause fluorescence of the phosphor screen.

Cathode-ray tubes may be considered to he extremely fast x-y plotters, which plot input voltage versus time. The stylus of this plotter is a luminous spot which traces the image. In the conventional oscilloscope, the voltage to be observed is applied to the y-axis (vertical) input which moves the spot up and down in accordance with the instantaneous values of the input voltage. The x-axis voltage is supplied from an internally generated ramp voltage, which moves the spot uniformly from left to right across the display area of the screen. The spot then traces an image which displays input-voltage variation (amplitude) as a function of time. When the y-axis is repeated at a fast rate, the display appears as a steady straight line. Oscilloscopes may also be used to display mechanical motion with the proper transducer.

11.92 How are the electrons in a cathode-ray tube caused to strike the fluorescent screen?—Referring to Fig. 11-92, it will be seen that the heart of a cathode-ray tube is its electron gun. The gun is mounted at the end of a glass envelope and projects its electron stream towards the opposite end of the envelope, which contains a fluorescent screen. The construction of the gun is not simple, but consists of a number of parts that must be accurately manufactured and aligned.

The electrons are emitted by a heated cathode A which surrounds a heater element B. The electrons in leaving the cathode pass through a control grid C. The control grid determines the number of electrons that can pass and, consequently, the intensity of the light emitted by the fluorescent screen L. After leaving the control grid, the electrons are focused into a pencillike beam by the focusing anode D. Their velocity is further increased by the accelerating anodes E and F. At this point the electron stream is caused to

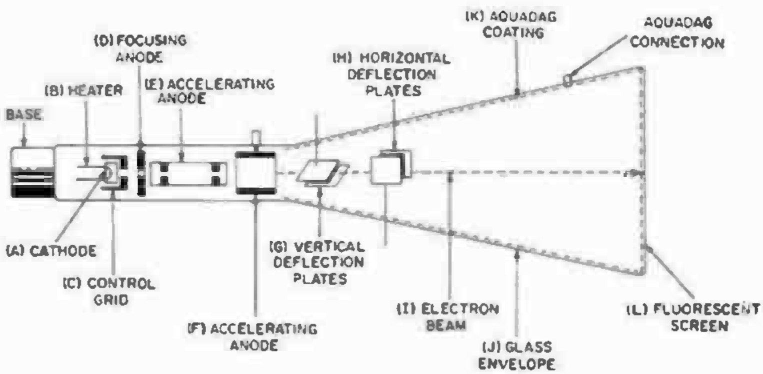


Fig. 11-92. Arrangement of the elements in an electrostatically deflected cathode-ray tube.

be deflected over various portions of the fluorescent screen by means of the horizontal and vertical deflection plates G and H.

Starting at the last anode F and continuing to the screen, the interior of the glass envelope is covered with a grounded graphite coating called aquadag. Its purpose is to bleed to ground stray electrons caused by secondary emission. (See Question 11.47.)

11.93 What is the method used to deflect the electron stream across the screen of a cathode-ray tube?—Two methods are used—the electrostatic and the electromagnetic. For oscilloscope work, the beam is generally deflected by the electrostatic method.

A front-end view of the deflecting plates, as would be seen looking through the fluorescent screen is shown in Fig. 11-93. The deflecting plates are arranged in pairs, two horizontal and two vertical. The electron beam passes through the center of the four plates as indicated by the spot in the center.

If a negative charge is applied to the left horizontal plate and a positive charge applied to the right horizontal plate, the beam will move to the right. When the charge is removed, the beam will return to the center again.

Removing the charge from the horizontal plates and charging the upper vertical plate negative and the lower vertical plate positive will move the beam downward. Reversing the polarities of the horizontal or vertical plates results in the beam moving in the opposite direction. Thus, if an alternating current is applied to the plates, the spot can be caused to oscillate back and forth either in a vertical or horizontal direction. The distance the spot is moved for either set of plates is directly proportional to the applied voltage. If equal voltages are applied to both the vertical and horizontal plates, the beam will be centered. With various combinations of voltage applied to the vertical and horizontal plates, the beam may be caused to move to any position on the screen. This is the basic principle on which a cathode-ray oscilloscope operates.

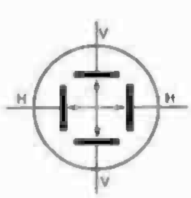


Fig. 11-93. The deflecting plates as seen when looking into the screen end of a cathode-ray tube.

11.94 How is the electron beam caused to draw out a pattern on an oscilloscope?—All oscilloscopes have an internal sawtooth oscillator for deflecting the beam in a linear manner across the screen. This motion is applied to the horizontal plates, while the signal is applied to the vertical plates. If the signal is an alternating one, and the sawtooth oscillator is adjusted for the fundamental frequency or a multiple of the applied waveform, one or more cycles of the waveform may be drawn out on the screen. This subject is further discussed in Question 22.69.

11.95 How is the sensitivity of a cathode-ray tube rated?—It is rated in terms of the voltage required to deflect the beam a given distance on the screen in millimeters or inches, with the signal applied directly to the deflecting plates (without amplification). Sensitivity is rated in rms volts per millimeter or peak-to-peak volts. For dc deflection, it is rated in a similar manner.

11.96 Are the characteristics of fluorescent screen the same for all types of cathode-ray tubes?—No. Tubes may be obtained with different degrees of persistence—long, medium, and short. Also, they may be obtained to reproduce the image in green, blue, or white.

11.97 What does the term "persistence" mean? — The various types of phosphor used in the manufacture of cathode-ray tubes (CRT) are many and varied. Basically there are at the present time 35 types, designated P1 to P35, although there are many more that are unregistered. The persistence of the cathode-ray tube screen is classified by the time it takes for the screen to fade to 10 percent of its original brightness.

The most common phosphors used for oscilloscopes are: P1, P2, P5, P31, and P35 with a grain size of 10 microns, with colors being yellowish-green, blue, green, and blue-white. Phosphors P8, P9 and P30 are no longer used. Luminescence occurs during the excitation of the screen. Luminescence that persists more than *ten nanoseconds* after the excitation has ceased is termed *phos-*

Phosphor	Trace Color		Low-Level Persistence	Writing Rate (aluminized 5BH at 10 kV)	Relative Burn Resistance	Relative Visual Brightness			Trace Contrast (without filter)
	Under Excitation	After-Glow				5AQ 2.5kV	5AM 5kV	5BH 10kV	
P1	green	green	180 ms	200 cm/μs	100	100	600	3000	good
P2	blue	yellow	1 s	600 cm/μs	150	90	550	2000	fair
P7	blu-wht	yellow	3 s	200 cm/μs	75	50	400	1500	poor
P11	blue	blue	20 ms	700 cm/μs	75	20	200	800	poor
P31	blu-grn	lt grn	0.5 s	400 cm/μs	250	110	1300	5000	fair to poor

Phosphor	Phosphor Application	Advantages	Disadvantages
P1	General purpose phosphor for visual observation of repetitive signals.	Good trace to background contrast; color near center of visual spectrum.	Lack of persistence; color too yellow for fast writing rate photography.
P2	General purpose phosphor for visual and photographic observation of slow and normal repetitive signals.	Persistence characteristic may be intensified with an amber filter; short persistence bluish trace very good for photography.	Low trace to background contrast; trace brightness lower than P1 and P31.
P7	For observation of low repetition rate and nonrepetitive phenomena.	Very long persistence; wide spectral separation between blue and yellow so that filters may emphasize either characteristic.	Sepcialized characteristics limit general use.
P11	For photographing all signal types, particularly those requiring fast writing rate.	Blue trace color near maximum sensitivity of film.	Specialized characteristics limit its general use.
P31	General purpose phosphor for visual and photographic observation of repetitive signals.	Provides highest visual brightness—particularly at 5 kV and above in aluminized tubes; listed characteristics combine to make it a good general purpose phosphor.	Low trace to background contrast under bright ambient light conditions.

Fig. 11-97. Characteristics of phosphors used for cathode-ray tube screens.

phorescence. The chemical composition of typical CRT screens is: zinc orthosilicate, zinc cadmium sulphide, zinc sulphide, and many other combinations.

Screens are available with variable persistence which will hold images from 0.1 second to several minutes, hours, and even days. Thus, multiple images may be recorded and held or wiped off with the push of a button. To eliminate parallax in photography and when viewing an image at an angle, CRT's may be obtained with a flat screen and an internal graticule. A table of CRT phosphor characteristics is given in Fig. 11-97 with their advantages and disadvantages.

11.98 How is electromagnetic deflection used with a cathode-ray tube?—It is an electronic system of deflecting the electron beam in the cathode-ray tube by means of electromagnets placed around the exterior of the neck near the front end of the electron gun. The gun construction is the same as that used in the electrostatically deflected one, except that the focusing anode is replaced by an external focusing coil and the deflection plates are replaced by deflection coils.

11.99 What is aquadog?—A graphite coating used on the inside of a cathode-ray tube to collect secondary emission electrons and conduct them to ground.

11.100 What is an intensifier element?—An element in a cathode-ray tube, consisting of a band of graphite (aquadag) on the inner surface of the glass envelope, connected to a source of high voltage. Its purpose is to accelerate the electron beam after it has been deflected.

11.101 What is an ultor element in a cathode-ray tube?—It is the element which receives the highest dc voltage for acceleration of the electron beam prior to its deflection.

11.102 What is a post-ultor element in a cathode-ray tube?—It is the element to which is applied a dc voltage higher than the ultor element voltage to accelerate the electron beam after its deflection.

11.103 What is germanium?—A rare metal discovered by Winkler in Saxony, Germany in the year 1896. Germanium is a by-product of zinc mining. Germanium crystals are grown from germanium dioxide powder. Germanium in its purest state behaves much like an insulator because it has very few electrical charge carriers. (See Question 11.111.) The conductivity of germanium may be increased by the addition of an impurity in small amounts.

11.104 What is the purpose of inducing impurities in semiconductor material?—To control its conductivity characteristics. By introducing arsenic or antimony in germanium, its free or negative mobile charges are increased. The use of gallium, indium, or aluminum in germanium, increases the number of positive mobile charges (called holes). This characteristic is taken advantage of in the manufacture of transistors. The adding of an impurity to a semiconductor is termed doping. (See Question 11.114.)

11.105 When was the phenomenon of asymmetric conduction in solids first observed?—The first known observation was by Munck and Henry in 1835, and later in 1874 by Braun. In 1905, Col. Dunwoody invented the crystal detector, used in the detection of electromagnetic waves. It consisted of a bar of silicon carbide or carborundum held between two contacts. However, in 1903, Pickard filed a patent application for a crystal detector in which a fine wire was placed in contact with the silicon. This was the first mention of a silicon rectifier, and was the forerunner of the present day silicon rectifier. Later, other minerals such as galena (lead sulphide) were employed as detectors. In 1883, Edison observed the flow of current between a hot filament and an anode placed in an evacuated bulb (Edison effect). In 1903 Fleming made use of Edison's discovery and devised the first vacuum diode detector. In 1906, Dr. Lee DeForest invented the three-element vacuum tube. During World War II, intensive research was conducted to improve crystal detectors used for microwave radar equipment. As a result of this research, the original point-contact transistor was invented at the Bell Telephone Laboratories in 1948. Thus, it may be seen that semiconductor devices preceded the vacuum tube by many years.

11.106 Name different type devices classed as semiconductors.—A few of the more commonly known devices are:

Chronistor—A high-speed switching transistor.

Contact protector—A device consisting of one or more diodes, which are connected across the circuit to reduce the effects of surge currents. For ac circuits, two diodes are connected in series back-to-back. For dc circuits, a single diode is used. (See Question 24.67).

Epitaxial annular transistor—A transistor in which a high resistivity silicon is epitaxially grown on a low resistivity silicon substrate in a single crystal relationship. (See Question 11.118.)

Field-Effect Transistor (FET) — A single-junction majority-carrier device and similar in several respects to a vacuum tube. The FET has a high input impedance, is voltage controlled, and has a high output impedance although in the latter respect, by different manufacturing processes, it can have a lower output resistance than a vacuum tube.

Photofet—A light-sensitive field effect transistor fitted with a lens system and used similarly to a photocell. (See Questions 11.161 and 19.170.)

Phototransistor—Similar to the above, except it is of the conventional transistor type, fitted with a lens system.

Photoconductive cell—Similar to the above.

Silicon controlled rectifier (SCR)—A three-element device used for changing alternating current to direct current. The third element is used to control the action of the rectifying elements. (See Question 11.150.)

Selenium transient-voltage suppressor—A selenium diode having a controlled reverse-breakdown characteristic. Connected across rectifier circuits to protect the rectifying devices from heavy transient currents. (See Question 11.158.)

Stabistor—A single- or multiple-pellet diode with controlled forward voltage characteristics, and always used in a forward biased condition. Stabistors are used in computer circuits and in low-voltage regulating circuits.

Thyrector—A protector similar to a zener diode. (See Question 11.148.)

Lumistor—A four-terminal high-temperatnre semiconductor

Thermistor—A thermally sensitive resistor that exhibits a change in electrical resistance with a change in its body temperature.

Raysistor—A device in which a source of light and photosensitive resistive elements are enclosed in a single light-tight case, and used for circuit control by controlling the illumination of light falling on the photosensitive surface.

Voltage-sensitive capacitor—A voltage-controlled diode that functions as a variable capacitor.

Varactor diode—Similar to the above.

Tunnel diode—So named from the nature of its characteristics, a process wherein a particle can disappear from one side of a barrier, and reappear on the other side instantaneously, as if the particle had tunneled under the barrier element.

Transistor—A device that may be used as an amplifier, oscillator, modulater, and many functions performed by vacuum tubes.

Zener diode—A semiconductor device, the equivalent of the gaseous voltage regulator tube. They may be obtained to operate over a wide range of voltages.

Thyrite varistor—A nonlinear resistor in which the current varies as the power of the applied voltage, and used as a surge protector.

The symbols used with semiconductors, and their construction, are shown in Fig. 11-114J.

11.107 Describe the basic structure of a transistor—The transistor is a semiconductor device, invented by Dr. Wm. Shockley, Dr. John Bardeen, and Dr. Walter H. Brattain of the Bell Telephone Laboratories in 1948. The name "Transistor" is coined from two words, transfer and resistor. The first transistor consisted of a particle of semiconductor material, such as germanium, mounted in a holder with two point contacts (Fig. 11-107A).

In this type structure, connections to the emitter and collector were made by means of leads with sharp points in contact with the surface of the crystal (much like the crystal detector used in the early days of wireless telegraphy). During the manufacturing process, two

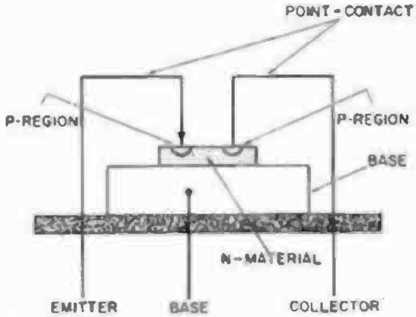


Fig. 11-107A. Cross-sectional view of point-contact transistor. This type of transistor was the original design, but has now been replaced by the junction-type transistor.

small areas of p-type material are produced. Therefore, the contact points of the leads are actually in contact with a p-type material. The n-material is between the p-material and the base. Point-contact construction is now obsolete, having been superseded by the junction-type construction.

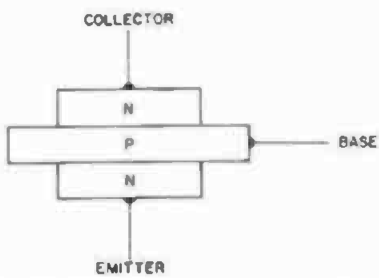


Fig. 11-107B. Basic construction for an npn transistor.

Junction transistors may be constructed using one of two methods, as shown in Figs. 11-107B and C. In the junction-type construction, the connecting leads are fused to the surface of the semiconductor material (Fig. 11-107D). The characteristics are control-

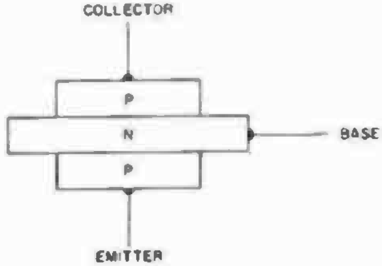


Fig. 11-107C. Basic construction for a pnp transistor.

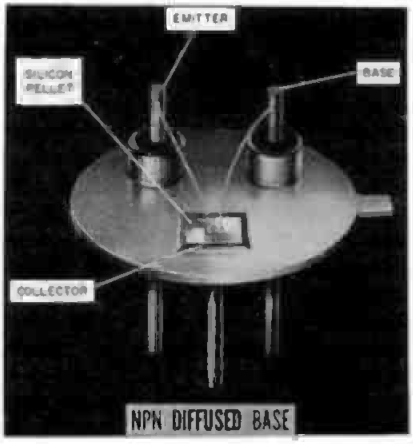


Fig. 11-107D. Interior view of General Electric diffused-base transistor.

led by doping the semiconductor material with impurities.

Since most semiconductor materials are photosensitive, it is necessary that the envelope enclosing the structure be opaque. Transistors do not require a vacuum for their operation. The characteristics of transistors are treated elsewhere in this section.

11.108 Does radiation affect the operation of transistors?—Yes, however the radiation caused by cathode-ray tubes or high-voltage rectifiers has little or no effect if the distance is greater than 12 inches. Radiation causes the semiconductor material to become *conductive*. Silicon devices are less affected than germanium. Cosmic, gamma, and nuclear radiations increase the leakage current and permanently affect the characteristics.

11.109 What is the difference between electron and conventional current flow?—In the field of electronics, two methods of dealing with current flow prevail. In the conventional school of thought, the current flows from *positive* to *negative*, while in the electronics

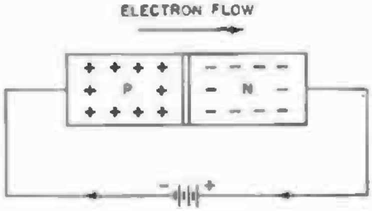


Fig. 11-109A. Direction of current flow in a reversed-biased junction diode.

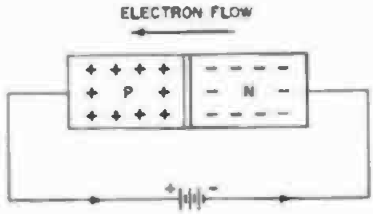


Fig. 11-109B. Direction of current flow in a forward-biased junction diode.

school of thought, the current flows with the electronic drift or *negative* to *positive*. Since the development of semiconductor devices, yet another theory of current flow has been developed—the hole movement positive to negative. In addition to this term are others such as forward current, reverse current, leakage current, saturation current and many others. Semiconductors with their opposite natures of npn and pnp combinations, and many more of multiple construction add to this confusion.

A better understanding of semiconductor current principles may be had by the diode concept. With an external battery connected across a pn junction (Fig. 11-109A), the current flow is determined by the polarity of the applied voltage and its effect on the space-charge region. In the illustration, the positive terminal of the battery is connected to the n material and the negative terminal to the p material. In this arrangement, the free electrons in the n material are attracted to the positive terminal of the battery and *away* from the junction. At the same time, holes from the p-type material are attracted *toward* the negative terminal of the battery and *away* from the junction. As a result, the space-charge region at the junction becomes effectively wider, and the potentional gradient increases until it approaches the battery potentional. Under these conditions the current flow is extremely small because

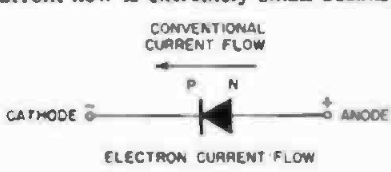


Fig. 11-109C. Symbol for a diode rectifier. The plus and minus terminals may be identified by arrow and bar.

no voltage difference exists across either the p-type or the n-type region and the pn junction is said to be *reverse-biased*.

If the connections are reversed as in Fig. 11-109B, the electrons in the p-type material near the positive terminal of the battery break their bonds and enter the battery, creating new holes. At the same time, electrons from the negative terminal of the battery enter the n-type material and diffuse toward the junction. The space-charge region then becomes effectively narrower, and the energy barrier decreases to an insignificant value. Excess electrons from the n-material can then penetrate the space-charge region, flow across the junction and move by way of the holes in the p-type material toward the positive battery terminal. This flow will continue as long as the external potential is applied. Under these conditions it is said the junction is *forward-biased* and maximum current will flow. In the construction of a diode (Fig. 11-109C), the anode is always positive and it consists of p-type material, and the cathode consists of n-type material. Thus, if a positive potential is applied to the anode (arrow symbol) conventional current flow is from the anode to the cathode, through the external circuit and back to the anode. Electron flow is in the opposite direction. (See Question 11.148.)

11.110 What is silicon?—A nonmetallic element used in the manufacture of diode rectifiers and transistors. Its resistivity is considerably higher than that of germanium.

11.111 Where does the resistance of germanium and silicon fall on the resistance scale of semiconductor material?—The relative position of pure germanium and silicon is given in Fig. 11-111. The scale indicates the resistance of conductors, semiconductors, and insulators per cubic centimeter. Pure germanium has a resistance of approximately 60 ohms per cubic centimeter. Germanium has a higher conductivity or less resistance to current flow than silicon, and is used in the majority of low- and medium-power diodes and transistors.

11.112 Define the terms, majority and minority carriers.—A majority carrier is the type of charge carrier that constitutes more than half of the total

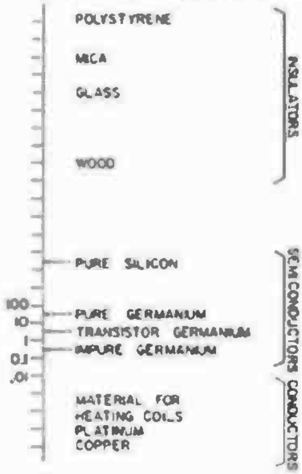


Fig. 11-111. Resistance of various materials per cubic centimeter. Pure germanium has about 60-ohms resistance per cubic centimeter.

charge-carrier concentration. Minority carriers constitute less than half the charge carriers.

11.113 Define the term junction as applied to semiconductors.—The term junction refers to the region between semiconductor regions of different properties, between metal and semiconductor material. An alloy of fused junction is formed by recrystallization on a base crystal from a liquid phase containing doping materials. A diffused junction is formed by a solid or vapor diffusion of an impurity within the semiconductor. Doped junctions are produced by the addition of an impurity that melts during the crystal growth. A grown junction is a junction produced during the growth of a crystal from a smelt. (See Question 11.118.)

11.114 What is the basic theory of a semiconductor?—To understand the theory of operation of a semiconductor device, it must be approached from the standpoint of its atomic structure.

The outer orbit of a germanium atom contains four electrons. The atomic structure for a pure germanium crystal is shown in Fig. 11-114A. Each atom containing four electrons forms covalent bonds with adjacent atoms. Therefore, there are no "free" electrous, and germanium in its pure state is a poor conductor of electricity. (See Fig. 11-111.) If a piece of "pure" germanium (the size used in a transistor) has a voltage applied to it, only a few microamperes of current will flow in the circuit. This current is caused by electrons which have been broken away from their bonds by thermal agitation, and will increase at an exponential rate with an increase of temperature.

Introducing into the germanium crystal an atom with five electrons, such as antimony or arsenic, changes the atomic structure to that of Fig. 11-114B. The extra electrons (called free electrons) will move toward the positive terminal of the external voltage source.

When an electron flows from the germanium crystal to the positive ter-

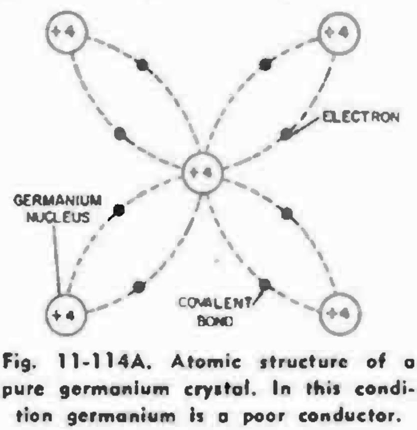


Fig. 11-114A. Atomic structure of a pure germanium crystal. In this condition germanium is a poor conductor.

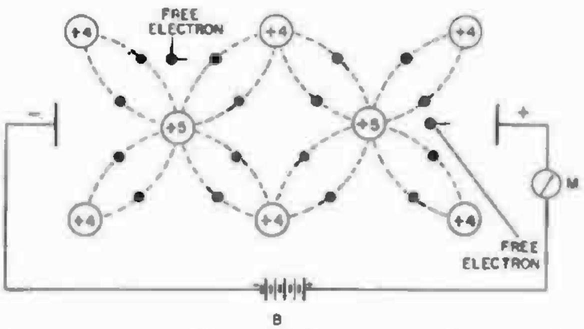


Fig. 11-114B. Atomic structure of a germanium crystal when a doping agent containing five electrons is induced. This type crystal is classified as n-type germanium.

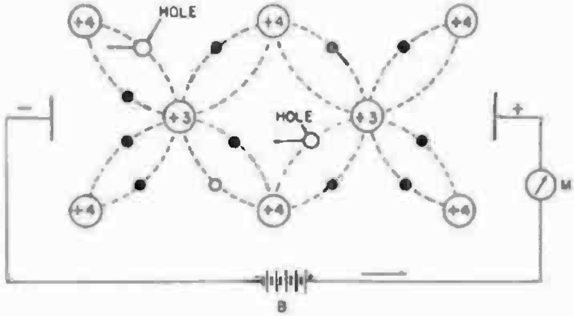


Fig. 11-114C. Atomic structure of a germanium crystal when a doping agent containing three electrons is induced. This type crystal is classified as p-type germanium.

minal of the external voltage source, another electron enters the crystal from the negative terminal of the voltage source. Thus, a continuous stream of electrons will flow as long as the external potential is maintained.

The atom containing the five electrons is called a "doping agent" or "donor." Such germanium crystals are classified as n-type germanium.

Using a doping agent of indium, gallium, or aluminum, which contain only three electrons in their outer orbit causes the germanium crystal to take the atomic structure of Fig. 11-114C. In this structure, there is a "hole" or "acceptor." The term hole is used to denote a mobile particle having a positive charge, and simulates the properties of an electron having a positive charge.

The use of the term hole is relatively new, but it is as useful as the term electron to explain the manner in which electricity is conducted by electrons. Electrons in semiconductors are also referred to as "carriers."

Pure germanium is also referred to as "intrinsic germanium," which means that the germanium has an equal number of donors and acceptors.

When a germanium crystal containing holes is subjected to an electrical field, electrons jump into the holes, and the holes appear to move towards the negative terminal of the external voltage source.

When a hole arrives at the negative terminal, an electron is emitted by the terminal and the hole is canceled. Simultaneously, an electron from one of the covalent bonds flows into the positive terminal of the voltage source. This new hole moves toward the negative terminal causing a continuous flow of holes in the crystal.

Germanium crystals having a deficiency of electrons are classified p-type germanium. Insofar, as the external electrical circuits are concerned, there is no difference between electron and hole current flow. However, the method of connection to the two types of transistors differs.

With an electrical field of one volt per centimeter in germanium, an electron will move at a velocity of 3600 centimeters per second, whereas a hole will move at a velocity of 1700 centimeters per second.

When a germanium crystal is so doped that it abruptly changes from an n-type to a p-type, and a positive potential is applied to the p region, and

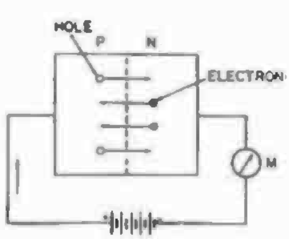


Fig. 11-114D. A germanium junction which consists of a p- and an n-type crystal.

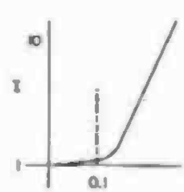


Fig. 11-114E. Voltage versus current characteristic of the junction shown in Fig. 11-114D.

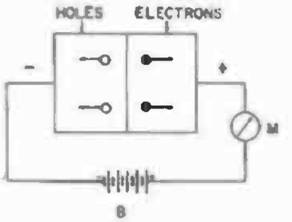


Fig. 11-114F. The junction transistor seen in Fig. 11-114D, with the battery polarities reversed.

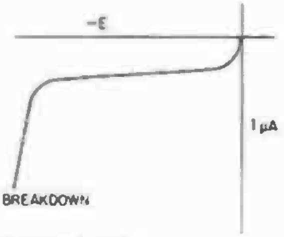


Fig. 11-114G. Voltage versus current characteristics of the junction transistor with the battery polarities shown in Fig. 11-114F.

a negative potential to the n region, the conditions will be as shown in Fig. 11-114D.

The holes move through the junction to the right and the electrons to the left, resulting in a voltage-current characteristic as shown in Fig. 11-114E. If the potential is reversed as shown in Fig. 11-114F both electrons and holes move away from the junction until the electrical field produced by their displacement counteracts the applied electrical field. Under these conditions zero current flows in the external circuit. Any minute amount of current which might flow is caused by thermal generated hole pairs. It will be noted in Fig. 11-114G, a plot of the voltage versus current for the reversed condition, the leakage current is essentially independent of the applied potential up to the point where the junction breaks down.

The basic construction of an npn-type transistor is shown in Fig. 11-114H. The left-hand n and center p section is forward-biased, and is termed the emitter junction. The other junction formed by the p and right-hand n section is reverse-biased, and is termed the collector junction. The p- type base is "lightly" doped in comparison to the n-type emitter; therefore, the majority of the current flowing from the emitter to the base is electron current with very little hole current. The major portion of the electrons emitted by the base diffuse across the collector junction and pass to the collector circuit. The ratio of emitter current to collector current is termed the "alpha" of a transistor.

It is essential that the value of alpha be as great as possible. This is achieved by a light doping of the base using a thin base region of the order of 1 mil, and minimizing the unwanted impurities in the germanium that might cause recombination of electrons before they transverse the base region. Alpha's of 0.95 to 0.99 are common in commercial transistors.

Only a small amount of leakage current flows in the collector circuit unless current is induced in the emitter. A small voltage of the order of 0.10 to 0.50 volt is all that is required to cause an appreciable amount of current flow; thus, the input power is quite small.

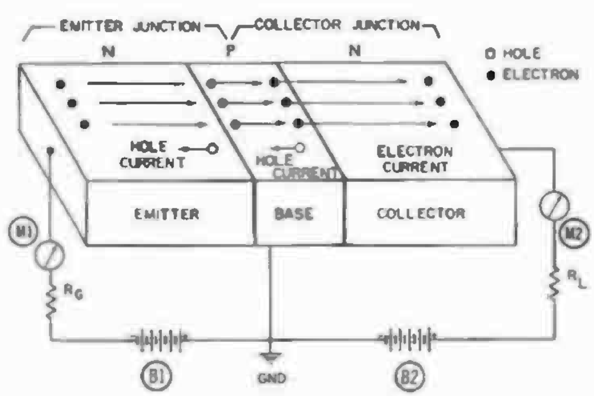


Fig. 11-114H. Basic construction and connections for an npn junction-type transistor.

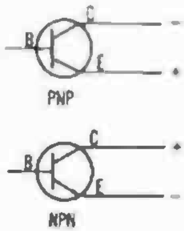


Fig. 11-114I. Symbols for a pnp and an npn transistor.

This permits a relatively large amount of power to be controlled in the external load circuit by a small amount of power at the input circuit.

The power gain of a transistor may be calculated:

$$P_o = \frac{P_{out}}{P_{in}}$$

$$P_{in} = \frac{E_{in}^2}{R_{in}}$$

$$P_{out} = \frac{E_{out}^2}{R_L}$$

where,

P_{out} is the output power,
P_{in} is the input power,
E_{in} is the voltage at the input,
E_{out} is the voltage at the output,
R_{in} is the input resistance of the transistor,
R_L is the load resistance.

The materials used in the manufacture of semiconductor devices are germanium and silicon. As germanium has a higher conductivity than silicon, it is used in most low- and medium-power diodes and transistors. Silicon is more suitable for high-power devices than is germanium since it may be used at much higher temperatures. Considerable research is being done in the development of gallium arsenide, which combines the most desirable characteristics of both germanium and silicon.

The npn-type junction transistor corresponds to the conventional triode vacuum tube using negative (electron) carriers. The pnp-type might be compared to a vacuum tube operating (if it were possible) with the electrons flowing from plate to cathode.

Graphical symbols for diodes and transistors are given in Figs. 11-114I and J. The arrows on the diagrams indicate the direction of electron flow. The base in a transistor corresponds to the control grid in a vacuum tube, the emitter to the cathode or filament, and the collector to the plate of a tube. However, transistors are entirely different from vacuum tubes as they are current-controlled devices, where a vacuum tube is voltage controlled and has a high input impedance. Transistors may have either a high- or low-input impedance, depending on the circuit design. The load impedance has a direct effect on the input impedance and vice versa. Although transistors will perform functions of the vacuum tube, their design requires a new concept in circuit design compared to that used for vacuum-tube circuit design. In Fig. 11-114J are given the symbols used for the more common types of transistors and diodes. However, this list is by no means complete, as the semiconductor field is undergoing continuous change.

11.115 What are the letter symbols used with semiconductor devices for identifying their electrical characteristics?—Although the letter symbols for transistors and diodes have not been completely standardized throughout the electronic industry, many symbols have become fairly well established by common usage. A tabulation of the symbols in use at the present time is given in Fig. 11-115. For letter symbols not given, the basic symbol is used and modified to suit the particular requirement.

Nomenclature used for the classification of semiconductor devices use the prefix 1N for diodes followed by the type number, such as 1N3493. Transistors use the prefix 2N followed by the type number. Letters appearing after the type number indicate a modification of the original design.

11.116 How are the polarities of semiconductor devices identified?—For diodes, the symbol representing an arrow and base is imprinted on the housing if large enough. The arrow always indicates the direction of *conventional current flow*, and is the *positive* (+) terminal. Electron flow is in the direction *opposite* to the arrow. If the device is too small for the symbol to read, a red dot is sometimes used to denote the positive terminal. For miniature-type diodes, the type number is indicated using the standard electronic industry color code. In this case the negative or cathode terminal is at the start of the nnmber, as shown in Fig. 11-116A. Diodes of early manufacture used a black ring or bar to denote the

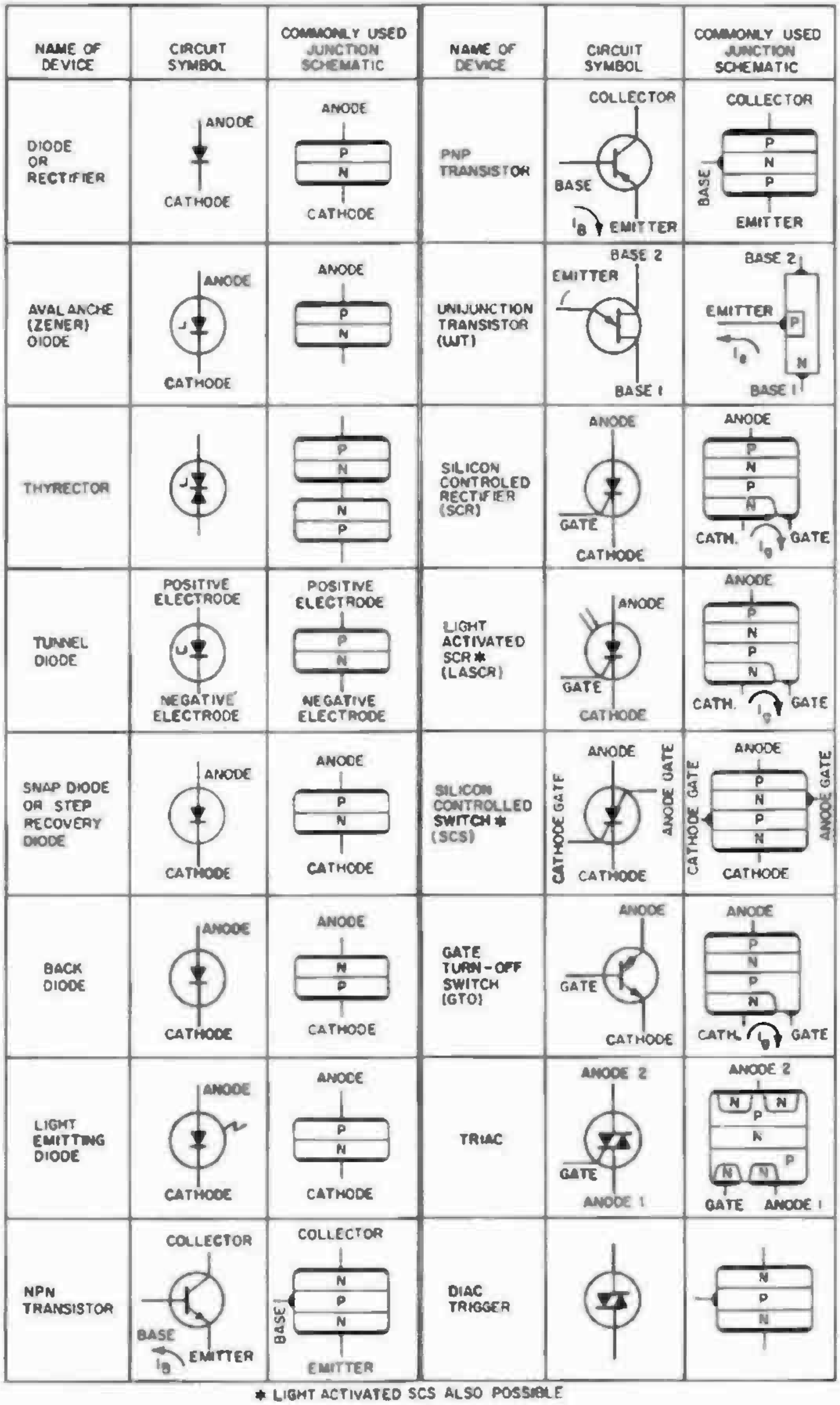


Fig. 11-114J. Graphical symbols for semiconductor devices, with their internal construction. The above symbols cover only the major designs. (Courtesy, General Electric Co., Semiconductor Div.)

GENERAL SEMICONDUCTOR SYMOLS

df	duty factor
η	efficiency (eta)
NF	noise figure
T	temperature
T_A	ambient temperature
T_C	case temperature
T_J	junction temperature
T_{MF}	mounting-flange temperature
T_{stg}	storage temperature
θ	thermal resistance
θ_{J-A}	thermal resistance, junction-to-ambient
θ_{J-C}	thermal resistance, junction-to-case
θ_{J-MF}	thermal resistance, junction-to-mounting flange
t_d	delay time
$t_d + t_r$	turn-on time
t_f	fall time
t_p	pulse time
t_r	rise time
t_s	storage time
$t_s + t_f$	turn-off time
τ	time constant (tau)
τ_s	saturation stored-charge time constant

TRANSISTOR SYMBOLS

$C_{b'c}$	collector-to-base feedback capacitance
C_c	collector - to - case capacitance
C_{cb}	collector-to-base feedback capacitauce
C_{ibo}	input capacitance, open circuit (common base)
C_{ieo}	input capacitance, open circuit (common emitter)
C_{obo}	output capacitance, open circuit (common base)
C_{oeo}	output capacitance, open circuit (common emitter)
$E_{s/b}$	second-breakdown energy
f_c	cutoff frequency
f_{hfb}	small-signal forward-current transfer-ratio cutoff frequency, short - circuit (common base)
f_{hfe}	small-signal forward-current transfer-ratio cutoff frequency, short - circuit (common emitter)
f_T	gain - bandwidth product (frequency at which small-signal forward-current transfer ratio, common emitter, extrapolates to unity)

TRANSISTOR SYMBOLS (Continued)

g_{me}	small-signal transconductance (common emitter)
G_{PB}	large-signal average power gain (common base)
G_{pb}	small-signal average power gain (common base)
G_{PE}	large-signal average power gain (common emitter)
G_{pe}	small-signal average power gain (common emitter)
h_{FB}	static forward-current transfer ratio (common base)
h_{fb}	small-signal forward-current transfer ratio, short circuit (common base)
h_{FE}	static forward-current transfer ratio (common emitter)
h_{fe}	small-signal forward-current transfer ratio, short circuit (common emitter)
h_{ib}	small-signal input impedance, short circuit (common base)
h_{IE}	static input resistance (common emitter)
h_{ie}	small-signal input impedance, short circuit (common emitter)
h_{ob}	small-signal output impedance, open circuit (common base)
h_{oe}	small-signal output impedance, open circuit (common emitter)
h_{rb}	small-signal reverse-voltage transfer ratio, open circuit (common base)
h_{re}	small-signal reverse-voltage transfer ratio, open circuit (common emitter)
I_B	base current
I_{B1}	turn-on current
I_{B2}	turn-off current
I_C	collector current
i_C	collector current, instantaneous value
I_{CB}	collector-cutoff current
I_{CBO}	collector - cutoff current, emitter open

Fig. 11-115. Letter symbols used with semiconductor devices.

TRANSISTOR SYMBOLS (Continued)

I_{CEO}	collector-cutoff current, base open
I_{CER}	collector-cutoff current, specified resistance between base and emitter
I_{CES}	collector-cutoff current, base short-circuited to emitter
I_{CEV}	collector-cutoff current, specified voltage between base and emitter
I_{CEX}	collector-cutoff current, specified circuit between base and emitter
I_{CM}	switching current (at minimum h_{FE} per specification)
I_E	emitter current
I_{EBO}	emitter-cutoff current, collector open
$I_{S/b}$	second-breakdown collector current
MAG	maximum available amplifier gain
MAG_c	maximum available conversion gain
MUG	maximum usable amplifier gain
P_{BE}	total dc or average power input to base (common emitter)
p_{BE}	total instantaneous power input to base (common emitter)
P_{CB}	total dc or average power input to collector (common base)
p_{CB}	total instantaneous power input to collector (common base)
P_{CE}	total dc or average power input to collector (common emitter)
p_{CE}	total instantaneous power input to collector (common emitter)
P	total dc or average power input to emitter (common base)
p_{EB}	total instantaneous power input to emitter (common base)
P_{IB}	large-signal input power (common base)
P_{ib}	small-signal input power (common base)
P_{IE}	large-signal input power (common emitter)
P_{ie}	small-signal input power (common emitter)
P_{OB}	large-signal output power (common base)
P_{ob}	small-signal output power (common base)
P_{OE}	large-signal output power (common emitter)
P_{oe}	small-signal output power (common emitter)
Q_s	stored base charge
r_{CE}(sat)	collector-to-emitter saturation resistance
$Re(h_{ie})$	real part of small-signal input impedance, short circuit (common emitter)
R_G	generator resistance
R_{ie}	input resistance (common emitter)
R_L	load resistance
R_{oe}	output resistance (common emitter)
R_s	source resistance
V_{BB}	base-supply voltage
V_{BC}	base-to-collector voltage
V_{BE}	base-to-emitter voltage
$V_{(BR)CBO}$	collector-to-base breakdown voltage, emitter open
$V_{(BR)CEO}$	collector-to-emitter breakdown voltage, base open
$V_{(BR)CER}$	collector-to-emitter breakdown voltage, specified resistance between base and emitter
$V_{(BR)CES}$	collector-to-emitter breakdown voltage, base short-circuited to emitter
$V_{(BR)CEV}$	collector-to-emitter breakdown voltage, specified voltage between base and emitter
$V_{(BR)EBO}$	emitter-to-base breakdown voltage, collector open
V_{CB}	collector-to-base voltage
V_{CB}(fl)	dc open-circuit voltage between collector and base (floating potential), emitter biased with respect to base
V_{CE}(fl)	dc open-circuit voltage between collector and emitter (floating potential), base biased with respect to emitter

(Courtesy, Radio Corporation of America)

TRANSISTOR SYMBOLS (Continued)

V_{CBO}	collector-to-base voltage (emitter open)
V_{CBV}	collector-to-base voltage, specified voltage between emitter and base
V_{CC}	collector-supply voltage
V_{CE}	collector-to-emitter voltage
V_{CEO}	collector-to-emitter voltage, base open
V_{CER}	collector-to-emitter voltage, specified resistance between base and emitter
V_{CES}	collector-to-emitter voltage, base short-circuited to emitter
V_{CEV}	collector-to-emitter voltage, specified voltage between base and emitter
V_{CE}(sat)	collector-to-emitter saturation voltage
V_{EB}	emitter-to-base voltage,
V_{EB}(fl)	dc open-circuit voltage between emitter and base (floating potential), collector biased with respect to base
V_{EBO}	emitter-to-base voltage, collector open
V_{EE}	emitter-supply voltage
V_{RT}	reach-through voltage
Y_{fe}	forward transconductance
Y_{ie}	input admittance
Y_{oe}	output admittance
Y_{re}	reverse transconductance

MOS FIELD-EFFECT TRANSISTOR SYMBOLS

A	voltage amplification ($= Y_{fs}/Y_{os} + Y_L$)
B_{os}	$= C_{ds}$
c_o	intrinsic channel capacitance
c_{ds}	drain-to-source capacitance (includes approximately 1-pF drain-to-case and interlead capacitance)

MOS FIELD-EFFECT TRANSISTOR SYMBOLS (Continued)

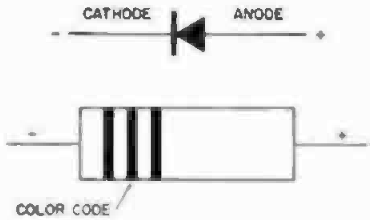
c_{gd}	gate-to-drain capacitance (includes 0.1-pF interlead capacitance)
c_{gs}	gate-to-source interlead and case capacitance
C_{iss}	small-signal input capacitance, short circuit
C_{rss}	small-signal reverse transfer capacitance, short circuit
g_{fs}	forward transconductance
g_{is}	input conductance
g_{os}	output conductance
I_D	dc drain current
I_{DS}(OFF)	drain-to-source OFF current
I_{GSS}	gate leakage current
NF	spot noise figure (generator resistance $R_G = 1$ megohm)
r_g	effective gate series resistance
r_d	active channel resistance
r_d'	unmodulated channel resistance
r_{DS}(ON)	drain-to-source ON resistance
r_{gd}	gate-to-drain leakage resistance
r_{gs}	gate-to-source leakage resistance
V_{DB}	drain-to-substrate voltage
V_{DS}	drain-to-source voltage
V_{GB}	dc gate-to-substrate voltage
v_{GB}	peak gate-to-subtrate voltage
V_{GS}	dc gate-to-source voltage
v_{GS}	peak gate-to-source voltage
V_{GS}(OFF)	gate-to-source cutoff voltage
Y_{fs}	forward transadmittance $\approx g_{fs}$
Y_{os}	output admittance $= g_{os} + jB_{os}$, $B_{os} = \omega C_{ds}$
Y_L	load admittance $= g_L + jB_L$

Fig. 11-115. (Continued)

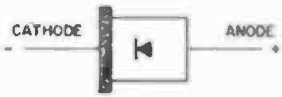
negative terminal. Top-hat types are marked as shown in Fig. 11-116B.

Transistors, due to their many types and shapes are sometimes hard to identify; therefore, it is good practice to consult the manufacturer's data sheet for the correct terminals and polarity. For three lead types, the center lead is usually the base. A dot indicates the collector; therefore, the remaining lead is the emitter.

11.117 What is the approximate resistance that may be expected between the elements of a transistor?—The ap-



(a) Tubular type with color-code stripes indicating the type number.



(b) Top-hat type.

Fig. 11-116. Diode polarity markings. The arrow indicates *conventional* current flow. *Electron* flow is in the opposite direction.

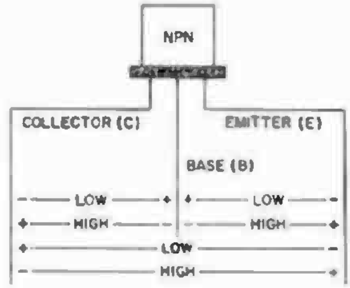


Fig. 11-117A. Approximate resistance between the elements of a transistor when measured with an ohmmeter. The polarities will be reversed for a pnp transistor.

proximate resistance that may be expected between any two elements of an npn transistor known to be operative is shown in Fig. 11-117A, and may be measured using a low current volt-ohmmeter. Before using the ohmmeter, the current flow between the leads of the ohmmeter must be measured with a microammeter or milliammeter; also measure the open-circuit voltage using a 1000 ohm-per-volt meter. For a typical 20,000 ohm-per-volt service meter on the $\times$ 100 resistance scale, the current measured 76 microamperes; on the $\times$ 1.0 scale, 800-microamperes; and 70 milliamperes on the 1/100 scale. The highest scale measured 16 volts, and the two lower scales 1.6 volts. Thus, it can be seen that a *good transistor could be damaged if measured on the low-resistance scale.*

Referring again to Fig. 11-117A, it is indicated that for a given polarity of the ohmmeter, either a high or low resistance may be expected between a given pair of terminals. Generally the ratio of high to low will be on the order of 200,000 ohms or greater; on the low side between 1 to 2 ohms. For a pnp type, the polarities are reversed from those shown. The effects of an internal short circuit for an operating transistor is shown in Fig. 11-117B, with a tabulation of the effect on the operating voltages shown in Fig. 11-117C. The operating voltages should, for accuracy, be measured with a meter of extremely high resistance, or better yet, with a dc vacuum-tube voltmeter.

11.118 Describe the methods of construction used in the manufacture of

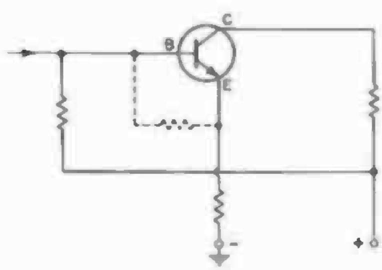


Fig. 11-117B. An npn-type transistor with its components. If two of the elements are shorted, as shown by the dotted line, the effects tabulated in Fig. 11-117C may exist.

Short Circuit Between Elements	*Base*	*Emitter*	*Collector*
B and E	increase	decrease	decrease
C and E	decrease	increase	increase
B and C	decrease	increase	increase

Fig. 11-117C. Tabulation of the effect on operating voltages across the resistive components of the transistor in Fig. 11-117B when certain elements of the transistor are short-circuited internally.

transistors.—The principal difference in the manufacturing of transistors is the method used in the construction of the junctions. Certain types of transistors employ up to eight elements, and there appears to be no limit to the various combinations possible. In Fig. 11-118A is shown the construction of a grown-junction transistor. An alloy-junction transistor is shown in Fig. 11-118B. During the manufacture of the material for a grown junction, the impurity content of the semiconductor is altered to provide npn or pnp regions. The grown material is cut into small sections and contacts are attached to the regions. In the alloy-junction type, small dots of n- or p-type impurity elements are attached to either side of a thin wafer of p- or n-type semiconductor material. After heating, the impurity dot alloys with the semiconductor material to form regions for the emitter and collector junctions. The base connection is made to the original semiconductor material. Drift-field transistors employ a modified alloy junction in which the impurity concentration in the wafer is diffused or graded, as shown in Fig. 11-118C. The drift field speeds up the current flow and extends the frequency response of the alloy-junction transistor.

A variation of the drift-field transistor is the microalloy diffused transistor (Fig. 11-118D). Very narrow base dimensions are achieved by etching techniques, resulting in a shortened current path to the collector. Mesa transistors, shown in Fig. 11-118E, use the original semiconductor material as the collector, with the base material diffused into the wafer and an emitter dot alloyed into the base region. A flat-topped peak or mesa is etched to reduce the area of the collector at the base junction. Mesa devices have large power-dissipation capabilities and can be operated at very high frequencies. Double-diffused epitaxial mesa transistors are grown by the use of vapor deposition to build up a crystal layer on a crystal wafer, and will permit the precise control of the physical and electrical dimensions independently of the nature of the original wafer. This technique is shown in Fig. 11-118F.

The planar transistor is a highly sophisticated method of constructing transistors. A limited area source is used for both the base diffusion and emitter diffusion, which provides a very small active area, with a large wire contact area. The advantage of the planar construction is its *high dissipation, lower leakage current, and lower collector-cutoff current,* which increases the stability and reliability. Planar construction is also used with several of the previously discussed base designs. A double-diffused epitaxial planar transistor is shown in Fig. 11-118G.

11.119 What are the general precautions to be observed with the use of semiconductor devices?—Limit the maximum junction temperature to prevent excessive leakage currents. Use low values of stability factors, such as using resistance in the emitter circuit. Reduce the value of the shunt resistance from base to emitter to a value as low as possible (taking into consideration the gain requirements). Use large values of collector current to minimize temperature changes. Use low values of source resistance to keep the stability factor low. In temperature-compensated circuits, provide a common heat sink for all semiconductor devices. *Do not* use an ohmmeter for checking a transistor or diode without first measuring the current flow between the ohmmeter leads. When soldering, use heat-sink pliers on the leads. Use a thermal derating factor when operating at temperatures above 25 degrees centigrade. The foregoing are but a few precautions to be observed in the use of semiconductor de-

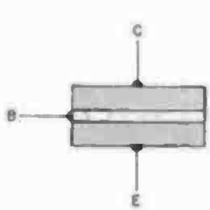


Fig. 11-118A. Grown-junction transistor.

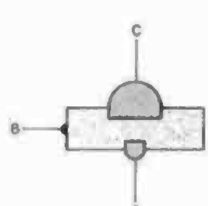


Fig. 11-118B. Alloy-junction transistor.

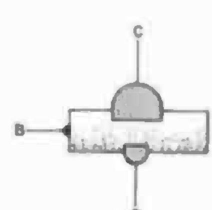


Fig. 11-118C. Drift-field transistor.

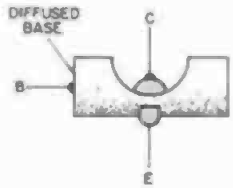


Fig. 11-118D. Microalloy-diffused transistor.

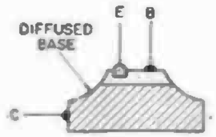


Fig. 11-118E. Mesa transistor.

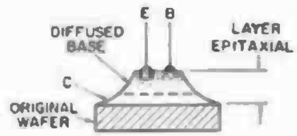
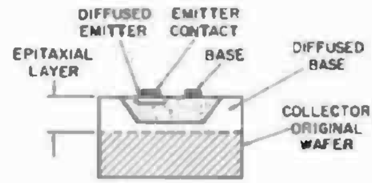


Fig. 11-118F. Epitaxial mesa transistor

Fig. 11-118G. Double-diffused epitaxial planar transistor.



vices. When in doubt, consult the manufacturer's data sheet.

11.120 What is meant by the term "punch-through"?—It is the widening of the space charge between the collector element and the base of a transistor. As the potential V_{cb} is increased from a low to a high value, the collector-base space charge is widened. This widening effect of the space charge, narrows the effective width of the base. If the diode space charge does not "avalanche" before the space charge spreads to the emitter section, a phenomenon termed "punch-through" is encountered. (See Fig. 11-120.)

The effect is: the base disappears as the collector-base, space charge layer contacts the emitter, creating a relatively low resistance between the emitter and the collector, causing a sharp rise in the current. The transistor action then ceases. As there is no voltage breakdown in the transistor, it will start functioning again if the voltage is lowered to a value below where punch-through occurs.

When a transistor is operated in the punch-through region, its functioning is not normal, and heat is generated internally which can cause permanent damage to the elements.

11.121 Define breakdown voltage in a transistor.—Breakdown voltage is that voltage value between two given elements in a transistor at which the crystal structure changes and current begins to increase rapidly. Breakdown voltage may be measured with the third electrode open, shorted, or biased in either the forward or reverse direction. A group of collector characteristics for different values of base bias are shown in Fig. 11-121. It will be observed the collector-to-emitter breakdown voltage increases as the base-to-emitter bias is decreased from the normal forward values through zero to reverse. As the resistance in the base-to-emitter circuit decreases, the collector characteristics develop two breakdown points. After the initial breakdown, the collector-to-emitter voltage decreases with an increasing collector current, until another breakdown occurs at the lower voltage.

Breakdown can be very destructive in power transistors. A breakdown mechanism, termed second breakdown, is an electrical and thermal process in which current is concentrated in a very small area. The high current, together with the voltage across the transistor, causes intense heating, melting a hole from the collector to the emitter, caus-

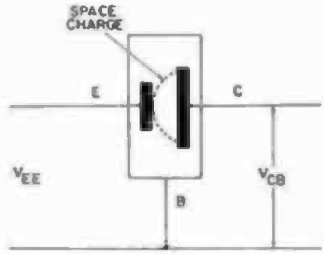


Fig. 11-120. Spreading of the space charge between the emitter and the collector. This phenomenon is called "punch-through."

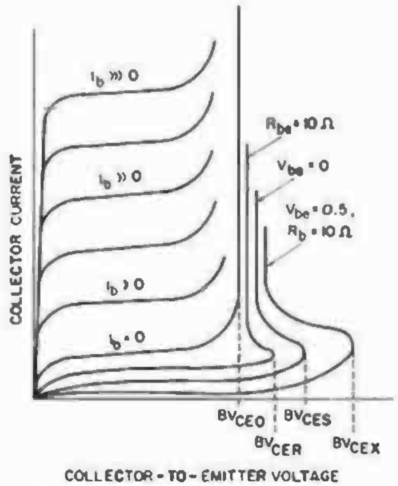


Fig. 11-121. Typical collector characteristic curves showing locations of various breakdown voltages. (Courtesy Radio Corporation of America)

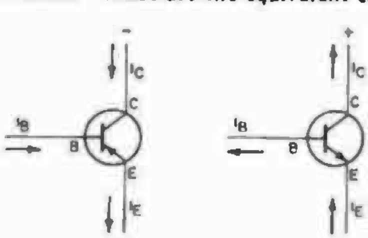
ing a short circuit and internal breakdown of the transistor.

The fundamental limitation to the use of transistors is the breakdown voltage, ($BV_{..r}$). Because the breakdown voltage is not sharp, it is necessary to specify the value of collector current at which breakdown will occur. This data may be obtained from the manufacturer's data sheet.

11.122 Define the term "extrinsic transconductance."—It is the quotient of a small change in collector current divided by a small change in emitter-to-base voltage producing it, under the condition that other voltages remain unchanged.

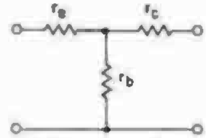
11.123 Show the flow of current through a transistor.—The current flow paths for both pnp- and npn-type transistors are given in Fig. 11-123. It will be observed that the flow of current is reversed from one type to the other.

11.124 What are the equivalent circuits for transistors?—Transistors may be considered to be a "T" configuration active network. Three such networks are given Fig. 11-124.

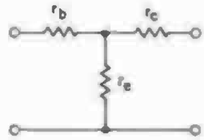


(a) In a pnp type. *(b) In an npn type.*

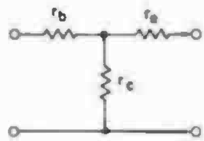
Fig. 11-123. Current flow paths as they occur in a transistor.



(a) Common base.

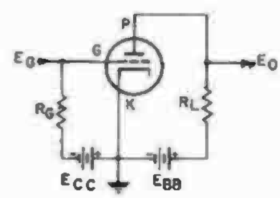


(b) Common emitter.

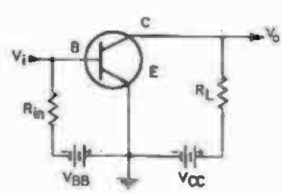


(c) Common collector.

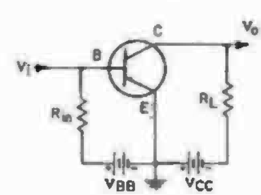
Fig. 11-124. Equivalent circuits for transistors.



(a) Vacuum tube.

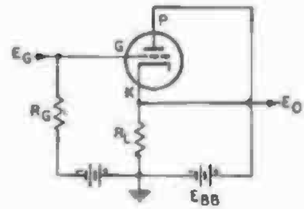


(b) Npn transistor.

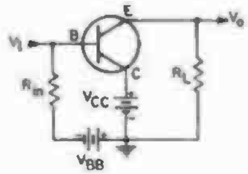


(c) Pnp transistor.

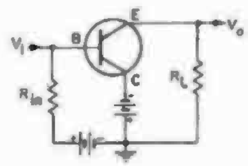
Fig. 11-125A. Comparison of a grounded cathode vacuum tube to a transistor using a grounded-emitter connection.



(a) *Vacuum tube.*

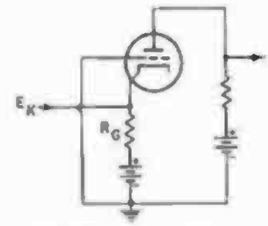


(b) *An npn transistor.*

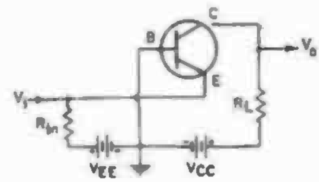


(c) *A pnp transistor.*

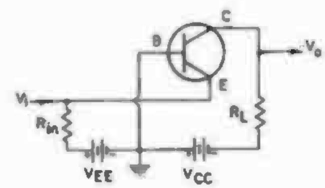
Fig. 11-125B. Comparison of a cathode-follower vacuum tube circuit (plate at ac ground) to a transistor using a grounded-collector connection.



(a) *Vacuum tube.*



(b) *An npn transistor.*



(c) *A pnp transistor.*

Fig. 11-125C. Comparison of a grounded-grid vacuum tube to a transistor using a grounded-base connection.

11.125 What are the phase relationships between the input and output of a transistor?—For the conventional transistor the *only* configuration that will provide a 180-degree phase reversal between input and output is the *grounded or common-emitter* configuration (Fig. 11-125A). Phase relations for the grounded or common-collector and common-base configurations are shown in Figs. 11-125B and C. These configurations have the same phase for both input and output. Equivalent vacuum tube circuits have been included for comparison. In the instance of the field-effect transistor (FET), Fig. 11-125D, for a given instantaneous polarity of the input signal at the gate, the output signal at the drain is reversed by 180 degrees. The signal at the source is of the same phase as the gate. This makes the FET ideal for phase-inverter applications.

11.126 What is the relationship between the input and output impedance of a transistor?—The input resistance for the common-collector and common-base configuration increases with an increase of the load resistance R_L. For the common-emitter, the input resistance decreases as the load resistance is increased; therefore, changes of input or output resistance are reflected from one to the other. Typical variations of output resistance for changes in generator impedance are given in Fig. 11-126. Voltage, power, and current gains are shown for a typical common-emitter configuration in Fig. 11-128.

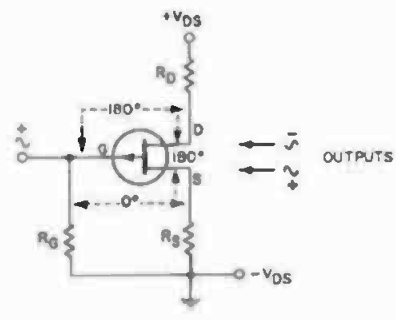
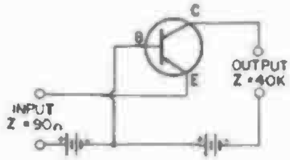
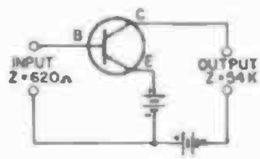


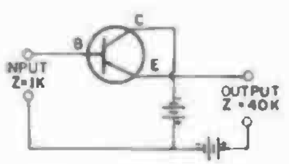
Fig. 11-125D. Signal-voltage polarities in a p-channel field-effect transistor (FET).



(a) Grounded base.



(b) Grounded emitter.



(c) Grounded collector.

Fig. 11-126. Junction-type transistor circuits showing the input and output impedances. The impedances shown are typical and will vary with different type transistors.

11.127 Show typical amplifier circuits for transistors.—Three simple amplifier circuits are shown in Figs. 11-127A, B, and C. In Fig. 11-127A is shown a two-stage resistance-capacitance (RC) coupled amplifier. In Fig. 11-127B, the circuit is direct-coupled, eliminating the coupling capacitors. Today's trends are toward the use of direct-coupled circuitry, since it reduces the phase-shift, extends the frequency range in both directions, and reduces the number of components. However, on the other hand, the gain is somewhat reduced over that of the RC coupled types, and the number of stages that can be direct-coupled is limited. Variations in the bias voltage are also reflected in the other stages. The term direct-coupled does not imply that it may be used for the amplification of direct current, as the extremely low-frequency signals are limited by factors other than the coupling method.

A transformer-coupled amplifier is shown in Fig. 11-127C. Although either pnp- or npn-type transistors may be used if the correct polarities are observed, it is quite important that the transformers match the transistor input and output impedances. Transistors are not necessarily confined to the circuits shown, but may be used in any amplifier configuration, such as single-ended types, push-pull, class-A, -B and -C, and other subscripts. Transistors may also be used for radio-frequency amplification up to several hundred megahertz. In fact, transistors may be used in almost any application where vacuum tubes are used, and in many circuits where it would be impractical to use tubes. Transistors may be used in an impedance-coupled circuit or with autotransformers. However, these circuits are generally only employed for special-purpose amplifiers.

Transistors may also be operated in parallel, similar to vacuum tubes, but unfortunately transistors, although of the same type and manufacture, do not always have similar characteristics. If the transistors are unmatched, when connected in parallel one transistor may be carrying most of the load. Two solutions are apparent—either use matched pairs of transistors or employ emitter resistors to balance the load. It has been found that for power transistors rated at 5 amperes or greater, a resistance of about 0.2 ohm connected in each emitter will generally be satisfactory. A rule of thumb is to use a resistor that will

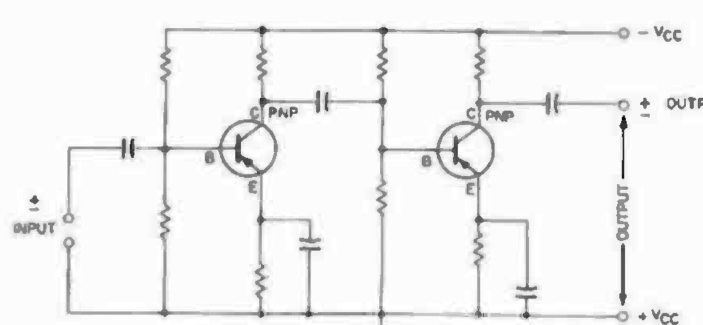


Fig. 11-127A. Typical resistance-capacitance coupled audio amplifier.

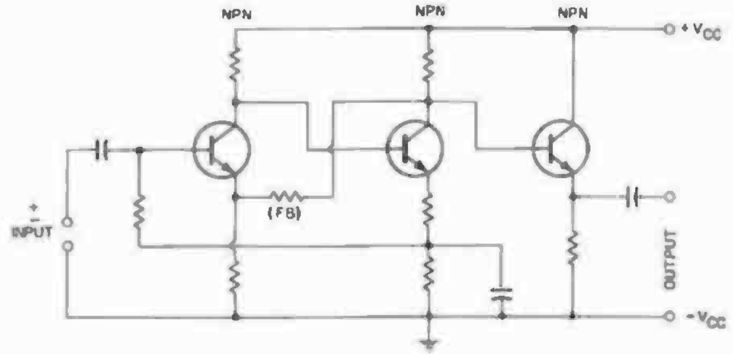


Fig. 11-127B. Direct-coupled amplifier using negative feedback between first and second stages.

cause a voltage drop of 1 volt at the maximum collector current. If possible, measure the characteristics of each transistor and then select two having nearly the same characteristics; also insert the resistor in the emitter circuits. If these precautions are not observed, one transistor can be badly damaged.

11.128 Show the voltage, power, and current gains for a typical transistor.—Typical curves for a conventional transistor using a common-emitter configuration are shown in Fig. 11-128. It will be noted the current gain decreases as the load resistance is *increased*. The voltage gain *increases* as the load resistance is *increased*. Maximum power gain occurs when the load resistance is approximately 40,000 ohms, and it may exceed unity. For the common-collector connection, the current gain decreases as the load resistance is increased, and while the voltage gain increases as the load resistance is increased, it never exceeds unity. Curves such as these help the designer to select a set of conditions for a specific result.

11.129 How is power gain for a transistor calculated?—The power gain varies as the ratio of the input to output impedance, and may be computed:

$$dB = 10 \operatorname{Log}_{10} \frac{Z_{out}}{Z_{in}}$$

where,

Z_{in} is the input impedance in ohms,
Z_{out} is the output impedance.

11.130 Can negative feedback be applied to transistor amplifiers?—Negative feedback is employed extensively in transistor amplifiers much in the same manner as vacuum tubes. One of the most common methods being used is to omit bypassing the emitter circuits. Referring to Fig. 11-130, here two feedback circuits are employed, one from the output to the emitter of Q1 and a second from the collector of Q2 to the emitter of Q1. In this manner a total of 40 dB of negative feedback is acquired —16 dB between Q1 and Q2, and 24 dB from the output of the emitter of Q1. If an output transformer is used, the feedback is taken from the output winding and returned to the proper phase point in the early amplifier stages. The subject of negative feedback and the amplifier circuit shown is further discussed in Question 12.255.

11.131 Define the terms alpha and beta as applied to transistors.—An im-

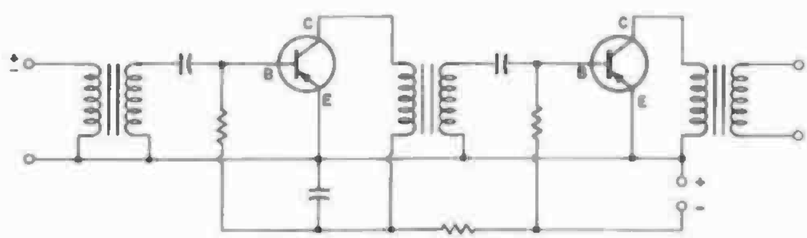


Fig. 11-127C. Typical transformer-coupled audio amplifier. The transformers are selected to match the impedance between stages.

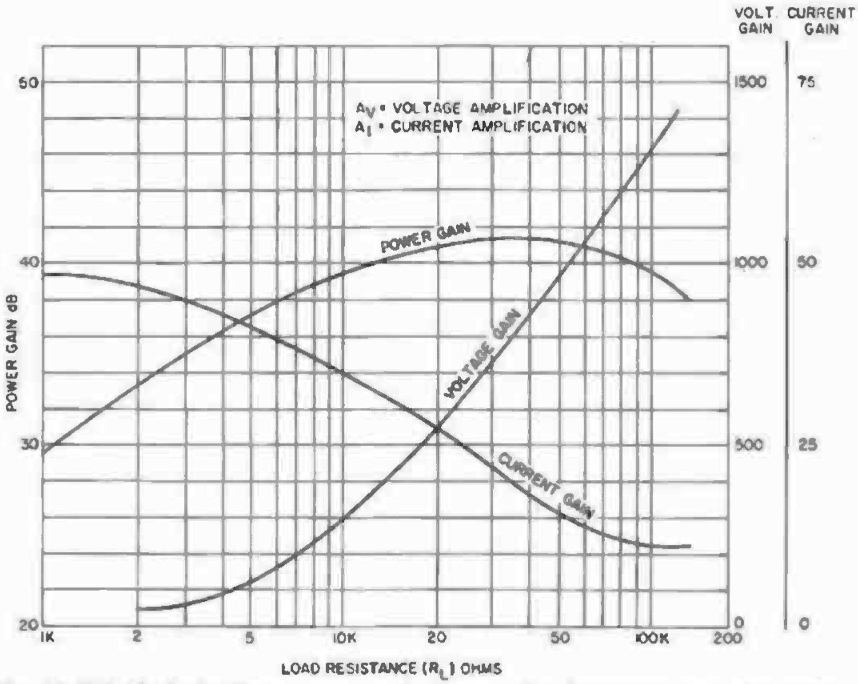


Fig. 11-128. Typical voltage, power, and current gains for a conventional transistor using a common-emitter configuration.

portant characteristic of a transistor is its forward current-transfer ratio, or the ratio of the current in the output to the current in the input element. Because of the many different configurations for connecting transistors, the forward transfer ratio is specified for a particular circuit configuration. The forward current-transfer ratio for the common-base configuration is often referred to as alpha (α) and the common-emitter forward current-transfer ratio as beta (β). In common-base circuitry, the emitter is the input element and the collector is the output element. Therefore, the dc alpha is the ratio of the dc collector current I_C to the dc emitter current I_E. For the common-emitter, the dc beta is then the ratio of the dc collector current I_C to the base current I_B.

Because the ratios are based on the dc currents, they are termed alpha and

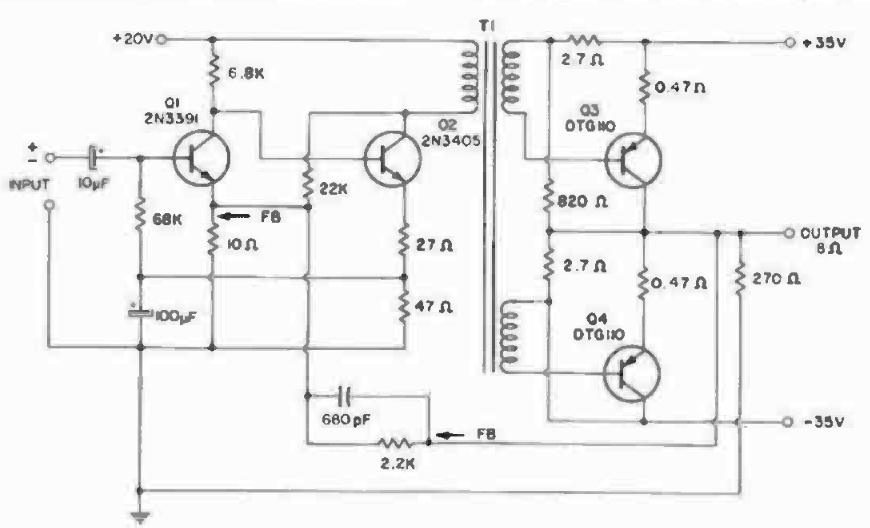


Fig. 11-130. A 50-watt push-pull amplifier. (Courtesy, Delco Radio Corp.)

beta. The ratios are also given in terms of the ratio of signal current, relative to the input and output, or as a ratio of change in the output current to the input current which causes the change.

The term alpha is also used to denote the frequency cutoff of a transistor, and is defined as the frequency at which the value of alpha for a common-base configuration, or beta for a common-emitter circuit, falls to 0.707 times its value at a frequency of 1000 Hz.

Gain bandwidth product is the frequency at which the common-emitter forward current-transfer ratio beta is equal to unity. It indicates the useful frequency range of the device and assists in the determination of the most suitable configuration for a given application.

11.132 What is cutoff current in a transistor?—When a transistor is biased to a nonconducting state, small reverse dc currents flow, consisting of leakage currents which are related to the surface characteristics of the semiconductor material and saturation currents. Saturation current increases with temperature and is related to the impurity concentration in the material. Collector-cutoff current is a dc current caused when the collector-to-base circuit is reverse-biased, and the emitter-to-base circuit is open. Emitter-cutoff current flows when the emitter-to-base is reverse-biased and the collector-to-base circuit is open.

11.133 Describe the various methods used for biasing transistors.—Several different methods of applying bias voltage to transistors are shown in Fig. 11-133A, with a master circuit for aiding in the selection of the proper circuit shown in Fig. 11-133B. Comparing the circuits shown in Fig. 11-133A their equivalents may be had by making the resistors in Fig. 11-133B equal to zero or infinity, for analyzation and study. As an example, the circuit of Fig. 11-133A, part (d) may be duplicated in Fig. 11-133B by shorting out resistors R4 and R5.

The circuit in part (g) of Fig. 11-133A employs a split voltage divider for R2. A capacitor connected at the junction of the two resistors shunts any ac feedback current to ground. The stability of circuits (a), (d), and (g) in Fig. 11-133A, may be poor unless the voltage drop across the load resistor is at least one-third the value of the power supply voltage V_{cc}. The final determining factors will be gain and stability. Stability may be enhanced by the use of a thermistor (see Question 11.157) to compensate for increases in collector current with increasing temperature. The resistance of the thermistor decreases as the temperature increases; thus, the bias voltage is decreased and the collector voltage tends to remain constant. Diode biasing may also be used for both temperature and voltage variations. The diode is used to

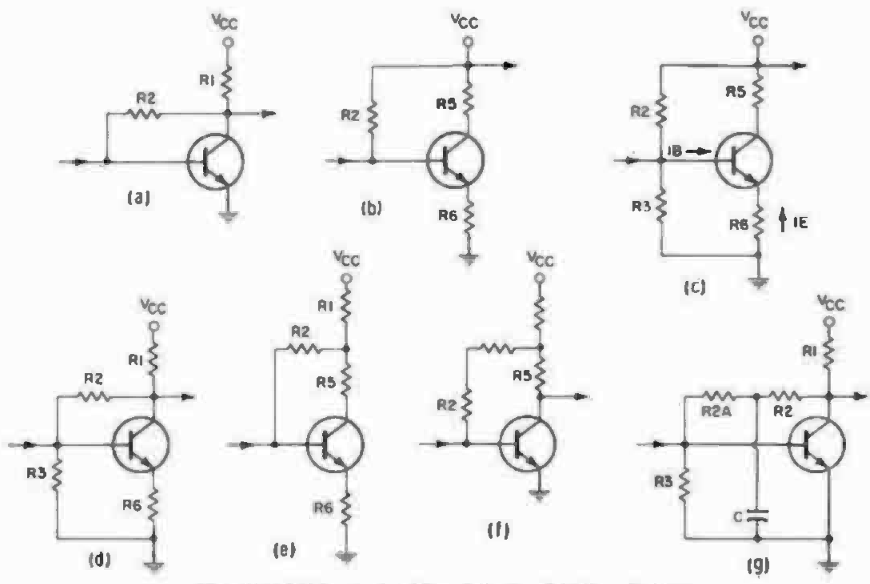


Fig. 11-133A. Basic bias circuits for transistors.

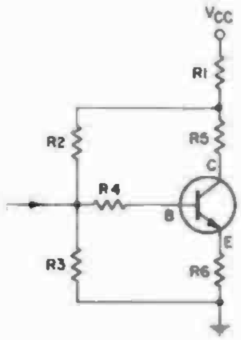


Fig. 11-133B. Basic design circuit for transistor bias circuits.

establish the bias voltage which sets the transistor idling current, or the current flow in the quiescent state. This subject is discussed at further length in Section 12.

11.134 Describe the distortion characteristics of a transistor.—The harmonic distortion increases in a transistor as the operating voltages and currents are increased and approach their maximum values. Output stages should be operated class-B push-pull whenever possible. A small amount of collector current must be permitted to flow at all times to reduce the effects of crossover or notch distortion at the zero signal point.

The addition of large amounts of negative feedback can be used to further reduce distortion. Transistor amplifiers have been developed to where both the harmonics and intermodulation are almost immeasurable.

Harmonic and intermodulation measurements must be made at levels 3 to 20 dB below the rated output of the stage under measurement. This procedure is discussed in Question 23.39.

11.135 How are gain controls connected in transistor circuits?—Three commonly used methods are shown in Figs. 11-135. The control should be connected in such a manner so that it has little or no effect on the transistor impedance, and not in a feedback loop.

11.136 Define the terms, small signal and large signal, in relation to a transistor.—The transistor like the vacuum tube is nonlinear, and can be classified as a nonlinear active device. Although the transistor is only slightly nonlinear, these nonlinearities become quite pronounced at very low and very high cnrrent and voltage levels. If an ac signal is applied to the base of a transistor without a bias voltage, conduction will take place on only one-half cycle of the applied signal voltage, resulting in a highly distorted output signal. To avoid high distortion, a dc bias voltage is applied to the transistor and the operating point is shifted to the linear portion of the characteristic curve. This improves the linearity and reduces the distortion to a value suitable for small signal operation. Even though the transistor is biased to the most linear part of the characteristic curve, it can still add considerable distortion to the signal if driven into the nonlinear portion of the characteristic. (See Question 11.139.)

Small signal swings generally run from less than 1 microvolt to about 10 millivolts under normal operation conditions. Therefore, it is highly important that the dc bias voltage be of sufficient value so that the applied ac signal is small compared to the dc bias current and voltage. Transistors are normally

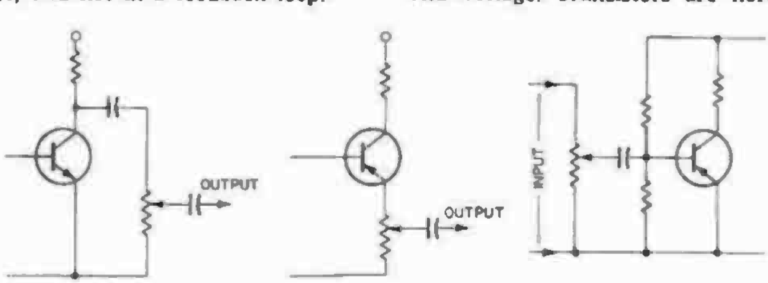


Fig. 11-135. Three methods used for connecting gain controls in transistor circuits.

biased at current values between 0.1 and 10 milliamperes. For large signal operation, the design procedures become quite involved mathematically and require a considerable amount of approximation and the use of nonlinear circuit analysis.

11.137 On what basis are audio transformers selected for transistor use? —Because of the wide difference of impedance between the input and output circuits of transistors, care must be taken to provide an impedance match between cascaded stages, or an appreciable loss of power will take place.

The maximum power amplification is obtained with a transistor when the source impedance matches the internal input resistance, and the load impedance matches the internal output resistance. The transistor is then said to be "image matched."

If the source impedance is changed in value it affects the internal output resistance of the transistor, necessitating a change in the value of the load impedance. When transistor stages are connected in tandem, except for the grounded-emitter connection, the input impedance is considerably lower than the preceding stage. Therefore, the interstage transformer must supply an impedance match in both directions.

When working between a grounded-base and a grounded-emitter circuit, a step-down ratio transformer must be used. Working into a grounded-collector stage, a step-up ratio is used. Grounded-collector stages are sometimes used as an impedance-matching device between other transistor stages. When adjusting the battery voltages for a transistor amplifier employing transformers, the battery voltage must be increased to compensate for the dc voltage drop across the transformer windings. The manufacturer's data sheets should be consulted before selecting a transformer, to determine the source and load impedances. Standard transformers are available, as for vacuum tubes. (See Question 11.139.)

11.138 How is the noise factor (NF) for a transistor amplifier calculated?— In a low-level amplifier such as a preamplifier, noise is the most important single factor, and is stated as the signal-to-noise ratio or *noise factor*. Most amplifiers employ resistors in the input circuit, which contribute a certain amount of measurable noise because of thermal activity. This power is generally about 160 dB below the power of 1 watt for a bandwidth of 10,000 Hz. When the input signal is amplified, the noise is also amplified. If the ratio of the signal power to noise power is the same, the amplifier is noiseless and has a noise figure of unity or more. However, in a practical amplifier some noise is present, and the degree of impairment is called the noise figure (NF) of the amplifier, expressed as the ratio of signal power to noise power at the output:

$$NF = \frac{S_i/N_i}{S_o/N_o}$$

where,

S_i is the signal power,
N_i is the noise power,
S_o is the signal power at the output,
N_o is the noise at the output.

The NF in dB is 10 times the logarithm of the power ratio. Thus, for an amplifier with a 1-dB noise figure, it decreases the signal-to-noise ratio by a factor of 1.26; a 3-dB NF by a factor of 2; a 10-dB NF by a factor of 10; and a 20-dB NF by a factor of 100. Amplifiers with an NF below 6 dB are considered excellent.

Low noise factors can be obtained by the use of an emitter current of less than 1 milliampere, a collector voltage of less than 2 volts, and a signal-source resistance below 2000 ohms.

11.139 Describe the procedure for using transistor load lines.—The use of transistor load lines is similar in every respect to that used with vacuum tubes. The transistor selected in this question has the following characteristics: maximum collector current, 10 milliamperes; maximum collector voltage, minus 22 volts; base current, zero to 300 microamperes; maximum power dissipation, 300 milliwatts. The base-current curves are shown in Fig. 11-139A. The amplifier circuit is to be class-A, using a common-emitter circuit (Fig. 11-139B). By proper choice of the operating point, with respect to the transistor characteristics and supply voltage, low distortion, class-A performance is easily obtained within the transistor power ratings.

The first requirement is a set of collector-current, collector-voltage curves for the transistor to be employed. Such curves can generally be obtained from the manufacturer's data sheets. Assum-

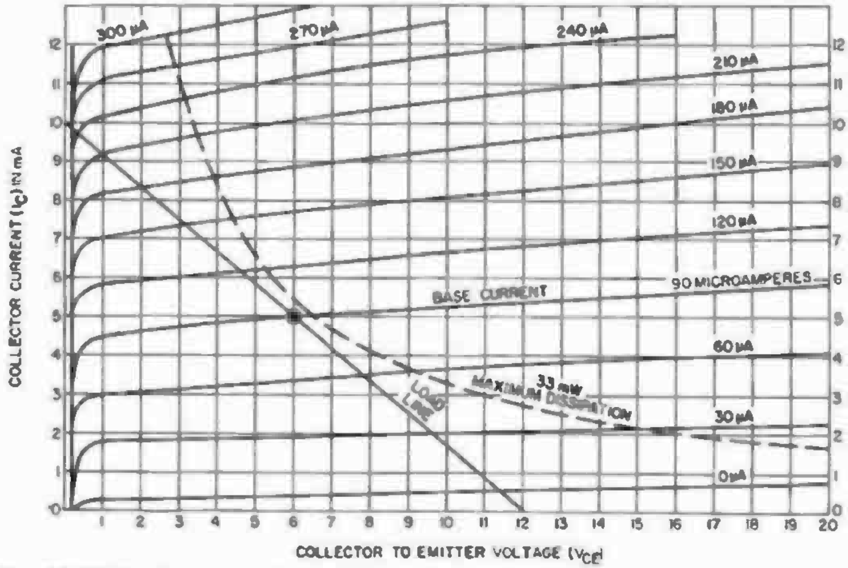


Fig. 11-139A. Common-emitter-collector family of curves, with load line and maximum dissipation power curve.

ing such data are at hand, and referring to Fig. 11-139A, a curved line is plotted on the data sheet, representing the maximum power dissipation by the use of the equation:

$$I_c = \frac{P_c}{V_c} \quad \text{or} \quad V_c = \frac{P_c}{I_c}$$

where,

I_c is the collector-current,
V_c is the collector-voltage,
P_c is the maximum power dissipation of the transistor.

At any point on this line at the intersection of V_cI_c, the product equals 0.033 watt or 33 milliwatts. In determining the points for the dissipation curve, voltages are selected along the horizontal axis and the corresponding current equated:

$$I_c = \frac{P_c}{V_{CE}}$$

The current is determined for each of the major collector-voltage points, starting at 16 volts and working backward until the upper end of the power curve intersects the 300 microampere base-current line.

After entering the value on the graph for the power dissipation curve, the area to the left of the curve encompasses all points within the maximum dissipation rating of the transistor. The area to the right of the curve is the overload region and is to be avoided. The operating point is next determined. A point that results in less than 33-milliwatts dissipation is selected some-

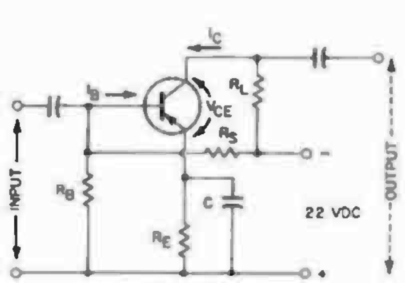


Fig. 11-139B. Amplifier circuit used for load-line calculations.

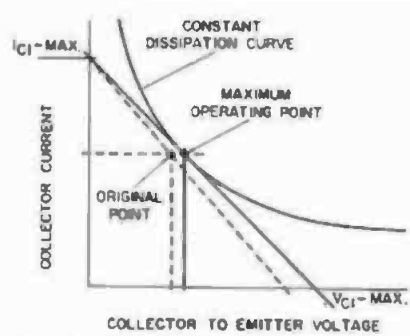


Fig. 11-139C. Load line moved to right for maximum power output. Dotted lines are the original load line and operating point.

where near the center of the power curve. For this example, 5 milliamperes collector current at 6 volts, or a dissipation of 30 milliwatts, will be used. The selected point is indicated on the graph and circled for reference. A line is drawn through the dot to the maximum collector current, 10 milliamperes, and downward to intersect the V_{CE} line at the bottom of the graph, which, for this example, turns out to be at 12 volts. This line is termed the load line. The load resistance R_L may be computed:

$$R_L = \frac{dV_{CE}}{dI_C} = \frac{0-12}{0-0.01}$$
$$= \frac{12}{0.01} = 1200 \text{ ohms}$$

where,

R_L is the load resistance,
dV_{CE} is the range of collector to emitter voltage,
dI_C is the range of collector current.

Under these conditions, the entire load line dissipates less than the maximum value of 33 milliwatts, with 90 microamperes of base current and 5 milliamperes of collector current. The required base current of 90 microamperes may be obtained by means of one of the biasing arrangements shown in Fig. 11-133.

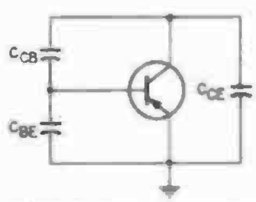


Fig. 11-140A. Internal capacitance of a transistor.

To derive the maximum power output from the transistor, the load line may be moved to the right and the operating point is placed in the maximum dissipation curve, as shown in Fig. 11-139C. Under these conditions, an increase in distortion may be expected. As the operating point is now at 6.5 volts and 5 milliamperes, the dissipation is 33 milliwatts. Drawing a line through the new operating point and 10 milliamperes (the maximum current), the voltage at the lower end of the load line is 13.0 volts; therefore, the load impedance is now 1300 ohms. It should be borne in mind, values shown on the graph are for a particular type transis-

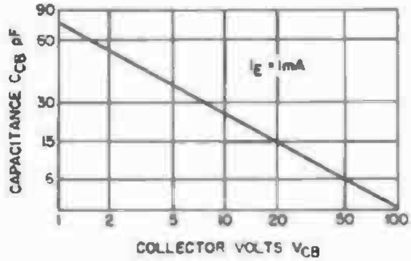


Fig. 11-140B. Variation of C_{CB} with collector voltage.

tor; other types will have a similar family of curves, but not of the same value.

11.140 How does the internal capacitance of a transistor affect its operation?—The paths of internal capacitance in a typical transistor are shown in Fig. 11-140A. Since the width of the p-n junction in the transistor will vary in accordance with voltage and current, the internal capacitance also varies. Variation of collector-base capacitance (C) with collector voltage and emitter current is shown in Figs. 11-140B and C. The *increase* in the width of the p-n junction between the base and collector, as the reverse bias voltage (V_{CB}) is increased, is reflected in lower capacitance values. This phenomenon is equivalent to increasing the spacing between the plates of a capacitor. An increase in the emitter current, most of which flows through the base-collector junction, *increases* the collector-base capacitance (C_{CB}). The increased current through the p-n junction may be considered as effectively *reducing* the width of the p-n junction. This is equivalent to decreasing the spacing between the plates of a capacitor, therefore increasing the capacitance.

The average value of collector-base capacitance (C_{CB}) varies from 2 pF to 50 pF, depending on the type transistor and the manufacturing techniques. The

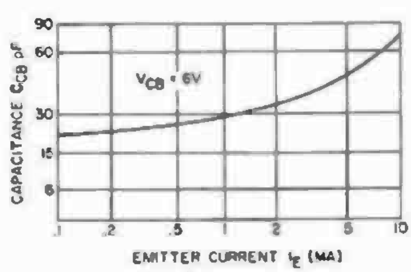


Fig. 11-140C. Variation of C_{CB} with emitter current.

collector-emitter capacitance is caused by the p-n junction, and normally is 5 to 10 times greater than that of the collector-base capacitance, and will vary with the emitter current and collector voltage. For certain applications advantage is taken of this effect for control purposes.

11.141 How is stability factor (SF) calculated for transistors?—The current through a transistor tends to increase with temperature; therefore it is necessary in the design of transistor circuits to include a stability factor (SF) to limit the current dissipation to a safe value, under the expected highest operating temperatures. Stability is expressed as the ratio between a change in dc collector current (I_C) and the corresponding change in dc collector cutoff current, with the emitter open circuit current (I_{CBO}). For a given set of operating voltages, the SF can be calculated for a maximum permissible rise in dc collector current from the ambient temperature value:

$$SF = \frac{I_{CMax} - I_{C1}}{I_{CBO2} - I_{CBO1}}$$

where,

- I_{C1} and I_{CBO2} are measured at 25 degrees centigrade,
- I_{CBO1}, at the maximum expected ambient or junction temperature,
- I_{CMax} is the maximum permissible collector current for a specified collector-to-emitter voltage at the maximum expected ambient or junction temperature.

The maximum value SF can never be *greater* than the value of beta. The stability from a feedback standpoint can be increased by the use of decoupling circuits in the power supply circuits, much in the same manner as for vacuum tubes, discussed in Question 12.36.

11.142 What are the frequency characteristics of transistors used for audio amplifiers?—As a rule, conventional transistors used for audio-amplifier application have bandwidths up to 4 MHz, the average being 1 to 2 MHz. When connected in complex circuitry, the response may be considerably reduced because of circuit capacitance and other design factors. This is particularly true for high-power output stages. If audio transformers are used, naturally the response is limited to the frequency range of the transformers.

Using a well designed circuit and mechanical layout, the frequency response may be designed to cover a spectrum from 5 to 50,000 Hz. In amplifiers using field-effect transistors, (FET) bandwidths from 5 to 250,000 Hz have been achieved. Generally speaking, the high frequency limit of a diode or transistor is a function of the base thickness.

Because of the wide frequency response of transistor amplifiers, precautions must be observed to prevent positive feedback (oscillation). Oscillation may occur in the megahertz range, and is often the cause of failure. By the use of properly designed printed circuit boards, oscillation can be prevented.

11.143 What type push-pull amplifier circuit is recommended for transistors?—One of the disadvantages of using transistors class-A, either single-ended or push-pull, is that the collector current flows at all times, and dissipation is the highest when no ac signal is present. This dissipation of power can be greatly reduced by the use of class-B push-pull operation. To achieve the full power output and economy of operation, the transistors should be biased to full cutoff. However, in so doing, a serious distortion problem arises, as shown in Fig. 11-143. Assuming that the transistors have identical characteristics, the variation in the collector currents are plotted against the base current under normal load conditions.

The overall dynamic characteristics are obtained by plotting individual curves of the two transistors back-to-back, and combining them as shown. Applying a sine wave to the input,

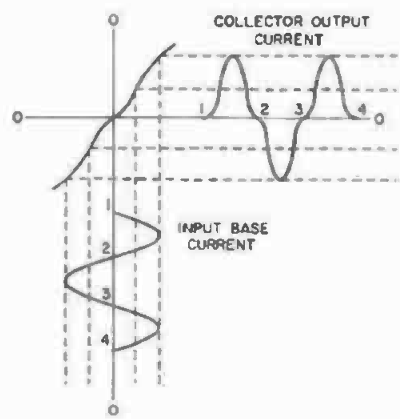


Fig. 11-143. Dynamic transfer characteristics curves of a class-B, transistor push-pull amplifier with zero bias, showing the distorted output waveform.

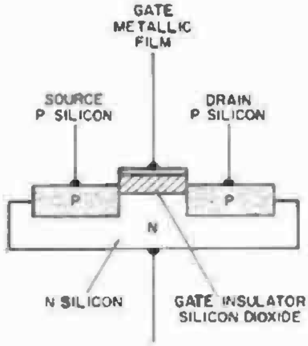


Fig. 11-144A. Internal construction of an insulated-gate transistor (IGT).

points of the input base current are projected into the dynamic transfer characteristics curve. Correspouding points are then projected to form the output collector-current waveform. It will be observed that severe distortion is generated at the *crossover* points, where the signal passes through the zero value. This is termed *crossover distortion* or *notch distortion*, and increases in severity as the signal level is lowered. Crossover distortion is eliminated or greatly reduced by the use of a small bias voltage on both transistors, resulting in a smooth crossover (applies to both transistors and vacuum tubes). In addition to the bias correction, negative feedback is used to further reduce the distortion. (See Question 12.230.)

11.144 Describe an insulated-gate transistor. (IGT) — Insulated-gate transistors (IGT) are also known as a field-effect transistor (FET), metal-oxide silicon or semiconductor field-effect transistors (MOSFET), metal-oxide silicon or semiconductor transistors (MOST), and insulated-gate field-effect transistors (IGFET). All of these devices are similar, and are simply names applied to them by the different manufacturers.

The outstanding characteristics of the IGT is its extremely high input impedance, running to 10^{18} ohms. IGT's have *three elements, but four connections.* The type elements are the gate, drain, and the source. The fourth connection goes to an n-substrate. The construction consists of an n-type substrate, into which two identical p-type silicon regions have been diffused. The source and drain terminals are taken from these two p regions, which form a capacitance between the n substrate and the silicon-dioxide insulator and the metallic gate terminals. A cross-sectional view of the internal construction appears in Fig. 11-144A, with a basic circuit shown in Fig. 11-144B. Because of the high input impedance, the IGT can easily be damaged by static charges. Strict adherence to the manufacturer's instructions must be followed, since the device can be damaged even before putting it into use.

IGT's are used in electrometers, logic circuits, and ultrasensitive electronic instruments. They should not be confused with the *conventional FET*, used in audio equipment and discussed in Question 11.145.

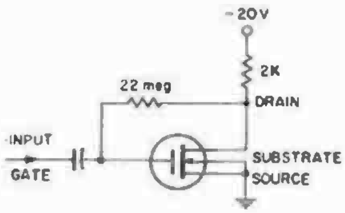


Fig. 11-144B. Typical circuit.

11.145 Describe the basic principles of a field-effect transistor (FET).—The field-effect transistor, or FET as it is commonly known, was developed by the Bell Telephone Laboratories in 1946, but was not put to any practical use until about 1964. The FET is often

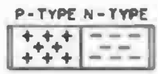


Fig. 11-145A. Creation of depletion region without bias voltage.

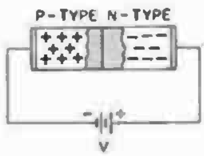


Fig. 11-145B. Creation of depletion region with reverse-bias voltage.

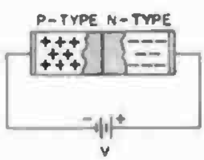


Fig. 11-145C. Effects due to increase in bias voltage.

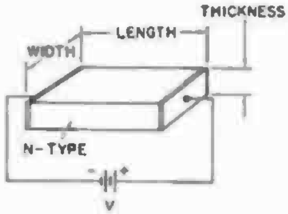


Fig. 11-145D. Plain semiconductor bar.

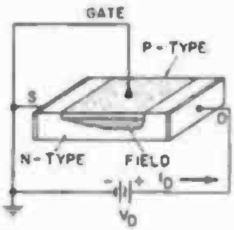


Fig. 11-145E. Bar with gate added and drain voltage applied.

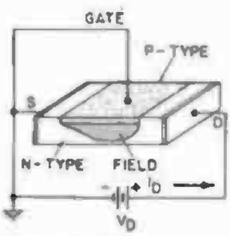


Fig. 11-145F. Field increases with increase of drain voltage.

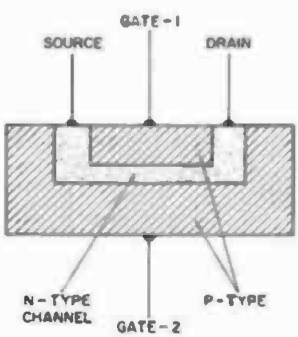


Fig. 11-145G. Cross-sectional view of the construction for a single- or double-gate FET.

quoted as being the semiconductor counterpart of the vacuum tube. The principal difference between a conventional transistor and the FET lies in the fact that the transistor is a *current-controlled* device, while the FET is *voltage-controlled*, similar to the vacuum tube. Conventional transistors also have a low input impedance, which may at times complicate the circuit designer's problems. The FET has a *high* input impedance with a *low* output impedance; this characteristic caused it to be likened to the vacuum tube.

The basic principles of the FET operation can best be explained by the simple mechanism of a p-n junction. The control mechanism is the creation and control of a depletion layer, which is common to all reverse-biased junctions. Referring to Fig. 11-145A, atoms in the n-region possess excess electrons which are available for conduction, and the atoms in the p-region have excess holes which may also allow current to flow (the reader is referred to Question 11-114 for the theory of holes and carriers in semiconductors.) Reversing the voltage applied to the junction, Fig. 11-145B, and allowing time for stabilization, very little current flows but a rearrangement of the electrons and holes will occur. The positively charged holes will be drawn toward the negative terminals of the voltage source, and the electrons which are negative will be attracted to the positive terminal of the voltage source. This results in a region being formed near the center of the junction having a majority of the carriers removed. This is true for both the p and n sides of the junction. The areas near the junction where the carriers have been removed are called the *depletion regions.*

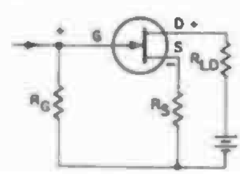


Fig. 11-145H. N-channel FET circuit.

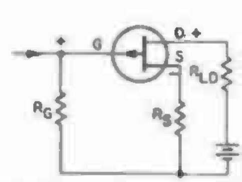


Fig. 11-145I. P-channel FET circuit.

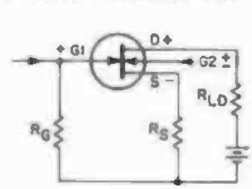


Fig. 11-145J. N-channel double gate FET circuit.

After stabilization, there will be an equal number of holes removed from the p region as there are electrons from the n region. Thus, the distauce that the depletion area extends from the junction is the fuuction of the relative carrier densities. For purpose of illustration, it will be assumed that the n region has less electrons than the p region has holes. The depletion region must then extend further into the n region to have the same number of electrons as there are holes, and in a much smaller volume of p-type material.

Increasing the reverse voltage causes the depletion region to reach farther into the semiconductor material. The result of increasing the voltage is shown in Fig. 11-145C. Referring to Fig. 11-145D, a simple bar composed of n-type semiconductor material is shown, with two nonrectifying contacts at each end. As there is always a certain amount of resistance (R) between the two end electrodes, (P) will be the function of the material sensitivity, (L) the length of the bar, (W) the width and (T) the thickness. Resistivity may be equated:

$$R = \frac{PL}{WT}$$

By varying one or more of the variables of the resistance of the semiconductor, the bar may be changed. Assume a p region in the form of a sheet is formed at the top of the bar shown in Fig. 11-145E. This may be accomplished by diffusion, alloying, or epitaxial growth, and a p-n junction is formed. Thus, a reverse voltage appearing between the p and n material produces two depletion regions. Current in the n material is primarily by means of excess electrons. Reducing the concentration of electrons or majority carriers, the resistivity of the material is increased. Removal of the excess electrons by means of the depletion region causes the material to become practically nonconductive.

Disregarding the p region and applying a voltage to the ends of the bar causes a current and creates a potential gradient along the length of the bar material, with the voltage increasing toward the right, with respect to the negative end or ground. Connecting the p region to ground causes varying amounts of reverse-bias voltage across the p-n junction, with the greatest amount developed toward the right end of the p region. A reverse voltage across the bar will produce the same depletion regions. If the resistivity of the p-type material is made much smaller than that of the n-type material, the depletion region will then extend much farther into the n material than into the p material. To simplify the following explanation, the depletion of p material will be ignored.

The general shape of the depletion is that of a wedge, increasing in size from left to right. Since the resistivity of the bar material within the depletion area is increased, the effective thickness of the conducting portion of the bar becomes less and less, going from the end of the p region to the right end. This indicates the overall resistance of the semiconductor material is larger, because the effective thickness is being reduced. Increasing the voltage applied to the ends of the bar (Fig. 11-145F) extends the depletion region, and further decreases the resistance. Continuing to increase the voltage across the ends of the bar, a point is reached where the depletion region is extended practically all the way through the bar, reducing the effective thickness to zero. Increasing the voltage beyond this point produces little change in current.

Because of the controlling action of the p region, it is termed a *gate*. The left end of the bar being the source of majority carriers, it is termed the *source*. The right end being where the electrons are drained off, it is called the *drain*. A cross-sectional drawing of a typical FET is shown in Fig. 11-145G, and three basic circuits are shown in Figs. 11-145H to J. Double-gate FET's are also in use. The chief advantage of the FET is its low internal noise, simplicity of construction, and its similarity to a vacuum tube, which makes it highly desirable for voltage-amplifier stages. One of the interesting characteristics of a FET is its ability to operate over a wide range of temperature. It has been stated by one manufacturer that FET's have been developed that perform satisfactorily even when submerged in liquid nitrogen at a temperature of minus 200 degrees centigrade, and have also been operated at a temperature of plus 150 degrees centigrade.

11.146 What is the procedure for derating a transistor? — Ratings have

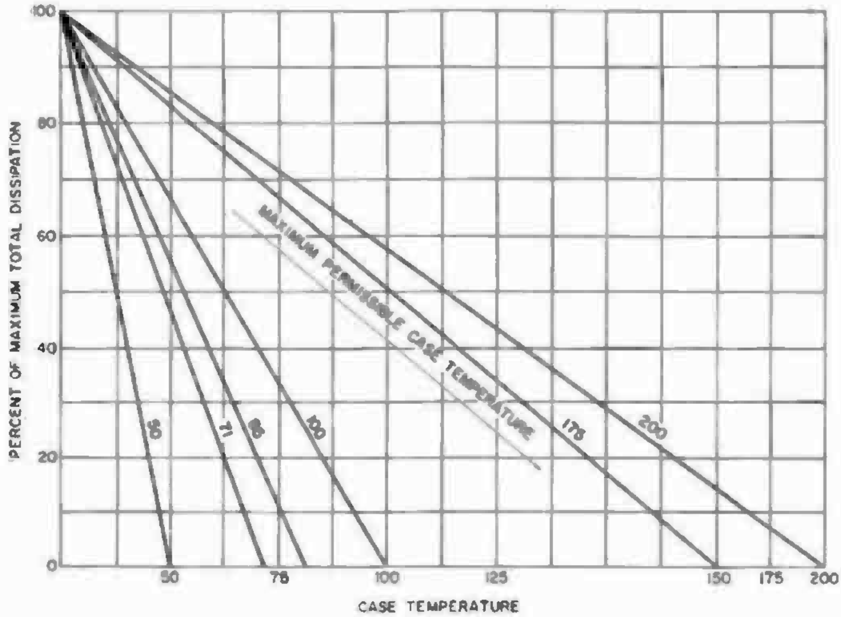


Fig. 11-146. Derating chart showing the maximum permissible percentage of maximum rated dissipation as a function of temperature.

been established for transistors and other semiconductor devices to prevent the overloading and possible damage to such devices, and have been defined by the Joint Electron Device Engineering Council (JEDEC) and the National Electrical Manufacturers Association (NEMA) and Electrical Industries Association (EIA). Absolute maximum ratings, electrode voltage, and current ratings are the limiting factors. Transistor dissipation is the power dissipated in the form of heat by the collector element. It is the difference between the power supplied to the collector and the power delivered by the transistor to the load. For many types of transistors the maximum dissipation is specified for ambient temperature, case, and heat-sink temperature up to 25°C. If the maximum operating temperature is to be higher than that specified by the manufacturer, the device must be *derated*. This may be done quite conveniently by the use of the chart in Fig. 11-146. Knowing the maximum dissipation rating and maximum operating temperature, a vertical line is drawn at the desired operating temperature on the abscissa to intersect the curve representing the maximum operating temperatures specified for the particular transistor. A horizontal line is drawn from this intersection to the ordinate establishing a new maximum dissipation for the device at the new operating temperature.

As an example: The maximum permissible dissipation for a transistor at a case temperature of 100°C is to be determined. The transistor in question has a maximum dissipation rating of 75 watts at a case temperature of 25°C, and a maximum permissible case temperature of 200 degrees. A perpendicular line is drawn from the 100-degree point to the 200-degree curve. The point is projected to the left and indicates a percentage of 57.5. Therefore, the maximum dissipation for this particular transistor at a case temperature of 100° C is 0.575 times 75, or approximately 43.2 watts.

11.147 Describe the basic construction of a solid-state diode rectifier.—Diodes are two element devices consisting of a pellet of either germanium or silicon, using point-contact or junction construction. However, the majority of diodes manufactured at the present time use the junction-type construction. A small diode pictured in Fig. 11-147 is used in radio and television receivers, harmonic generators, frequency discrimiuators and many applications too numerous to mention. Diodes require no heater or filament and have no contact potential, but have a

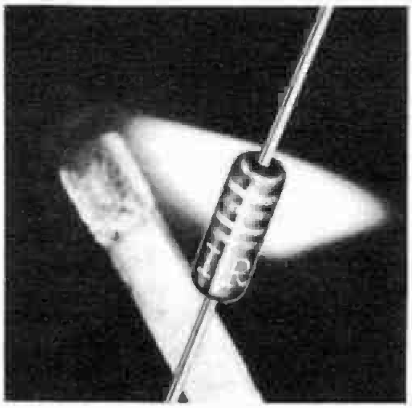


Fig. 11-147. External appearance of a silicon or germanium diode.

high forward conductance, with a high reverse or back resistance. Diodes that will carry a few milliamperes up to several hundred amperes over a wide range of voltages are available. The diode concept is discussed in Question 11-114. Applying a positive voltage to the plus terminal of the diode causes a current through the junction and back to the source. Reversing the voltage, the current is blocked by the high back resistance of the junction.

The conduction characteristics of a silicon diode differ from its counterpart, the germanium diode. Silicon diodes have unusually high back resistance, thereby passing a much lower reverse current. As an example, a 1N251 silicon diode passes in the reverse direction only 0.2 microampere at 10 volts. Applying the same voltage to a 1N34 germanium diode of similar characteristics, 50 microamperes of current are passed in the reverse direction. However, germanium diodes are employed extensively in many different types of equipment. (See Questions 11.148 to 11.155.)

11.148 Describe the characteristics of a zener diode.—The zener diode, named for Dr. Carl Zener, is a semiconductor equivalent of the gaseous regulator tube discussed in Question 11-30. Whereas the gaseous tube regulator is confined to a limited range of voltage operation, zener diodes are available in a wide range of operating voltages, ranging from 2 to several hundred volts, with large power-dissipation capabilities.

A zener diode is a semiconductor device with a substantially constant reverse-voltage characteristic over a wide range of reverse current or temperature. When the reverse voltage of a semiconductor diode is increased, its internal resistance remains at a high value for a short time. When a critical voltage is reached, the resistance falls suddenly. The current flow up to the critical point is on the order of microamperes; however, at breakdown it may be several millamperes, or in the instance of a large diode, it may be several amperes.

The point of breakdown is termed the zener or avalanche point (see Question 11.149). Once this point is reached, the voltage drop across the diode remains fixed, and the current changes only if more voltage is applied. The breakdown of a diode should not be construed as damaging, as such diodes are designed to be operated in and out of the breakdown region without damage or a change in their characteristics. In this manner the diode may be used to provide a constant voltage drop or reference voltage across its internal resistance.

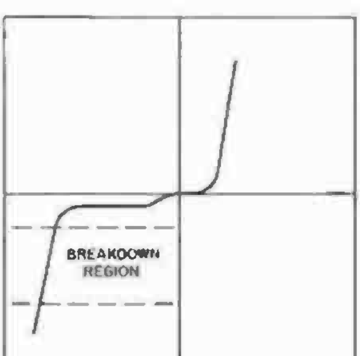


Fig. 11-148A. Zener forward and reverse current characteristics.

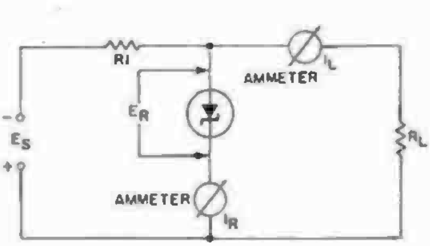


Fig. 11-148B. Voltage-regulator circuit using a zener diode.

WHERE:

R_1 IS THE SERIES RESISTOR
E_S IS THE SOURCE VOLTAGE
E_R IS THE REGULATOR VOLTAGE
I_R IS THE REGULATOR CURRENT
I_L IS THE LOAD CURRENT
P_R IS THE REGULATOR POWER DISSIPATION

$$R_1 = \frac{E_S - E_R}{I_R + I_L}$$

$$I_R = \left(\frac{E_S - E_R}{R_1}\right) - I_L$$

$$P_R = \left(\frac{E_S - E_R}{R_1} - I_L\right) E_R$$

FOR VARIABLE SOURCE VOLTAGE AND LOAD CURRENT:

$$R_1 = \frac{E_S(MIN.) - E_R}{I(MAX.) + .1I_L(MAX.)}$$

$$P_R(MAX.) = \left(\frac{E_S(MAX.) - E_R}{R_1} - I_L(MIN.)\right) E_R$$

FOR CONSTANT LOAD CURRENT BUT VARIABLE INPUT VOLTAGE:

$$R_1 = \frac{E_S(MIN.) - E_R}{I_L + .1I_L}$$

$$P_R(MAX.) = \left(\frac{E_S(MAX.) - E_R}{R_1} - I_L\right) E_R$$

FOR CONSTANT INPUT VOLTAGE BUT VARIABLE LOAD CURRENT:

$$R_1 = \frac{E_S - E_R}{I_L(MAX.) + .1I_L(MAX.)}$$

$$P_R(MAX.) = \left(\frac{E_S - E_R}{R_1} - I_L\right) E_R$$

NOTE:

THE ABOVE EQUATIONS ALLOW A TOLERANCE OF 10% TO COMPENSATE FOR LOAD REGULATION. IF BREAKDOWN IMPEDANCE IS A SIGNIFICANT PERCENTAGE OF THE VALUE OF R, THIS MUST BE TAKEN INTO CONSIDERATION. A HIGH IMPEDANCE SOURCE PRESENTS ADDITIONAL PROBLEMS AND MUST BE CONSIDERED IF IT IS SIGNIFICANT COMPARED TO R_1.

Fig. 11-148C. Equations for designing zener-diode regulator circuit of Fig. 11-148B.

Zener diodes may be constructed using germanium or silicon, with point-contact or junction construction. However, the majority are of the junction type. The conduction characteristics of the silicon diode differ from its counterpart, the germanium diode. Silicon diodes have unusually high back resistance, passing much lower reverse current than the germanium diode. A typical plot of the conduction characteristics for a germanium and silicon diode is shown in Fig. 11-148A. As a comparison, a 2N251 silicon diode passes in the reverse direction only 0.20 microampere at 10 volts, while a germanium diode type 1N34 passes 50 microamperes for the same voltage.

A voltage regulator circuit, employing a silicon diode is given in Fig. 11-148B. It can be seen that a relatively small increase in voltage across the diode causes a large increase in current through resistor R1, thus increasing the voltage drop. This voltage drop across R1 effectively maintains a constant volt-

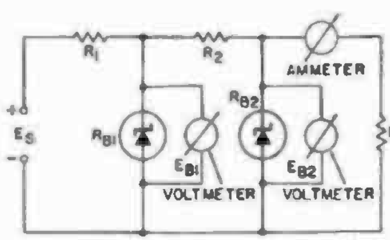


Fig. 11-148D. Zener-diode regulator circuit using a two-stage regulator.

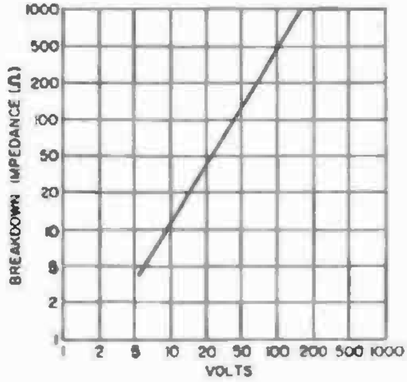


Fig. 11-148E. Breakdown impedance versus zener voltage.

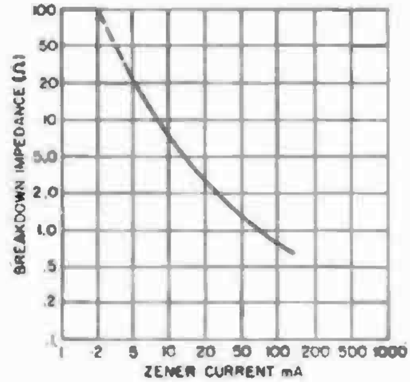


Fig. 11-148F. Breakdown impedance versus zener current.

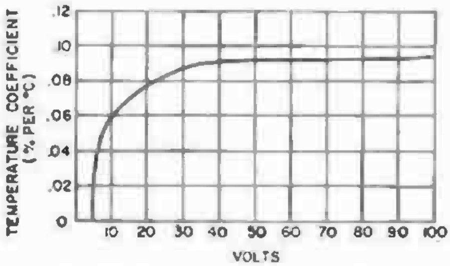


Fig. 11-148G. Temperature coefficient versus zener volts.

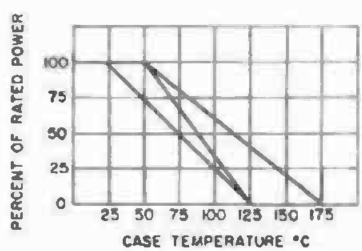


Fig. 11-148H. Typical zener derating curves.

age across the load. The value of resistor R1, regulator current, and power rating, is given in the equations in Fig. 11-148C. The circuitry shown will compensate for the normally encountered variations of input and output load voltages. Extreme variations may require the addition of a second diode regulator as shown in Fig. 11-148D.

Breakdown impedance offers a reasonable measurement to the diode's ability to maintain a nearly constant breakdown voltage, and, consequently, its regulation characteristics. The breakdown voltage and current for a 1-watt 7.5-volt regulator diode is plotted in Figs. 11-148E and F respectively. Variations with temperature are shown in Fig. 11-148G for a range of 5.6 to 100 volts. The curve, as will be noted, is quite constant above 50 volts, then changes rapidly at the lower voltages.

Silicon diodes are rated relative to their case temperature, therefore the manufacturer's data sheet should be consulted and the diode derated accordingly. Typical derating curves are given in Fig. 11-148H. Stud-type diodes require cooling fins to dissipate the heat. Cooling fins will dissipate about 8 milliwatts per square inch, per degree centigrade. Therefore, a 10-watt regulator diode operating close to maximum rating would require an area of $10/0.008 \times 60 \times 20$ square inches to limit the rise to 60 degrees centigrade. Since both sides of the fins radiate, the dimensions would be approximately 3⅛ inches by ⅛-inch thick.

Zener diodes are also used as reference voltage devices and are quite useful in voltage-regulated power supplies or any place requiring an accurate source of reference voltage. The standard tolerance for voltage-reference diodes is plus-minus 10 percent, but may be obtained for 1 and 5 percent accuracy. In a silicon reference diode, the reverse current remains quite small until the breakdown voltage point is reached, then increases rapidly with little further increase in voltage. The breakdown point being the function of the semiconductor material and its construction, it can be controlled in manufacture to supply a given voltage ranging from one to several hundred volts, at various current and power ratings and the cooling method used. Design procedures for zener diodes are given in Question 21-118.

11.149 Define the term "avalanche" as applied to semiconductor devices.— Avalanche is the breakdown point in a semiconductor material caused by the multiplication of the carriers through ionization, and is termed the avalanche point. The avalanche point was proposed by McKay, and elaborated on by many others.

In the early development stages of the zener diode, it was believed the breakdown was due to the Zener effect, thus its name. However, it is now thought that both the Zener effect and avalanche are operative, with the latter process being the most predominant, especially at voltages greater than 6 volts. Zener diodes are sometimes referred to as avalanche diodes; however, the name zener diode prevails.

Both the inverse-peak voltage and avalanche point increase with temperature; therefore, as a safety measure the circuitry is designed for an inverse-peak voltage about 20 to 25 percent *lower* than the breakdown voltage of the diode. Manufacturers take advantage of these characteristics, using doping techniques to control the resistivity, thus controlling the breakdown-voltage point and operating temperature.

11.150 Describe the basic principles of a silicon controlled rectifier (SCR).—

Silicon controlled rectifiers (SCR) are semiconductor devices that may be used as a latching static switch, a sensitive amplifier, a controlled rectifier, and many devices where the precise control of output current is required. An SCR may be turned on in 1 to 4 microseconds and turned off in 10 to 20 microseconds. Units with current capabilities of ½ ampere to several hundred

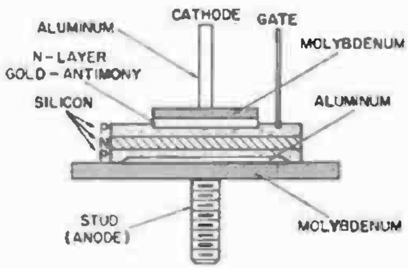


Fig. 11-150A. Cross-sectional view of the interior construction of a silicon controlled rectifier (SCR).

amperes are available with comparable voltage ratings. The basis of the SCR is a disc of four alternate layers of n- and p-type silicon (Fig. 11-150A). Besides the two terminals of the conventional diode (a cathode and anode), the SCR has a gate terminal for controlling the conduction cycle. A cutaway view of its interior construction is shown in Fig. 11-150B.

Referring again to Fig. 11-150A, to protect the silicon junction against thermal and mechanical injury, the silicon discs are braised between plates of molybdenum or tungsten. These plates have the same coefficient of expansion as the silicon. In the small lead-mounted type SCR, the bottom plate acts as the base of the housing enclosure. In the larger types, above 1 ampere, the bottom plate is soldered to a copper stud which acts as the anode terminal and a thermo path for the heat losses to the outside ambient temperature. For this reason, the stud is usually screwed into the base of a cooling fin or heatsink, as shown in Fig. 11-159B.

The operation of the SCR may be explained as follows. On the forward-blocking region of the characteristic, increasing the forward current does not tend to increase the current until a point is reached where avalanche multiplication begins. Beyond this point, the current increases rapidly until the total current through the device is sufficient to maintain itself. At this point, the SCR will go into high conduction, provided the current through the device remains greater than the minimum current, called the holding current. If the current through the device drops below the holding current, the SCR will return to the forward-blocking region again. In the reverse direction, the SCR has essentially two back-biased p-n junctions in series, so that it exhibits

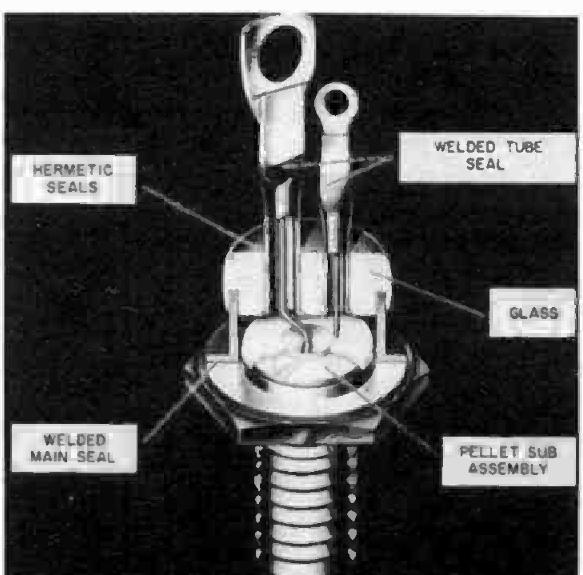


Fig. 11-150B. Interior view of a silicon controlled rectifier (SCR) as manufactured by General Electric Co.

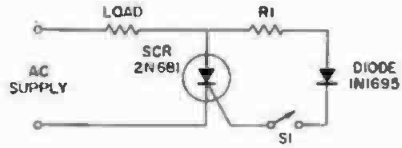


Fig. 11-150C. Firing circuit for a silicon controlled rectifier. (SCR)

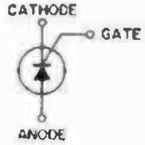


Fig. 11-150D. Symbol for a silicon controlled rectifier.

characteristics very similar to the ordinary back-biased silicon rectifier.

A simple method of obtaining gate current for firing an SCR from the main ac supply whenever the anode is positive with respect to the cathode is shown in Fig. 11-150C. As soon as the SCR 2N681 has fired, the anode voltage drops to the conduction value and the gate current decreases to a low value. Resistor R1 limits the peak gate current and has a value greater than the ac peak voltage divided by two amperes. The 1N1695 diode in the gate circuit is provided to prevent reverse voltage from being applied between the cathode and gate during the reverse half of the cycle. If desired, the diode may be connected between the gate and cathode rather than in series with resistor R1. Conduction in initiated by closing switch S1. Interruption of the load current occurs within one half cycle after opening switch S1.

11.151 What factors determine the selection of a silicon or selenium rectifier?—The selection of a suitable rectifier depends on a number of factors, namely the voltage and current requirements, temperature range of operation, and the space availability. In addition, the type circuit and equipment it is to be operated with must be taken into consideration. The subject is discussed more fully in Section 21.

11.152 Describe the basic principles of a tunnel diode.—Tunnel diodes are two-terminal semiconductor devices consisting of a single p-n junction. The basic difference between the conventional diode and a tunnel diode is in its conductivity. The p-n material used in tunnel diodes has a conductivity of over 1000 times greater than the conventional diode. Such high conductivity is obtained by increasing the amount of donor and acceptor impurities in the semiconductor material during its formation.

The tunnel diode takes its name from the tunnel effect, a process wherein a particle can disappear from one side of the barrier and reappear on the other side instantaneously, as though it had tunneled through the barrier element. Such devices are used at very high frequencies in radio receivers, in computers, and in many other devices. An interior view of a typical tunnel diode is pictured in Fig. 11-152. Although three leads are shown coming from the base, only two terminals are connected to the diode; the third lead is connected to the case for grounding. The tunnel diode was first introduced in 1959, by General Electric.

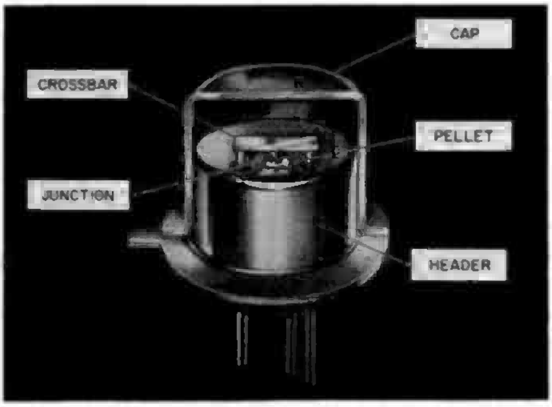


Fig. 11-152. Interior view of a tunnel diode. (Courtesy, General Electric Co.)

11.153 What is the rectifying efficiency of silicon diodes?—Because of the high forward-to-reverse current of the silicon diode, the efficiency is on the order of 99 percent. When properly used, silicon diodes have long life and are not affected by aging, moisture, or temperature when used with the proper heat sink.

11.154 Can diode rectifiers be connected in series or parallel?—Yes, the connection of several diodes in series is permissible when the applied voltage exceeds the peak-inverse voltage (PIV) for a single unit. As an example, four individual diodes of 400-volts PIV may be connected in series to withstand a PIV of 1600 volts. In a series arrangement, the most important consideration is that the applied voltage be equally distributed between the several units. The voltage drops across each individual unit must be very nearly identical. If the instantaneous voltage is not equally divided, one of the units may be subjected to a voltage exceeding its rated value.

Uniform voltage distribution can be obtained by the connection of capacitors or resistors in parallel with the individual rectifier unit. Shunt resistors are used for steady-state applications,

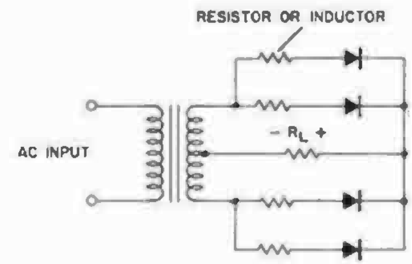


Fig. 11-154D. Diode rectifiers connected in parallel, with series resistances or small inductors.

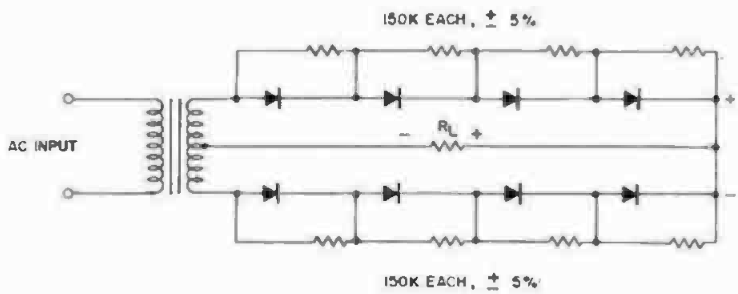


Fig. 11-154A. Diodes connected in series, with balancing resistors connected across each rectifier.

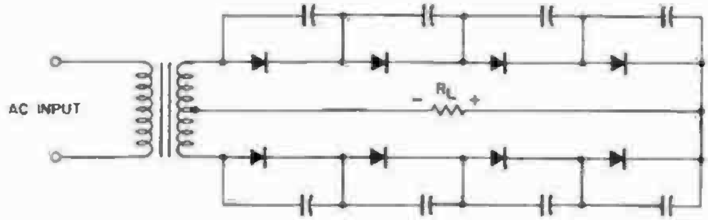


Fig. 11-154B. Diode rectifiers connected in series, with capacitors connected in parallel to reduce the effect of transient voltages.

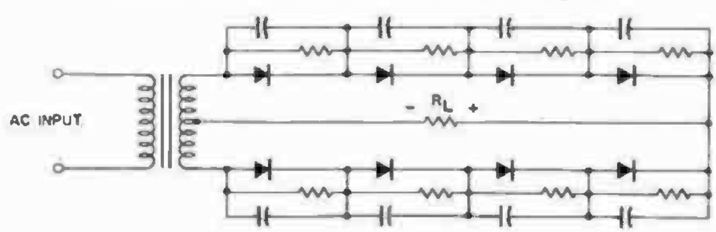


Fig. 11-154C. Diode rectifiers connected in series, with both resistors and capacitors in parallel to equalize voltage distribution and transient effects.

and shunt capacitors are used in applications where transient voltages are expected. If the circuit is exposed to both dc and ac, both shunt capacitors and resistors should be employed.

When the maximum current of a single diode is exceeded, two or more units may be connected in parallel. To avoid differences in voltage drop across the individual units, a resistor or small inductor is connected in series with each diode. Of the two methods, the inductance is favored because of the lower voltage drop and consumption of power. Parallel arrangements should be avoided if possible, by the employment of polyphase circuits or the use of rectifier units large enough to handle the required current.

Series-resistor values will vary with the current ratings of the diodes; however, resistors ranging from 1 to 10 ohms are the usual value, with the dc resistance of series chokes held to 1 ohm or less. The four methods of connection are given in Figs. 11-154A to D.

11.155 What are hot carrier diodes? —The hot carrier diode is distinguished from the more conventional semiconductor devices in that the junction consists of a metal and a semiconductor rather than two different semiconductors. The junction in a hot carrier diode is made to be rectifying rather than ohmic-resistant through a choice of materials, with suitably related work functions. In diodes thus formed, current occurs mainly by means of majority carriers. When the diode is forward biased, the majority carriers injected into the metal have a much higher energy level than those that are in thermal equilibrium with the metal. Such diodes have extremely rapid recovery time, and are used for fast switching circuits, detectors, mixers and many other uses, particularly at high frequencies. A cross-sectional view of its construction is given in Fig. 11-155.

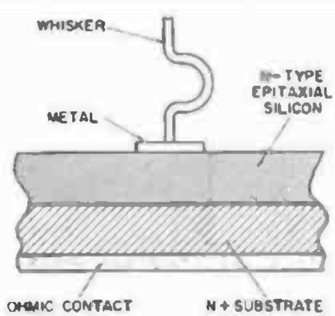


Fig. 11-155. Cross-sectional view of a hot carrier diode construction.

11.156 What is a triac semiconductor? —It is a three-lead, eight-element semiconductor developed by RCA, that passes both halves of the ac waveform when properly triggered. It is similar to an SCR, the difference being the bi-polarity of the Triac assembly, which amounts to having two SCRs back-to-back. The construction of the junctions and an experimental circuit are shown in Fig. 11-156.

11.157 Describe the construction and characteristics of a thermistor. —Thermistors are thermally sensitive resistors whose primary function is to exhibit a change in electrical *resistance* with a change of body *temperature*. They are passive semiconductors, with their electrical resistance being between that of a conductor and an insulator. An important characteristic is their extreme sensitivity to relatively minute temperature changes. While most metals have small positive coefficients of resistance, thermistors exhibit a wide range of either negative or positive temperature coefficients. This unique characteristic permits them to be used in electronic apparatus for control of

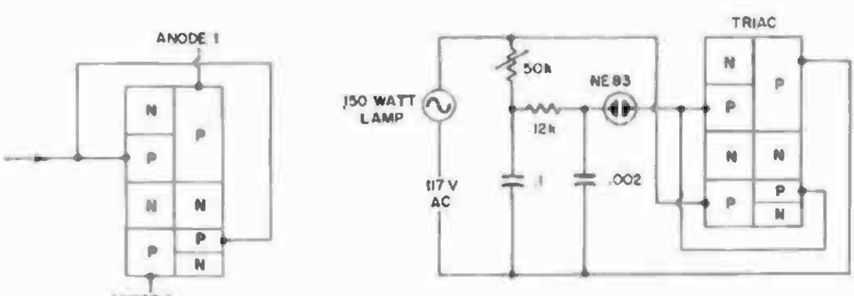


Fig. 11-156. A triac three lead, 8-element semiconductor switch for controlling the intensity of a light bulb or other devices. (Courtesy, Radio Corporation of America.)

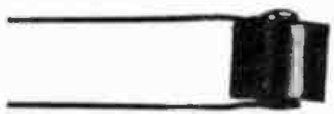


Fig. 11-157A. Thermistor manufactured by General Electric Co.

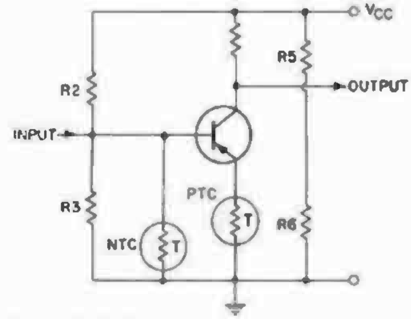


Fig. 11-157B. Thermistor used for temperature compensation in a transistor circuit. A thermistor with a negative temperature coefficient may be connected in parallel with R3, or one with a positive temperature coefficient in place of R6.

voltage and current. A typical thermistor, manufactured by General Electric Co., is shown in Fig. 11-157A.

Thermistors are widely used in transistor circuitry for voltage-temperature compensation, particularly in bias circuits, as seen in Fig. 11-157B. Here a transistor is connected in parallel with a resistor R3 for control of the bias voltage, or it may be connected in the emitter circuit for stabilization. The characteristics for parallel and series connection are shown in Fig. 11-157C, with their change in resistance for a change in temperature in Fig. 11-157D.

The standard reference temperature for thermistors is 25°C. Thermistors are manufactured in a variety of sizes and shapes, such as rods, cubes, washers, discs, beads, etc.

11.158 Describe a surge or transient-voltage suppressor.—Silicon rectifiers are sensitive to voltage, and overvoltage conditions for even a short period of time can cause their failure.

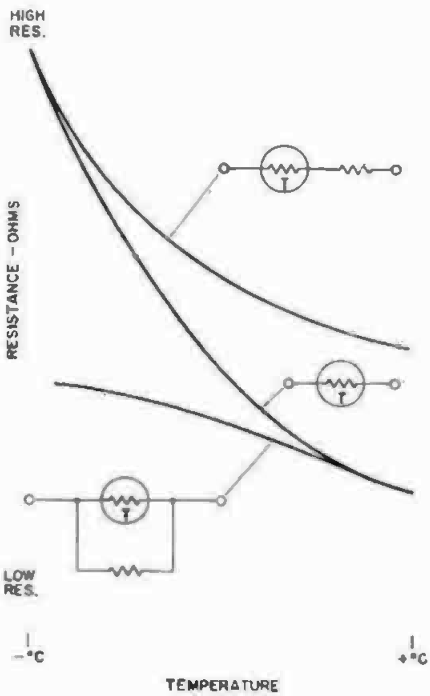


Fig. 11-157C. Temperature versus resistance for parallel series connection. (Courtesy, General Electric Co.)

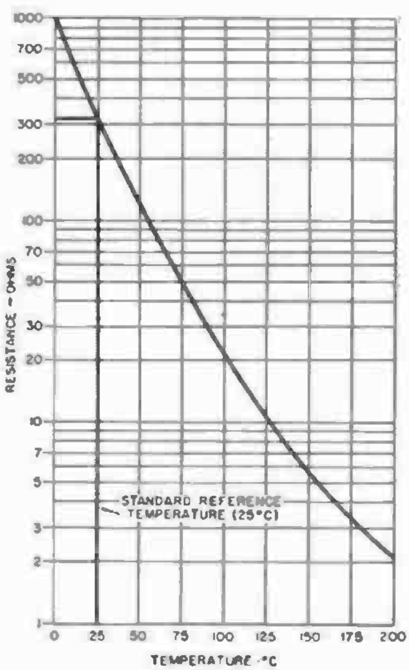
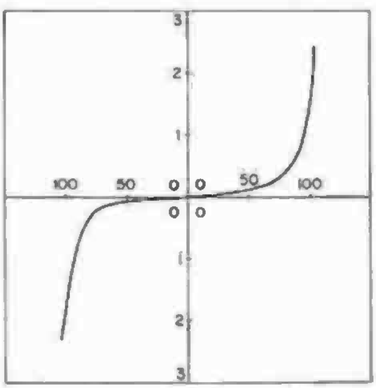


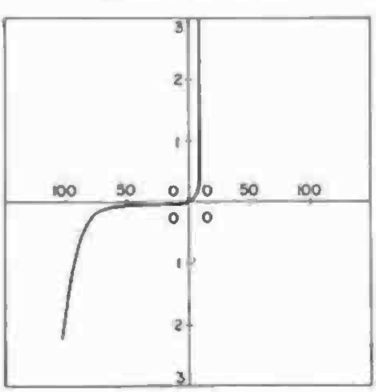
Fig. 11-157D. Resistance in ohms versus change in temperature. (Courtesy, General Electric Co.)

Magnetic devices like transformers, reactors, and relays generate voltage-transient peaks far in excess of the normal input, and are frequently the cause of rectifier failure. As a result, voltage suppressors or clippers are used to increase the reliability of rectifier units, without affecting the circuit operation.

Transient suppressors are selenium devices with *nonlinear* reverse characteristics. The thick dielectric barriers, large area dissipation, and the polycrystalline structure make them capable of handling transient peak voltages of two orders of magnitude or more, above their steady-state rating, and for in-rush currents far in excess of normal densities. The characteristics for a typical *nonpolarized* suppressor are shown in Fig. 11-158A. The similarity of wave shape on both halves of the input circuit will be noted. The characteristic for a *polarized* suppressor are shown in Fig. 11-158B. Nonpolarized



(a) *Nonpolarized.*



(b) *Polarized.*

Fig. 11-158. Typical reverse characteristics for surge suppressors.

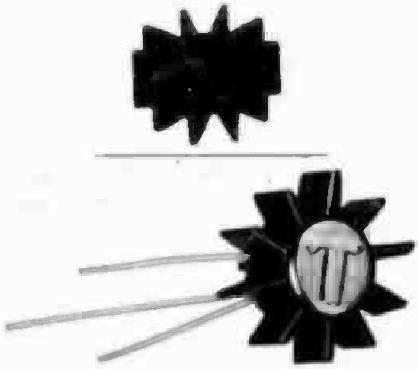
suppressors may be used in either ac or dc applications. Polarized suppressors can be used only in dc circuits, with the suppressor in a normally nonconducting state, but offering a low-impedance discharge path for a collapsing field. Nonpolarized suppressors will clip transient voltages to approximately twice the peak-inverse voltage ratings. Nonpolarized suppressors with low impedance in the nonblocking direction will clip to very low voltage values.

The most common cause of transients is when a switch is opened in the load side of a rectifier circuit at nearly the maximum instantaneous current, generating a high voltage. The peak of the voltage spike can be *hundreds of times the peak of the steady-state voltage.* In this case, the suppressor is connected across the inductance. In rectifier circuits where only the primary circuit of the transformer is broken, the suppressor is connected across the transformer primary. The recovery time for the suppressor is within microseconds. Typical devices of this nature are the Klipvolt, manufactured by Sarkes-Tarzian Inc., and Klip-Sels, manufactured by the International Rectifier Corp. The use of such devices is discussed in Section 21.

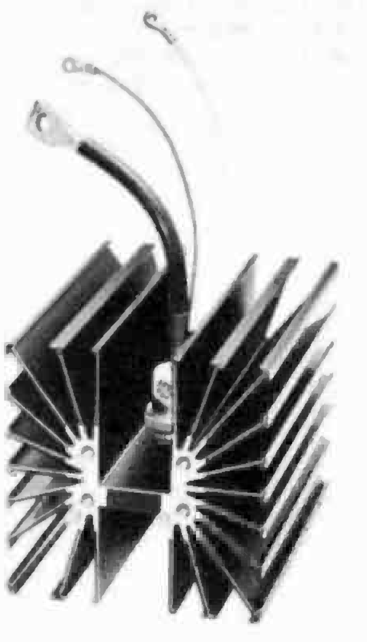
11.159 Describe the different type heat sinks used with semiconductor devices.—Two types of heat sinks used with semiconductor devices are shown in Fig. 11-159. The units in part (a) are used with small low-power transistors, and are smaller in diameter than a dime. The large unit shown in part (b) is used with high-current silicon diodes where considerable heat must be dissipated. The stud of the diode unit is screwed into the center piece of the heat sink. Heat-sink design is discussed in Questions 21.125 to 21.127.

11.160 What are steering diodes?—Diodes connected in a circuit for the purpose of permitting circuit operation only when the actuating signal is of a given polarity. In Fig. 11-160 is shown a steering diode connected in the input of a transistor amplifier to allow the circuit to respond only when a positive-going signal is applied to the input. If negative-signal operation is required, the diode polarity is reversed.

11.161 Describe a photofet or phototransistor.—A photofet is an epitaxial semiconductor photocell, using a



(a) Small heat sinks used with diodes and transistors. Their diameter is less than a dime.



(b) Large heat sink for use with heavy-current diode rectifiers. The stud of the diode is screwed into the center fin of the sink.

Fig. 11-159. Conduction-type heat sinks used for cooling diodes and transistors. (Courtesy, Wakefield Engineering Co.)

field-effect transistor, manufactured by Crystalonics Inc. Incident light is focused on the gate element of the FET through a lens, an integral part of the device. When light strikes the gate area, the electrons on the outer valence of the impurity atom are excited and flow out of the junction through the gate.

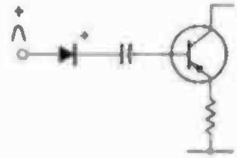


Fig. 11-160. Steering diode connected in the input of a device for permitting its operation only from a positive-going signal. For negative-signal operation the polarity of the diode is reversed.

To state it simply, a given amount of incident light generates a proportional amount of gate current, known as a λ I_g. Thus, the gate junction acts as a current generator and will continue to deliver the same amount of current, regardless of the load, for a fixed source of illumination. It is claimed for such devices that the sensitivity is greater than 10:1 over conventional bipolar photosensitive devices, with 4:1 increases in bandwidth. Such devices are used for high-speed switching, logic, and for many other various control functions.

A circuit suitable for a light-controlled attenuator is shown in Fig. 11-161. The circuitry is designed to take advantage of the photofet's ability to function as a variable resistor. Here, the drain-to-source resistance is a function of the light input obtained from a small incandescent lamp. The bias voltage is adjusted for the desired drain-to-source resistance under quiescent conditions. The spectral response extends from 4000 to 12,000 angstroms, with maximum sensitivity occurring at 9500 angstroms. The recommended light source is a tungsten lamp operated at 2000 degrees Kelvin color temperature.

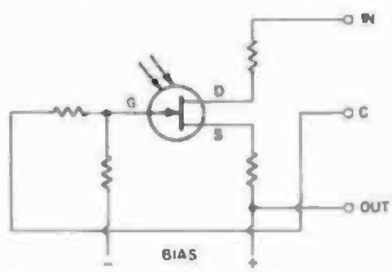


Fig. 11-161. Control circuit for a remote light-operated attenuator using a photofet manufactured by Crystalonics Inc. To distinguish the photofet from a conventional transistor, two arrows are placed above the gate.

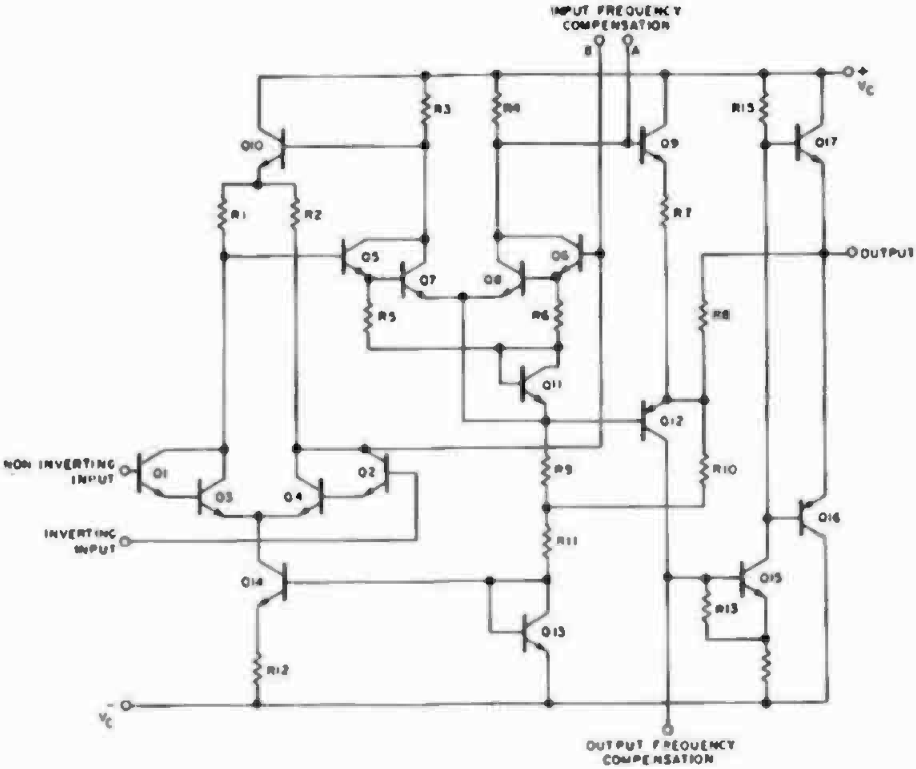


Fig. 11-162. Typical schematic circuit for a monolithic integrated amplifier similar to that used in operational amplifiers.

Because of the light sensitivity of certain semiconductor materials, it is necessary to encase the standard transistor in an opaque container. If this is not done, it is possible to pick up hum from lighting fixtures. This effect has been observed when painted plastic envelopes are used and the paint has become scratched. (See Question 19.169.)

11.162 Describe the basic concept of an integrated circuit (IC).—It is a combination of active and passive semiconductor components fabricated on a single semiconductor chip. The internal components may consist of several transistors, diodes, and resistors. External reactive components can be connected to the internal circuits to form amplifiers, oscillators, and a multiplicity of devices. Negative feedback is generally induced for stability. Integrated circuits are employed in operational amplifiers, although they may also be used for other purposes.

A typical schematic circuit for a monolithic integrated circuit similar to that used in operational amplifiers is given in Fig. 11-162. Operational amplifiers are discussed in Questions 12.196 and 12.261.

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Section 12

Audio Amplifiers

The subject of audio amplifiers seems to be endless because of the many designs and functional aspects. In this Section both theoretical and practical amplifier circuits are given for semiconductor and vacuum tube types, with simplified design equations. Resistance and transformer coupled amplifiers with different types of negative feedback are explained. The latter are the outgrowth of the research of H. S. Black, I. G. Wilson, and others.

Of special interest to the audio engineer will be the discussion of semiconductor ac and dc operational amplifiers suitable for adoption to sound mixer console design. Covered are such items as compressor amplifiers, graphic equalizers, high- and low-frequency equalizers, bandpass filters, oscillators, mixer and line amplifiers, and many others. Also discussed are field-effect transistors (FET's) and the possibility of their direct use in place of vacuum tubes.

12.1 What is a resistance-coupled amplifier?—One of the most widely used methods of coupling vacuum tubes is by means of resistance coupling. Resistance-coupled amplifiers are characterized by their simplicity of design, construction, wide frequency range, and their low cost of manufacture. In addition, they are less susceptible to hum and noise pickup from surrounding electrical fields and are especially suited for use with high-mu triode and pentode tubes. A typical two-stage resistance-capacitance coupled amplifier is shown in Fig. 12-1A.

It will be noted that a load resistor R_p is connected in the plate circuit of tube V1 and is used for developing the signal voltage for the following stage. The developed signal is conducted from the plate of V1 to the control grid of V2 through the coupling capacitor C. A grid resistor R_g returns the control grid of V2 to ground. The sole purpose of the coupling capacitor C is to isolate the control grid of V2 from the high positive voltage on the plate of V1.

When a tube in a static or quiescent state (no signal) is connected to the proper grid, cathode, and plate resistors (screen resistor if a pentode is used), and supplied with plate voltage, there is a steady voltage drop across the plate-load resistor. If an instantaneous positive-going signal is applied to the control grid, the plate current will increase. This causes the steady plate voltage to increase in a negative direction (voltage drops at the plate) and generate a negative signal voltage at the plate. Now, if an instantaneous negative-going signal is applied to the control grid, the plate current decreases and the voltage at the plate increases in a positive direction, thus generating a positive signal.

When applying a sine-wave signal to the control grid, the voltage at the plate will rise and fall in accordance with the signal at the control grid, but it will be reversed 180 degrees. Therefore, if the tube is not driven into overload the signal at the plate will be an inverted replica of the signal voltage at the control grid, but of greater magnitude depending on the amplification factor of the tube. The quiescent plate current is selected for a value in the most *linear portion* of the plate-current characteristic. (See Questions 11.52 and 12.101.

As a rule, resistance coupled amplifiers are designed to use self-bias; that is, they have a resistance R_k connected in the cathode circuit through which

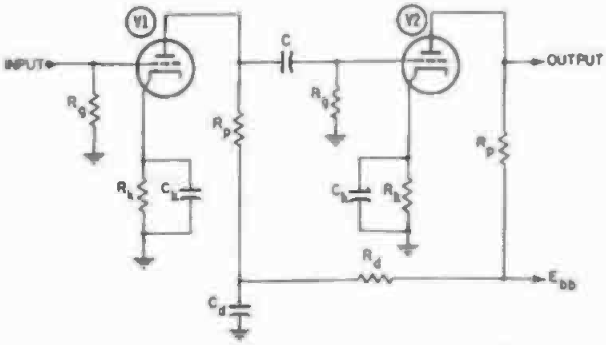


Fig. 12-1A. A two-stage, resistance-capacitance coupled triode amplifier.

the plate and screen-grid (pentode tube) currents pass. This flow of current through resistor R_k to the plate causes the cathode to become positive, with respect to the ground, by the amount of the voltage drop across R_k.

The control grid of V1 is returned to ground through the grid resistance R_g which places the grid at ground potential. This makes the control grid negative with respect to the cathode by the amount of voltage drop across the cathode resistor R_k. This same reasoning is also true for the control grid of V2. Large values of capacitance are connected across the cathode resistors to provide a low impedance path for the signal voltage in the cathode circuit to ground.

In multistage amplifiers, decoupling circuits consisting of capacitors C_d and resistor R_d are connected in the plate circuits of each amplifier stage to prevent common coupling of the tubes through the power supply.

A two-stage resistance-capacitance coupled amplifier using pentodes is shown in Fig. 12-1B. The circuit is similar in design to Fig. 12-1A, except for the addition of the screen-grid dropping resistor R_{sg} and the screen-grid bypass capacitor C_{sg}.

12.2 What are the symbols used for the design of resistance-capacitance coupled amplifiers?

C (C_c)	Coupling capacitor between stages
C_k	Cathode bypass capacitor
C_{sg} (C_{g2}) (C_{sc})	Screen-grid bypass capacitor
E_{bb}	Supply voltage
E_p	Actual voltage at the plate
E_{sg}	Actual voltage at the screen grid
I_p (I_b)	Plate current
I_{sg} (I_{g2})	Screen grid current
R_k	Cathode self-biasing resistor
R_{sg} (R_{g2}) (R_{sc})	Screen-grid dropping resistor
R_g	Grid resistor
R_p (R_L) (R_L)	Plate load resistor
E_{sig} (E_{sg})	Signal voltage at the control grid
V_g	Voltage gain of the stage
E_o	Output signal voltage

Variations of these symbols are given in parentheses. (See Question 11.34.)

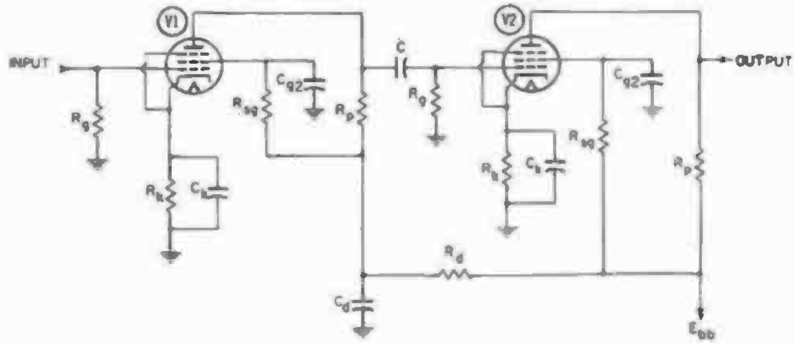


Fig. 12-1B. A two-stage, resistance-capacitance coupled pentode amplifier.

12.3 What are the gain-frequency characteristics of a resistance-capacitance coupled amplifier?—Generally speaking, resistance-capacitance coupled amplifiers have three sets of frequency characteristics. They are: low, medium, and high frequency. With proper design the frequency response may be made quite uniform over a range from direct current to several megahertz. Resistance-coupled amplifiers for audio use are designed to cover a frequency range from 10 to 20,000 Hz and above. (See Question 12.6.)

12.4 What are the factors governing the gain of a resistance-coupled amplifier?—To obtain a large voltage gain in a resistance-coupled amplifier stage, the plate-load resistor R_p must have as large a value as is practical. The higher the value of the plate-load resistance, the greater will be the dc voltage drop across it and the lower will be the actual voltage at the plate of the tube. Therefore, there is a practical limit to the value of the load resistance.

The dc voltage drop across the plate-load resistor must be subtracted from the supply voltage E_{bb} to arrive at the actual value of voltage at the plate of the tube. If the value of the load resistance is high, only a small portion of the supply voltage is available at the plate. This may be too low for proper operation.

Fig. 12-4 illustrates how the supply voltage is distributed in a triode resistance-coupled amplifier. It will be noted that for a supply voltage of 300 volts and a plate-load resistance of 30,000 ohms, 180 volts of the supply voltage is lost in the dc voltage drop across the load resistance and only 117 volts appear between the cathode and the plate, with 3-volts drop across the cathode resistor R_k.

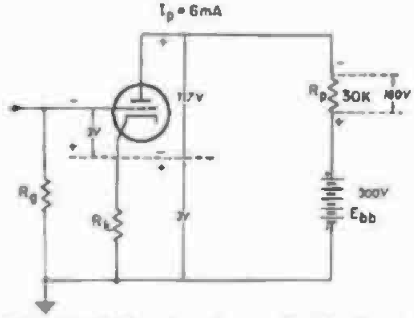


Fig. 12-4. The dc voltage distribution in a resistance-coupled amplifier.

From this illustration it may be readily seen that with a 300-volt supply, increasing the plate-load resistance to a high value results in an excessive voltage drop across the load resistance, and that for a very high value of resistance, the supply voltage would have to be increased to a value out of proportion to the small amount of increased gain achieved by increasing the plate-load resistance.

12.5 What is the relationship between gain and the value of plate-load resistance?—Under the best conditions, only about 80 percent of the tube amplification can be attained. This relationship is shown in Fig. 12-5. It will be noted that, after a certain value of resistance has been reached, the gain increases very slowly and never quite reaches the full amplification factor of the tube.

12.6 What is the variation in the frequency characteristics for different values of plate-load resistance?—Typical gain-frequency characteristic curves for a resistance-coupled amplifier using different values of plate load are shown in Fig. 12-6. As will be seen, the frequency response falls off at both the high and low frequency ends as the plate-load resistance is increased. A

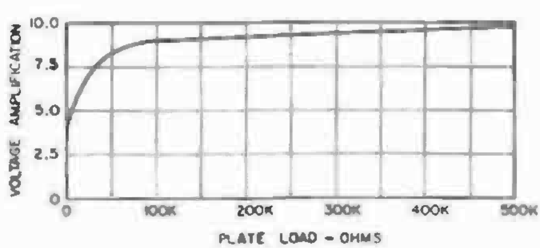


Fig. 12-5. Voltage amplification versus load resistance for a triode having a plate resistance of 10,000 ohms, an amplification factor of 10, and a transconductance of 1000 micromhos.

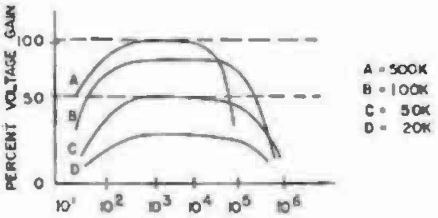


Fig. 12-6. Variation in voltage gain of an RC coupled amplifier for different values of R_p.

wide frequency response can only be obtained with low values of plate load resistance and at the expense of gain.

12.7 What factors affect the gain at the lower frequencies in a resistance-coupled amplifier?—Referring to Fig. 12-1A, a loss of gain at the low frequencies is caused by the high reactance of the coupling capacitor C. The equivalent circuit for a single stage, triode, resistance-coupled amplifier is shown in Fig. 12-7. At the low frequencies, the signal voltage across R_p appears across the series combination of capacitor C and the grid resistor R_g, which is applied to the input of the following stage (μE_g represents the vacuum tube.)

The lower the frequency, the higher becomes the reactance of the coupling capacitor C and more and more of the signal voltage developed across R_p appears across the capacitor C. Therefore, less voltage is applied to the grid of the following stage because of the voltage divider action of R_g and C. The best low-frequency response is obtained when the coupling capacitor C is large in value, so that its reactance is negligible at the lowest frequency to be amplified.

12.8 What are the gain-frequency characteristics at the midfrequencies in a resistance-coupled amplifier?—At the midfreqnencies little or no frequency discrimination is noted because the reactance of the coupling capacitor C is

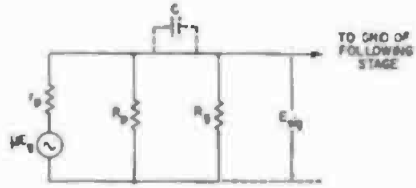


Fig. 12-8. Equivalent circuit for a triode RC coupled amplifier at the mid-frequencies.

small. In Fig. 12-8 (an equivalent circuit) the coupling capacitor has been omitted as its reactance is negligible and may be considered to be a short circuit. Under these conditions, all the voltage (E_{sig}) developed across the plate-load resistor R_p is delivered to the following stage, resulting in a uniform frequency response.

12.9 What are the gain-frequency characteristics at the high frequencies in a resistance-coupled amplifier?—The high-frequency response falls off because of the interelectrode capacitance of the tube and the distributed capacitance of the tube socket and associated wiring. These capacities are considered to be in shunt with the plate and control-grid circuits and act as a low-impedance path to ground at the higher frequencies. (See Fig. 12-9.) As the frequency increases, the reactance of the shunt capacitance C_s becomes less and less, eventually reaching a value equal to that of the load resistance R_p. As a result, the impedance is lowered, a lower voltage is developed across the load resistor R_p, and thus less voltage appears across the input of the following stage. Therefore, the gain falls off as the frequency is increased. Because the reactance of the coupling capacitor is small at the high frequencies, it may be considered to be out of the circuit.

12.10 What is the equation for calculating the amplification of a resistance-coupled amplifier stage at the mid-

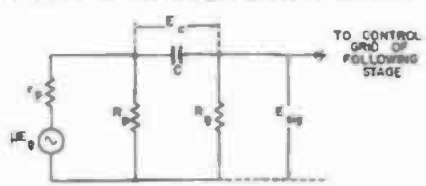


Fig. 12-7. Equivalent circuit for a triode RC coupled amplifier at the low frequencies.

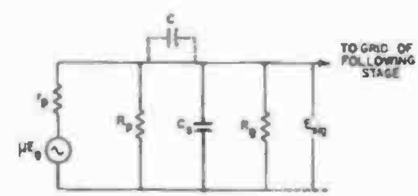


Fig. 12-9. Equivalent circuit for a triode RC coupled amplifier at the high frequencies.

frequencies?—The general equation for a triode tube is:

$$A = \frac{R_p}{r_p + R_p}\mu$$

where,

R_p is the plate load resistor,
r_p is the plate resistance of the tube,
μ is the amplification factor of the tube.

The foregoing equation will give the gain of the tube when the plate-load resistor R_p is not shunted by the grid resistance R_g and coupling capacitor C of a second stage. If the plate resistor R_p is shunted by the components of a second stage, this must be taken into consideration when computing the gain.

The equation is then written:

$$A = \frac{R_{eq}}{r_p + R_{eq}}\mu$$

where,

r_p is the plate resistance of the tube,
R_{eq} is the equivalent resistance of the second stage grid resistor R_g, and the coupling capacitor C, in parallel with the plate load resistor R_p.

A practical example using the above equation would be: Assume a triode is to be operated using a plate-load resistor R_p of 47,000 ohms. The plate resistance of the tube is 7700 ohms and the mu is 17. If the grid resistor R_g of the second stage is several times the value of the plate load resistor R_p, only the value of R_p need be considered. The reactance of the coupling capacitor C, is generally small at the midfrequencies and for practical purposes may be considered to be a short circuit. Therefore, the amplification will be:

$$A = \frac{47,000}{7700 + 47,000} \times 17$$
$$= \frac{799,000}{54,700}$$
$$= 14.62$$

12.11 What is the equation for calculating the amplification of a resistance-coupled amplifier stage at the low frequencies?—The gain of a resistance-coupled amplifier at the low frequencies is more difficult to compute because the reactance of the coupling capacitor to the second stage must be included. The amplification at the low frequencies for a triode tube may be computed:

$$A = \frac{1}{\sqrt{1 + (X_c/R_{eq})^2}}$$

where,

X_c is the capacitive reactance of the coupling capacitor C at the lowest frequency to be passed by the amplifier,
R_{eq} is the equivalent resistance of the plate load resistor R_p in parallel with the reactance of the coupling capacitor in series with the grid resistor R_g of the second stage.

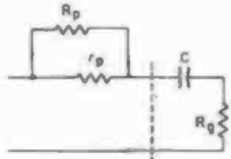


Fig. 12-11. Equivalent circuit for a triode RC coupled amplifier at the low frequencies.

An equivalent diagram for the foregoing equation is shown in Fig. 12-11.

The reactance of the coupling capacitor C increases with a decrease of frequency and is in series with the grid resistor R_g of the second stage. The control grid of the second stage is fed from the junction of these two circuit elements; therefore, less signal voltage is applied to the secoud stage control grid as the frequency is lowered, because of the voltage-divider action of these two circuit elements.

At a frequency where the capacitive reactance of capacitor C is equal to the value of the grid resistor R_g, the signal voltage at the control grid of the second stage will be reduced 3 dB. This is called the half-power point and is taken to be the lower frequency limit of the amplifier. The response at the lower frequencies may be extended by the methods described in Questions 12.21 and 12.37.

12.12 What is the equation for calculating the amplification of a resistance-coupled amplifier stage at the high frequencies?—For a triode the amplification is:

$$A = \frac{1}{\sqrt{1 + (R_{eq}/X_c)^2}}$$

where,

R_{eq} is the equivalent resistance as described in Question 12.10,
X_c is the capacitive reactance of the stray capacitance due to wiring, sockets, and the internal capacitance of the tube.

At the high frequencies the reactance of the coupling capacitor C, is quite small and may be neglected. At a frequency where the reactance of the stray capacitance equals one-third the value of R_{eq}, the gain will drop 3 dB. This is called the upper half-power point and is taken as the upper frequency limit of the amplifier.

12.13 What is the equation for calculating the amplification of a resistance-coupled pentode amplifier stage?—Because of the high plate resistance of a pentode tube (generally of the order of 1 megohm), the value of the plate load resistor R_p is not important. The formula then becomes:

$$A = G_m R_{eq}$$

where,

G_m is the transconductance of the tube in micromhos,

R_{eq} is the equivalent load resistance.

The transconductance is equal to mu/r_p.

12.14 What type tubes are recommended for resistance-coupled amplifiers?—Pentodes or high-mu triodes. For high-frequency amplification, the pentode is preferred because of its lower internal-electrode capacitance. This keeps the shunt capacitive reactance high in comparison to the load resistance R_p. As a rule, the load resistance is made low enough that it is lower than the shunting impedance of the distributed capacitance caused by the wiring, socket, and internal capacitance of the tube.

Lowering the load resistance R_p also decreases the amplification of all frequencies passed by the amplifier, but such design is necessary if the amplifier is to cover a wide frequency band.

The data supplied by the tube manufacturers cover the audio frequency range from 20 to 20,000 Hz. If the low-frequency end is to be extended, the coupling capacitor must be increased in size as discussed in Question 12.21. Also, compensating circuits are generally added as described in Question 12.37.

12.15 How are the circuit constants for resistance-coupled amplifiers selected?—The circuit constants may be calculated individually; however, charts are available from the tube manufacturer that provide all the necessary data except for very special cases. Typical charts are shown in Fig. 12-15. Values for the circuit elements are selected on the basis of the plate supply voltage E_{bb}, plate-load resistance R_p, and the grid resistance R_g, remembering that the higher the plate-load resistance, the greater the loss at the high frequencies. Referring to Fig 12-15, chart 2, it will be noted that under a given plate-load resistance are tabulated the value of the cathode resistance R_k, the screen-grid dropping resistor R_{c2} (or R_{sg}), the coupling capacitor C, and the screen grid and cathode bypass capacitors C_{c2} (or C_{sg}) and C_k.

The values of capacity given are the minimum values for a given set of conditions. The circuit constants are predicated on a given signal voltage E_o at the plate for a given value of harmonic distortion. The amplification or voltage gain V_g given in the charts may be converted to decibels by the equation:

$$dB = 20 \operatorname{Log}_{10} V_g$$

where,

V_g is the voltage gain stated in the charts.

As an example, a given tube has a voltage gain of 14. What is the gain in decibels?

$$\begin{aligned} dB &= 20 \operatorname{Log}_{10} 14 \\ &= 20 \times 1.46 \\ &= 22.92 \text{ dB.} \end{aligned}$$

12.16 Explain how the circuit constants given in a resistance-coupled amplifier design chart are determined.—They are based on a given frequency response, distortion, gain, and signal output voltage. The frequency response on which the circuit constants are predicated is shown in Fig. 12-16. The frequency f_2 is that value at which the high-frequency response begins to fall off. The frequency f_1 is that value at which the low-frequency response drops below a satisfactory value, as discussed in the questions to follow. For most of the types shown, the data pertain to the use of a cathode bias resistor where feasible, a series screen-dropping resistor where applicable, and offers several advantages over fixed voltage operation. These advantages are:

Operation over a wide range of plate supply voltages without appreciable change in gain;

The effects of possible tube differences are minimized;

The low frequency at which the amplifier cuts off is easily changed;

The tendency toward motorboating is minimized.

Tube	E_{bb}	R_p	R_g	R_{g2}	R_k	C_{g2}	C_k	C	E_o	V.G.
6AQ6, 6AQ7-GT, 6AT6, 6Q7, 6Q7-G, 6Q7-GT, 6SL7-GT•, 6SZ7, 6T7-G, 6T8, 12AT6, 12Q7-GT, 12SL7-GT•, 19T8	90	0.1	0.1	–	4200	–	2.5	0.025	5.4	22●
			0.22	–	4600	–	2.2	0.014	7.5	27●
			0.47	–	4800	–	2.0	0.0065	9.1	30●
		0.22	0.22	–	7000	–	1.5	0.013	7.3	30■
			0.47	–	7800	–	1.3	0.007	10	34■
			1.0	–	8100	–	1.1	0.0035	12	37★
		0.47	0.47	–	12000	–	0.83	0.006	10	36■
			1.0	–	14000	–	0.7	0.0035	14	39★
			2.2	–	15000	–	0.6	0.002	16	41★
	180	0.1	0.1	–	1900	–	3.6	0.027	19	30★
			0.22	–	2200	–	3.1	0.014	25	35
			0.47	–	2500	–	2.8	0.0065	32	37
		0.22	0.22	–	3400	–	2.2	0.014	24	38
			0.47	–	4100	–	1.7	0.0065	34	42
			1.0	–	4600	–	1.5	0.0035	38	44
		0.47	0.47	–	6600	–	1.1	0.0065	29	44
			1.0	–	8100	–	0.9	0.0035	38	46
			2.2	–	9100	–	0.8	0.002	43	47
	300	0.1	0.1	–	1500	–	4.4	0.027	40	34
			0.22	–	1800	–	3.6	0.014	54	38
			0.47	–	2100	–	3.0	0.0065	63	41
		0.22	0.22	–	2600	–	2.5	0.013	51	42
			0.47	–	3200	–	1.9	0.0065	65	46
			0.1	–	3700	–	1.6	0.0035	77	48
		0.47	0.47	–	5200	–	1.2	0.006	61	48
			1.0	–	6300	–	1.0	0.0035	74	50
			2.2	–	7200	–	0.9	0.002	85	51
6AU6, 6SH7, 12AU6, 12SH7	90	0.1	0.1	0.07	1800	0.11	9.0	0.021	25	52
			0.22	0.09	2100	0.1	8.2	0.012	32	72
			0.47	0.096	2100	0.1	8.0	0.0065	37	88
		0.22	0.22	0.25	3100	0.08	6.2	0.009	25	72
			0.47	0.26	3200	0.078	5.8	0.0055	32	99
			1.0	0.35	3700	0.085	5.1	0.003	34	125
		0.47	0.47	0.75	6300	0.042	3.4	0.0035	27	102
			1.0	0.75	6500	0.042	3.3	0.0027	32	126
			2.2	0.8	6700	0.04	3.2	0.0018	36	152
	180	0.1	0.1	0.12	800	0.15	14.1	0.021	57	74
			0.22	0.15	900	0.126	14.0	0.012	82	116
			0.47	0.19	1000	0.1	12.5	0.006	81	141
		0.22	0.22	0.38	1500	0.09	9.6	0.009	59	130
			0.47	0.43	1700	0.08	8.7	0.005	67	171
			1.0	0.6	1900	0.066	8.1	0.003	71	200
		0.47	0.47	0.9	3100	0.06	5.7	0.0045	54	172
			1.0	1.0	3400	0.05	5.4	0.0028	65	232
			2.2	1.1	3600	0.04	3.6	0.0019	74	272
	300	0.1	0.1	0.2	500	0.13	18.0	0.019	76	109
			0.22	0.24	600	0.11	16.4	0.011	103	145
			0.47	0.26	700	0.11	15.5	0.006	129	168
		0.22	0.22	0.42	1000	0.1	12.4	0.009	92	164
			0.47	0.5	1000	0.098	12.0	0.007	108	230
			1.0	0.55	1100	0.09	11.0	0.003	122	262
		0.47	0.47	1.0	1800	0.075	8.0	0.0045	94	248
			1.0	1.1	1900	0.065	7.6	0.0028	105	318
			2.2	1.2	2100	0.06	7.3	0.0018	122	371

●– At 2 volts (rms) output. ■ At 3 volts (rms) output. ★ At 4 volts (rms) outpt.
• One triode unit.

Fig. 12-15. A typical resistance-coupled amplifier data chart. (Courtesy, Radio Corporation of America)

The output peak voltage E_o is the voltage available across the grid resistance R_g of the following stage at any frequency within the flat portion of the frequency response, and where the input signal is of such magnitude as to swing the control grid to a point where grid current just starts to flow.

12.17 How critical is the supply voltage for a resistance-coupled amplifier?—For supply voltages differing by 50 percent from those listed in the

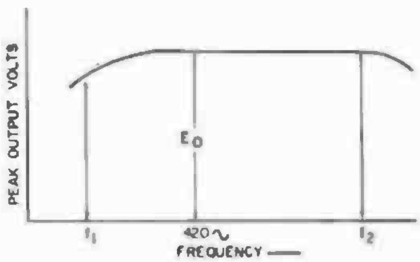


Fig. 12-16. Frequency response for the design of RC amplifiers. (Courtesy, Radio Corporation of America.)

charts of Fig. 12-15, the voltage gain is approximately the same.

12.18 How critical are the values of the circuit components in a resistance-coupled amplifier?—The values of the circuit components in the charts are approximately correct for changes up to 50 percent of the stated supply voltage. Ten-percent tolerance capacitors and resistors are quite satisfactory to be used.

12.19 What is the frequency response when the data in the tube charts are used with a directly heated triode amplifier?—Capacitors C and C_k have been chosen to give a signal output voltage equal to 0.8 E_o for a frequency f_1 of 100 Hz (Fig. 12-16). For any other value of f_1 multiply C and C_k by $100/f_1$. The values given for C_k are for an amplifier using direct current on the heaters. When alternating current is used, it will be necessary to increase the value of C_k to minimize hum disturbances. The voltage at f_1 of "n" stages equals 0.8^n E_o, where E_o is the peak signal voltage at the final stage for any value of R_p.

12.20 What is the frequency response when the data in the tube charts are used with a filament-type pentode tube?—Capacitors C and C_{sg} have been chosen to give an output signal voltage equal to 0.8 E_o for a frequency f_1 of 100 Hz (Fig. 12-16). For any other value of f_1, multiply values of C and C_{sg} by $100/f_1$. The voltage output at F_1 for "n" like stages equals 0.8^n E_o, where E_o is the peak signal output voltage of the final stage. For an amplifier of typical construction and for R_p values of 100,000, 250,000, and 500,000 ohms, approximate values of f_2 are 20,000, 10,000, and 5,000 Hz, respectively. The values of the input coupling capacitor C in microfarads and of the grid resistor R_g in megohms should be such that their product lies between 0.02 and 0.10.

12.21 What is the frequency response when the data in the tube charts are used with a heater-type pentode?—Capacitors C, C_k, and C_{sg} have been chosen to give an output voltage equal to 0.7 E_o for a frequency f_1 of 100 Hz (Fig. 12-16). For any other values of f_1, multiply values of C, C_k, and C_{sg} by $100/f_1$. In the cases of capacitor C_k, the values shown in the charts are for an amplifier with dc heater excitation. When ac is used, depending on the character of the associated circuits, voltage gain, and the value of f_1, it may be necessary to increase the value of C_k to minimize hum disturbances. The voltage output at f_1 for "n" like stages equals 0.7^n E_o, where E_o is the peak output signal voltage at the final stage. For an amplifier of typical construction and for R_p values of 0.1, 0.25, and 0.5 megohm, approximate values of f_2 are 20,000 10,000, and 5000 Hz, respectively.

12.22 Is it necessary to bypass a self-bias resistor?—No, not unless the maximum gain is required from the tube. When the cathode or self-bias resistance is left unbypassed, degeneration (negative-current feedback) is developed by the flow of the audio signal through the bias resistor because of the ac component of the plate current flowing through the bias resistor.

The flow of signal current causes a voltage drop across the bias resistor, which increases the normal bias voltage on the control grid. This increase of bias voltage reduces signal voltage at the control grid and reduces the stage gain from 4 to 6 dB. This requires the input signal amplitude be increased to produce a given output signal level at the plate circuit by the factor:

$$E_{sig} = \frac{A \times R_k}{R_{eq}}$$

where,

- A is the amplification factor of the tube with the cathode resistor bypassed,
- R_{eq} is the combined resistance of the plate-load resistor, R_p and the grid resistor R_g of the following stage,
- R_k is the cathode resistor.

Leaving the cathode resistor unbypassed results in a more uniform frequency response with lower distortion for a given set of conditions, but with higher plate impedance. Unbypassed cathode circuits are more susceptible to hum pickup from ac heater circuits.

The foregoing statements apply to single-stage amplifiers only. For push-pull amplifiers see Question 12.106.

12.23 What is the relationship between the coupling capacitor C and the grid resistor R_g?—When the reactance of the coupling capacitor C at the lower frequencies becomes equal to the grid resistor R_g, the frequency response will be down 3 dB, compared to a reference frequency of 420 Hz.

When designing a capacitive-coupled circuit, it is useful to be able to predict what proportion of the voltage across the plate resistor R_p will appear across the grid resistor R_g. This may be determined by the use of the nomograph in Fig. 12-23. A straight edge is laid from the load resistance at the left to the value of the capacitive reactance at the right. The loss in voltage gain, relative to 100-percent gain, is read from the percentage scale at the center. The loss in dB for a range of 0.5 to 6 dB may be read at the left of the center scale. The capacitive reactance is that of the coupling capacitor at the highest and lowest frequency of interest.

12.24 How does the cathode capacitor C_k affect the frequency response?—Normally the cathode capacitor size is selected for a value that will have a reactance of one-tenth the resistance of the cathode resistor. Practical sizes are 25 to 50 μF. The purpose of the cathode capacitor is to provide a low impedance path for the signal voltage to ground and remove it from the cathode resistor so far as possible. As the size of the cathode-bypass capacitor is reduced, degeneration (negative-current feedback) is increased and, if this capacitor is made small enough, the gain begins to fall at the lower frequencies and the frequency characteristic starts to rise at the higher frequencies. A small capacitor across a cathode resistor is often used to increase the high-frequency response. This subject is discussed in Question 6.96.

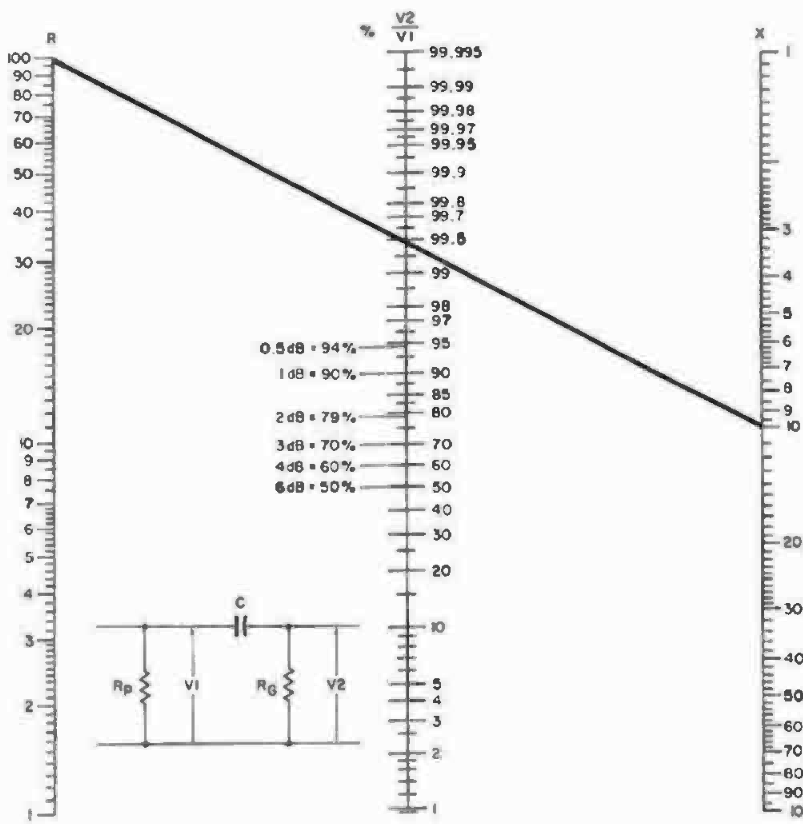


Fig. 12-23. Nomograph for determining the loss in gain that occurs through a capacitive-coupled circuit.

12.25 How may the size of the coupling, screen, and cathode capacitors be approximated?—By the use of the equations:

$$C = \frac{1.6 \times 10^6}{f \times R_g} = \mu F$$

$$C_k = \frac{1.6 \times 10^6}{f \times R_k} = \mu F$$

$$C_{sg} = \frac{1.6 \times 10^6}{f \times R_{sg}} = \mu F$$

$$X_c = \frac{10^6}{2\pi f}$$

where,

C is the coupling capacitor,
C_k is the cathode-bypass capacitor,
C_{sg} is the screen-bypass capacitor,
f is the lowest frequency to be amplified,
R_g is the grid resistor,
R_k is the cathode resistor,
R_{sg} is the screen grid dropping resistor,
X_c is the capacitive reactance.

This statement does not take into consideration effects of phase shift.

12.26 What effect does a volume control have on the frequency response of an amplifier?—It adds to the impedance seen by the input of the following tube reducing the high-frequency response; however, generally it is of no great consequence. Volume controls should not be included in a negative-feedback loop of an amplifier because of its internal capacitance. (See Question 12.191.)

12.27 What factors affect the distortion characteristics of a resistance-coupled amplifier?—They are the same as for any other type amplifier and are a function of the signal amplitude and operating parameters. Negative feedback may be used over a number of stages to obtain the most desirable distortion characteristics.

12.28 How is self-bias achieved in an amplifier?—By connecting a resistor in series with the cathode element to ground. The flow of plate current (and screen grid in the case of a pentode) through this resistor causes a voltage drop across the resistor making the cathode positive with respect to ground. The control grid is connected to ground making the grid negative with respect to the upper end of the cathode. Thus, the control grid is made negative by the amount of voltage drop across the cathode resistor. The value of the cathode resistor, R_k, may be calculated:

$$R_k = \frac{E_g}{I_k}$$

where,

I_k is the total current through the cathode resistor,
E_g is the required grid-bias voltage.

For pentodes, the screen-grid current is added to that drawn by the plate. (The screen-grid current also flows through the cathode resistor.)

The cathode capacitor may be eliminated by use of the circuit shown in Fig. 12-28 in which the cathode element is connected to the junction of a voltage divider circuit formed by a resistor R_d and the cathode resistor R_k.

This circuit places the cathode close to ground and eliminates the need for the usual heavy bypass capacitor. This circuit is quite useful in low-level high-gain amplifier stages in reducing hum. As an example: a given tube requires a bias voltage of 1.2 volts. Connecting a 68,000 ohm resistor from a source of 200 volts dc, to a cathode resistor of 390 ohms places the cathode element only 390 ohms above ground.

Because the cathode resistor is not bypassed, degeneration will take place across the cathode resistor; however, the benefits gained by the reduction in hum in this stage offsets the loss of gain suffered from the degeneration.

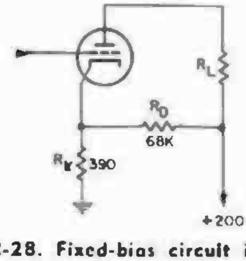


Fig. 12-28. Fixed-bias circuit in a low-level, high-gain amplifier stage.

12.29 What is a grid-leak resistor?—A high-value resistor connected from the control grid of a vacuum tube to ground or the negative side of the circuit. Its purpose is to permit excessive electrons on the control grid to leak off to ground and thus prevent the tube from becoming blocked due to an excessive negative charge on the grid. Grid-leak resistors are used in amplifiers, oscillators, radio detectors (demodulators), and similar devices.

12.30 What is a stopping capacitor?—It is another name for a coupling capacitor used between the stages of a resistance- or impedance-coupled amplifier.

12.31 What is the effect of a leaky coupling capacitor?—Referring to Fig. 12-1A, if the coupling capacitor C is leaky, the positive potential at the plate of the tube V1 will be applied to the control grid of tube V2. The positive potential will offset the negative bias on the control grid of V2 causing the tube to draw grid current and an excessive amount of plate current which may cause the tube to be permanently damaged. In addition, because the tube will not be operating on the linear portion of its characteristic curve, excessive distortion will result.

12.32 What are the limiting factors for determining the maximum resistance that may be inserted in the control-grid circuit of a vacuum tube?—The space current. Because of the current flow, electrons collect on the control grid forming a negative charge. This charge has no way of leaking off except through the resistance between the grid and ground. If the grid resistance is in the order of one-fourth to one-half megohm, a considerable bias voltage is created, ranging from 0.25 to 2.0 volts. The effect of the space charge relative to the grid resistance is often referred to as contact potential because of its similarity to the condition resulting when dissimilar metals placed in contact generate a current.

The maximum value of resistance that may be used with a particular tube may be obtained from the manufacturer's sheet. Higher values of grid-circuit resistance may be used with cathode bias (self-bias) than fixed bias. The reason for this difference is that in a self-bias or automatic-bias circuit, the bias voltage increases with an increase of plate current, thus preventing the creeping of plate current, because of grid emission and gas, between the grid and cathode elements. (See Questions 11.44 and 11.62.)

12.33 What is the overall frequency response of two resistance-coupled amplifier stages connected in tandem?—If two amplifier stages each having a frequency response of 0.80 at a given frequency (f_1) and relative to a reference frequency of 420 Hz are connected in tandem, the overall frequency response for the two stages will be:

$$0.8^2 \text{ or } 0.8 \times 0.8 = 0.64$$

Therefore, the frequency (f_1) will show a loss of 4 dB compared to the reference frequency of the response at (f_1) and is 64 percent of the reference frequency. This will hold true regardless of the type of interstage coupling employed. Percentage compared to decibel loss may be read directly from the chart in Fig. 25-133.

12.34 What is the permissible variation of supply voltage (E_{bb}) for resistance-coupled amplifiers?—Using a given group of circuit components, the supply voltage may vary up to about 50 percent without seriously affecting the operation of the amplifier. However, this should be the exception rather than the rule, as the distortion will be increased and the output signal voltage will drop.

The change of output signal voltage with a change of supply voltage (E_{bb}) may be calculated:

$$E_{o1} = \frac{E_o E_{bb2}}{E_{bb1}}$$

where,

E_{o1} is the new output signal voltage,
E_o is the original output signal voltage,
E_{bb1} is the original supply voltage,
E_{bb2} is the new supply voltage.

12.35 What are the practical factors affecting the frequency response of a resistance-coupled amplifier?—Low-frequency response can be affected by any one or more of the components of the circuit. The reactance of the coupling capacitor C rises rapidly below a frequency of 100 Hz causing a loss at the lower frequencies. As an example, a 0.10-μF capacitor has a reactance at 50 Hz of 30,000 ohms and a reactance at 20 Hz of 80,000 ohms. This increase of reactance reduces the signal voltage at the control grid of the second stage. Increasing the capacity of the coupling capacitor to reduce its reactance increases its physical size adding to the shunt capacities of the circuit and reducing the high-frequency response. This is particularly true if the coupling capacitor is in a metal case. The loss at the lower frequencies causes phase shift which rises to an appreciable amount. One microfarad appears to be the practical limit of a coupling capacitor.

To obtain a uniform frequency response at a frequency of 20 Hz will require a cathode-bypass capacitor of at least 100 μF and preferably 1000 μF. Such values are practicable as they are of low voltage and the units are small in physical size. If these values are impractical, the cathode-bypass capacitor may be eliminated, with a resulting loss of gain because of degeneration in the cathode circuit. However, this will reduce frequency discrimination and phase shift. In some instances it may be desirable to use battery bias with the cathode connected to ground.

Another offender is the screen-grid bypass capacitor. If of insufficient size, degeneration is induced in both the screen and control grid circuits, reducing the stage gain. The screen-grid bypass capacitor should be at least 10 μF and larger if practicable. In any case, the reactance of the capacitor should be one-tenth the resistance of the screen-grid dropping resistor.

Another source of trouble is common coupling of the amplifier stages through the internal impedance of the power supply causing low-frequency motorboating. If the power supply is of the regulated type, the greater portion of the trouble is eliminated. The use of a large bypass capacitor at the output of an unregulated power supply will reduce common coupling. (See Question 12.36.)

12.36 What is a decoupling circuit? —When three or more resistance-coupled stages are used, decoupling resistors R_d and capacitors C_d are connected in the plate-circuit returns of each stage as shown in Fig. 12-36A. The purpose of these decoupling networks is to prevent common couplings between the stages through the common impedance of the power supply. Decoupling may be dispensed with if the amplifier consists of only two stages, as the plate current for the two tubes is out of phase and will not couple through the power-supply impedance.

In a three-stage amplifier, the plate currents of the first and third stages are in phase and will couple through the power supply. This condition manifests itself by a low-frequency oscillation called motorboating. This undesirable coupling is eliminated by the decoupling capacitor C_d which offers a low impedance path to ground to the signal voltage, while the resistor R_d offers a high impedauce to ground through the power supply route. If the ratio of capacitance C_d to resistance R_d is high, the sigual is forced to return

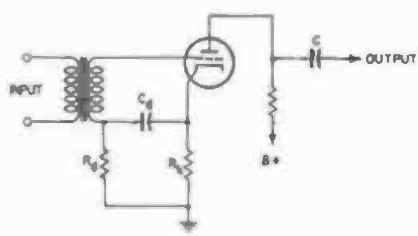


Fig. 12-36B. Decoupling circuit in the control-grid return circuit of a transformer-coupled amplifier.

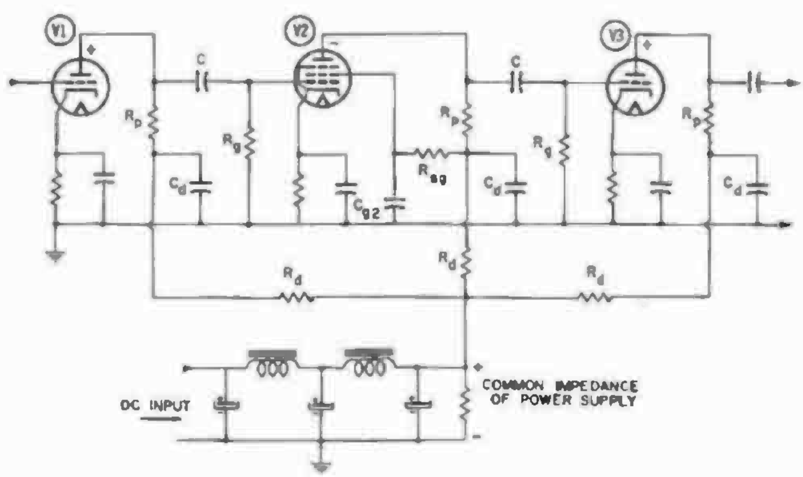


Fig. 12-36A. A three-stage, resistance-coupled amplifier with decoupling circuits C_d and R_d.

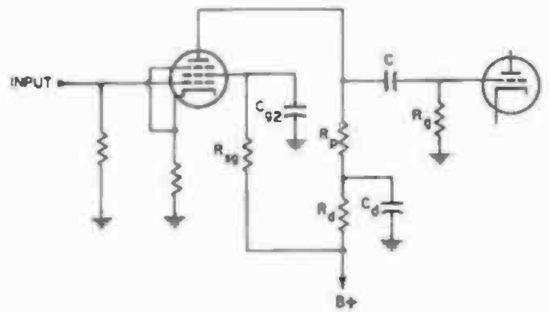


Fig. 12-37. A resistance-coupled amplifier with a low-frequency compensating circuit consisting of resistor R_d and capacitor C_d.

to ground through capacitor C_d rather than through the power supply. Practical values of resistance for R_d are 10,000 to 25,000 ohms. Capacitance values for C_d are 10 μF or greater.

The internal impedance of the power supply may be lowered by shunting the output with a capacitor of 100 microfarads; however, this is often impractical. Another form of decoupling is shown in Fig. 12-36B. A resistor R_d and capacitor C_d are connected in the grid return of the input transformer. The capacitor C_d has a low reactance to the signal voltage while the resistor R_d offers a high-impedance path forcing the signal to return back to the cathode circuit.

12.37 What is a low-frequency compensating circuit in a resistance-coupled amplifier?—Elements connected in the plate circuit to increase the low-frequency response. The circuit elements are those of the decoupling circuit shown in Fig. 12-37. By the proper choice of values, the circuit may be used for both decoupling and compensation. As a compensator circuit it serves two purposes; (1) it introduces phase shift in the plate circuit, compensating for the phase shift of the coupling circuit C and R_g, and (2) it effectively increases the plate-load impedance at the lower frequencies, thus maintaining the gain at the lower frequencies.

The resistor R_d should be kept as large as practicable, say 10,000 to 20,000 ohms. The larger the R_d, the lower the compensation may be carried; however, the value must not be made too great or the drop in voltage at the plate will be excessive. The most effective compensation is obtained when the time constant of the compensating circuit is the same as for the coupling capacitor C and the grid resistor R_g.

12.38 How is the time constant for the compensating circuit in Fig. 12-37 controlled?—For compensation down to 20 Hz, time constants of 0.02 to 0.5 are used. Knowing the time constant desired, the capacitance may be calculated:

$$C_d = \frac{T}{R_p}$$

where,

T is the time in microseconds,
R_p is the plate-load resistance in ohms,
C_d is the decoupling capacitor in microfarads.

The grid resistance may be calculated:

$$R_g = \frac{T}{C}$$

where,

C is the coupling capacitor,
R_g is the grid resistor,
T is the time in microseconds.

12.39 What is a direct-coupled amplifier?—A resistance-coupled amplifier in which the coupling capacitor C has been omitted. In such amplifiers, the plate of the driving tube is connected directly to the control grid of the following stage. This type coupling is shown in Fig. 12-39. The only function of the coupling capacitor is to isolate the control grid of the second stage from the dc potential on the plate of the preceding stage.

In direct-coupled amplifiers, the cathode of the second stage is made sufficiently positive above ground that the difference in potential between the cathode and plate is equivalent to the normal grid bias for the control grid of the second stage.

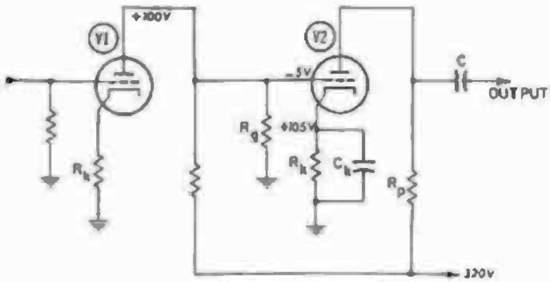


Fig. 12-39. A direct-coupled amplifier. The negative bias voltage at the control grid of V2 is the cathode voltage of V2 minus the plate voltage of V1.

In the circuit shown the cathode is made plus 105 volts above ground. The actual voltage at the plate of V1 is 100 volts positive. The control grid of V2 is returned to ground. Subtracting the plate voltage of V1 from the cathode voltage of V2 makes the control grid minus 5 volts, its correct bias voltage.

12.40 What is a Loftin-White direct-coupled amplifier?—This well known amplifier is shown in Fig. 12-40. The plate of V1 is directly connected to the control grid of V2. The voltage distribution may be traced by starting at the negative end of the power supply. The grid of V1 is connected to ground through the grid resistor R_g. The proper bias for this tube is obtained by connecting the cathode to point A on the voltage divider so that when the circuit is in operation, the total current flow through the resistance between the negative end of the voltage divider and point A gives the required voltage drop.

The plate of V1 is connected to point C on the voltage divider through the plate load resistor R_p which also serves as a grid resistor for the following tube V2. Since plate current flows through resistor R_p, it must be returned to a point on the voltage divider which will supply the proper voltage to the plate; therefore, it is returned to point C.

The plate of V2 is returned through a suitable load resistance or transformer to the positive end of the voltage divider. The cathode of V2 is returned to a point on the voltage divider where the proper bias voltage is obtained. The bias voltage at the control grid will be the voltage drop across the plate load resistor R_p minus the 53-volt drop across the voltage divider between points B and C. This results in a negative voltage (minus 16 volts) at the control grid of V2.

With the proper voltages established and the amplifier operating class-A, a low-distortion amplifier with a wide frequency range is possible.

Direct-coupled amplifiers of the foregoing type can be used to amplify very low frequencies and are particularly suited to pulse amplification, as the phase shift is reduced to a minimum.

12.41 What is a cathode follower?—A vacuum-tube circuit in which the signal voltage is applied to the control grid and the output signal is taken from the cathode circuit. The characteristics of a cathode follower are high input

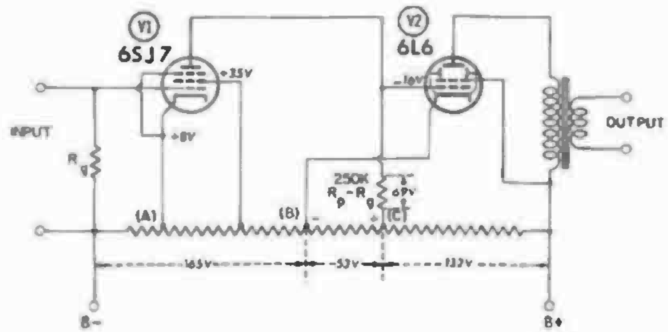


Fig. 12-40. The Loftin-White direct-coupled amplifier.

impedance and low output impedance. Because of degeneration in the cathode circuit, the amplification is always less than one. The circuit in Figs. 12-41A, B, and C are typical cathode-follower circuits. Cathode followers are practically distortionless if they are operated with the correct load impedance. However, if they are overloaded, limiting occurs and the distortion increases.

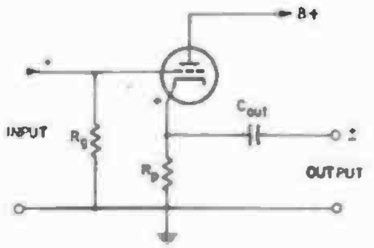


Fig. 12-41A. A cathode follower with 100% negative-current feedback.

In Fig. 12-41A the output signal is taken across the full load resistance R_p (the load resistance is the cathode resistor). In this circuit, 100-percent feedback exists because all of the output voltage is fed back to the input. The output capacitor C_{out} is selected on the basis of output reactance and frequency. The output capacitor prevents the load impedance from shorting the bias load resistor R_p to ground.

The circuit in Fig. 12-14B is used when the load resistor R_p is too large for the required bias voltage. The control grid is returned to a point on the load resistor R_p where the correct operating bias is obtained. The method used to obtain the bias voltage when the load resistor R_p is too small for the correct control-grid bias is shown in Fig. 12-41C. A resistor R_k is connected in series with the load resistor R_p and

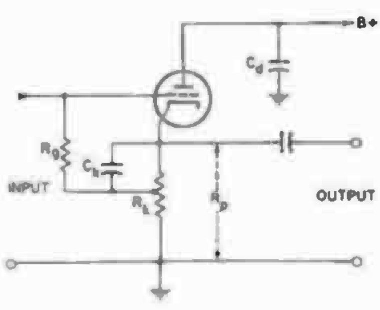


Fig. 12-41B. A cathode follower with a grid-bias tap on the load resistor.

bypassed in the usual manner. If the bias resistor is not bypassed, negative-current feedback will take place across all the resistance in the cathode circuit. In a properly designed cathode follower, the signal appears only in the cathode circuit. At times, it may be necessary to connect a large bypass capacitor from the plate to ground to prevent the possibility of the signal voltage getting

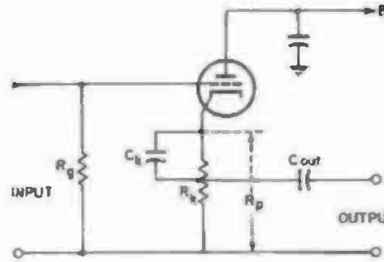


Fig. 12-41C. A cathode follower with bias resistor R_k.

into the power supply and causing common coupling to other parts of the amplifying system. Common coupling is discussed in Question 12.36.

12.42 How are tubes selected for cathode-follower circuits?—By first calculating the required transconductance using the formula:

$$g_m = \frac{10^6}{Z}$$

where,

Z is the output impedance desired in ohms,

g_m is the transconductance in micromhos.

12.43 What is the equation for calculating the load resistor R_p for a cathode follower?—For triodes the equation is:

$$R_p = \frac{Z \times r_p}{r_p - Z(1 + \mu)}$$

where,

R_p is the load resistor,

Z is the output impedance desired,

r_p is the plate resistance of the tube,

μ is the amplification factor of the tube.

For pentodes the equation is:

$$R_p = \frac{Z}{1 - (g_m \times Z)}$$

where,

R_p is the plate load resistor,

Z is the output impedance desired,

g_m is the transconductance of the tube in micromhos.

12.44 What is the equation for calculating the voltage gain of a cathode follower?—For triodes the equation is:

$$V_g = \frac{\mu \times R_p}{r_p + R_p\,(\mu + 1)}$$

where,

μ is the amplification factor of the tube,
R_p is the load resistor,
r_p is the plate resistance of the tube.

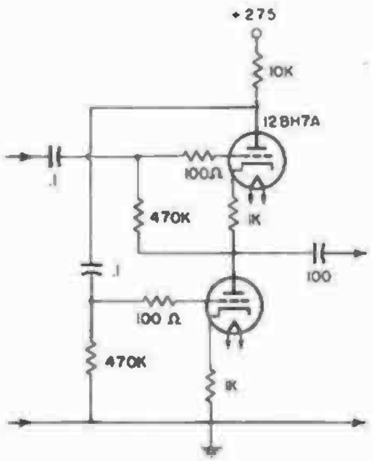
For pentodes the formula is:

$$V_g = \frac{g_m \times R_p}{1 + (g_m \times R_p)}$$

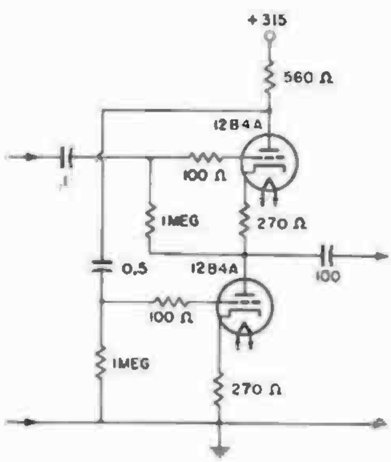
where,

R_p is the load resistor,
g_m is the transconductance in micromhos.

12.45 Describe a double cathode follower.—A double cathode follower, or as nicknamed "Totem Pole," is used in circuits where there is a large mismatch of load impedance, or a very low value of source impedance is required. The double cathode follower has about one-twentieth the output impedance of the conventional cathode follower. Typical circuits for a double cathode follower are shown in Fig. 12-45.



(a) Using a single 12BH7A tube.



(b) Using two 12B4A tubes.

Fig. 12-45. Double cathode-follower circuits.

12.46 Is it permissible to operate a cathode follower with a high load impedance?—Yes; this may be done if high output voltage is required. However, it will be necessary to use the circuit in Fig. 12-41B. The cathode capacitor C_1 must have a reactance that is negligible at the lowest frequency to be amplified. With a high resistance in the cathode circuit, the amplification factor approaches unity provided the plate voltage is increased above normal by the value of the grid bias voltage.

12.47 How far may the output circuit of a cathode follower be carried?—If the output impedance is in the order of 1000 ohms or less, 50 feet—more if coaxial cable is used. (See Question 25.212.)

12.48 Describe the procedure for plotting vacuum-tube load lines.—Load lines for vacuum tubes are made in the same manner as for a transistor described in Question 11.139, the principal difference being the names of the elements. In the transistor terminology the base corresponds to the control grid of a vacuum tube, the emitter to the cathode, and the collector to the plate element. In the instance of FET's, the gate corresponds to the control grid, the source to the cathode, and the drain to the plate.

12.49 What are the phase relationships in a cathode-follower circuit?—The input and output signals are in phase with each other.

12.50 What is a phase inverter?—Two tubes connected as shown in Fig. 12-50 and used for driving a push-pull amplifier stage. With proper design, two signals of equal amplitude and 180 degrees out of phase with respect to each other may be obtained. The incoming signal is applied to the control grid of tube V1, amplified, and then applied to the control grid of the upper tube of the push-pull stage V3. A portion of the signal at the grid of V3 is taken from the grid circuit by means of potentiometer P and applied to the control grid

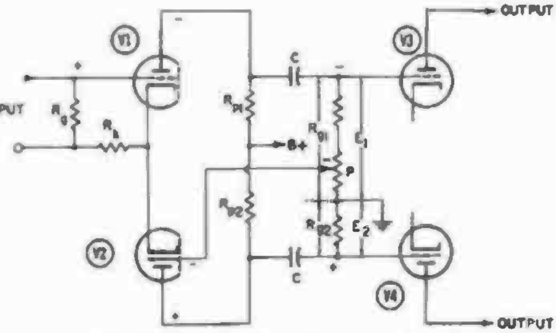


Fig. 12-50. A phase-inverter circuit.

of a phase-inverter tube V2. The signal, after inversion, is applied to the control grid of the lower push-pull tube V4. When the potentiometer P is properly adjusted, voltages of equal amplitude but 180 degrees out of phase are obtained at the control grids of tubes V3 and V4.

The term phase inverter is generally associated with a circuit employing two tubes, one as an amplifier and the other as an inverter, for the purpose of obtaining two voltages which are out of phase, although very frequently the terms phase inverter and phase splitter are used interchangeably.

12.51 What is a phase splitter?—A single-ended amplifier stage with the plate load divided between the cathode and plate circuits, as shown in Fig. 12-51. The incoming signal is applied to the control grid. Two signals of equal amplitude but 180 degrees out of phase are obtained from the plate and cathode circuits which may be used for driving a push-pull amplifier stage.

To assure that the two driving voltages will be exactly the same amplitude, the plate and cathode-load resistors must be matched to within one percent of their values. It is also equally important that the coupling capacitors C between the phase-splitter stage and the push-pull stage be of as nearly the same value as possible. If the values vary over too great a range, voltages developed at the control grids of the push-pull stage will not be of the same amplitude at frequencies below 50 Hz, because of the change in the reactance of each capacitor with frequency.

12.52 Can a transformer be used as a phase splitter?—Yes; a push-pull interstage transformer is, in reality, a phase splitter as shown in Fig. 12-52. The transformer delivers to the control grids of the output stage two voltages which are of equal amplitude and 180 degrees out of phase.

12.53 What is a floating paraphase inverter circuit?—Two vacuum tubes connected to produce two voltages 180 degrees out of phase as shown in Fig. 12-53A. The incoming signal voltage is fed to the control grid of tube V1, amplified, and applied to the control grid of V3. The signal voltage for V2 is taken from the junction of a voltage-divider network consisting of resistors R_{g1} and R_{g2}. Because of the phase reversal of the signal in passiug through V1, the signal is of the correct phase for V2. The signal is again reversed in passing through V2 and then applied to the control grid of V4. Resistor R_{g3} acts as a balancing resistor because the out-

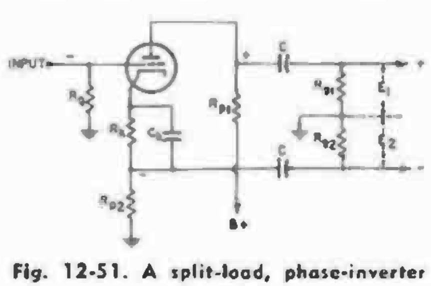


Fig. 12-51. A split-load, phase-inverter circuit.

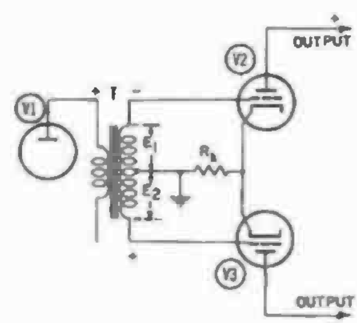


Fig. 12-52. A transformer phase splitter.

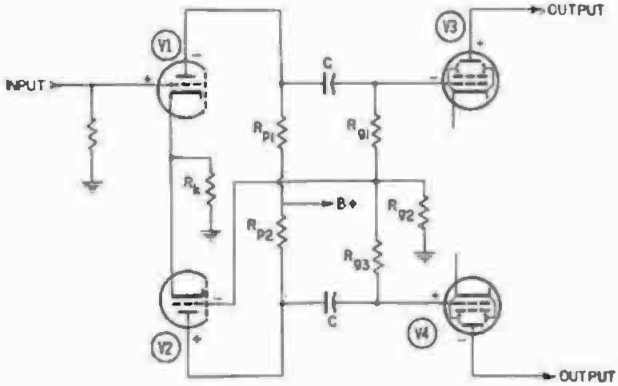


Fig. 12-53A. Floating paraphase inverter circuit.

put signals of both V1 and V2 are opposite in phase and flow through R_{g3} to ground, thus tending to cancel until a push-pull balance is achieved.

This circuit has good stability and low distortion. The common cathode resistor R_k is not bypassed. Without a signal at the input of V1, there is no feedback in the cathode circuit; however, for an unbalanced condition, feedback is generated.

A modified floating paraphase phase-inverter circuit designed to minimize out of balance conditions because of variations in tube characteristics is shown in Fig. 12-53B.

The plate currents for both halves of the tube pass through a common resistor R_{p3}. The grid swing for the second half of the tube requires only a fraction of the plate voltage swing to drive it; therefore, the control grid of the second tube is connected at the junction of the load resistors and the common resistor R_{p3}. With the proper values of resistors, the two halves of the circuit function with the same gain. If the plate currents of the two tubes were exactly the same, there would be no voltage variation across R_{p3}. The voltage variation at this point automatically adjusts itself so that the plate currents are slightly out of balance, sufficiently so to drive the second half of the tube.

12.54 What is a long-tailed phase inverter?—It is a phase inverter of a type similar to the floating paraphase and so called because of the long extension of the circuits from the plate. The inverter is essentially a cathode-coupled stage which will produce two voltages at the output having equal amplitudes and 180 degrees out of phase, as is shown in Fig. 12-54.

If a positive-going signal is applied to the grid of V1, plate current in this section will increase and a larger voltage will appear across R_1. The signal on the plate of V1 will be negative going. The increased voltage across R_1

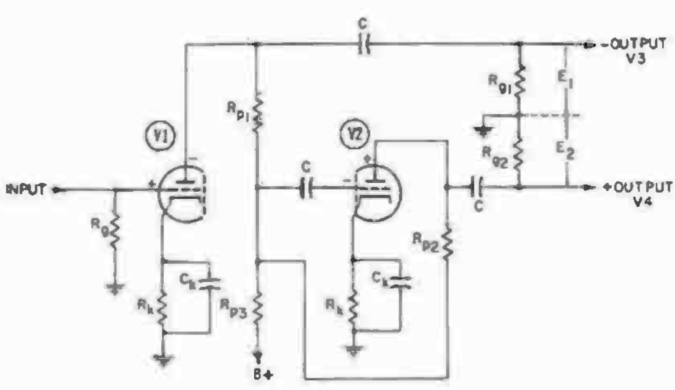


Fig. 12-53B. A floating paraphase inverter designed to minimize the variations in tube characteristics.

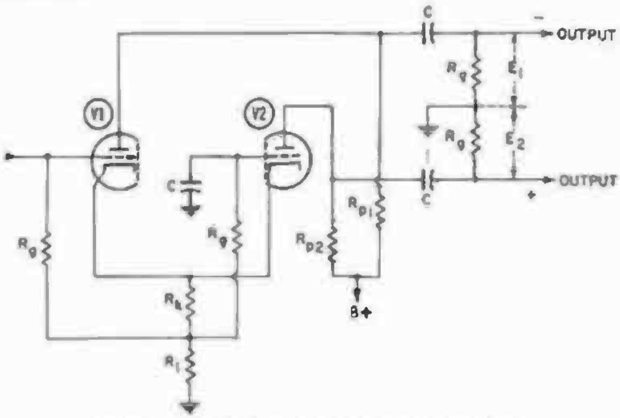


Fig. 12-54. A long-tailed phase inverter.

makes the cathode of V2 more positive than its grid. This, in effect, makes the grid of V2 more negative, the plate current in V2 decreases, and the plate voltage increases.

A common bias resistor R_k is employed together with suitable grid-leak resistors R_g. Capacitor C maintains the grid of V2 at ac ground. The size of this capacitor is important. If the capacitor is not large enough, phase unbalance will be introduced at the low frequencies, because then, in effect, the grid would not be returned to ground but to a tap on the voltage divider.

12.55 What is a fixed-bias, split-load phase splitter?—A tube connected as shown in Fig. 12-55A with two matched resistors, R_{p1} and R_{p2}, connected one in the cathode circuit and the other in the plate circuit. Signal voltages of equal amplitude, 180 degrees out of phase, are developed in the cathode and plate circuits and applied through coupling capacitors C to the control grids of a push-pull amplifier stage. Because of the high resistance in the cathode circuit of V1, the cathode is approximately 69 volts positive in respect to ground. The normal bias for a 6SJ7 tube is 9 volts. To obtain this bias voltage, a positive voltage is applied to the control grid through a 1-megohm resistor R_1 connected to the junction of the decoupling resistor R_d and the plate-load resistor R_{p1}. This places the grid 60 volts positive. With the cathode at 69 volts positive and the control grid at 60 volts positive, the control grid is 9 volts negative with respect to the cathode.

This circuit is quite stable and has good frequency characteristics with low distortion. The disadvantage of this circuit is that the gain is less than 1 for each half of the circuit. The voltage gain may be calculated:

$$V_g = \frac{2\mu R_p}{R_p(\mu + 2)r_p}$$

where,

R_p is the plate-load resistance,
r_p is the plate resistance of the tube,
μ is the amplification factor of the tube.

For best results, the load resistors R_p and the grid resistors R_g should be 1-percent resistors.

A split-load, phase-splitter circuit utilizing the diode-detector circuit of a radio receiver is shown in Fig. 12-55B. The two signals for driving the push-pull stage are taken from across the two 100,000-ohm resistors connected to the cathode of the diode rectifier. The capacitors C1 and C2 are small radio-frequency bypass capacitors and should be selected for a value that will not affect the high-frequency response.

12.56 What is an extended cathode-coupled phase inverter? — A self-balancing cathode-coupled phase inverter used for driving a single-ended push-pull amplifier such as that described in Question 12.126.

12.57 What is the equation for calculating the values of the coupling, screen-grid, and cathode capacitors for a resistance-coupled amplifier?

$$C = \frac{1}{2\pi f X_c}$$

where,

f is the lowest frequency to be amplified,
X_c is the reactance of the capacitor at the lowest frequency.

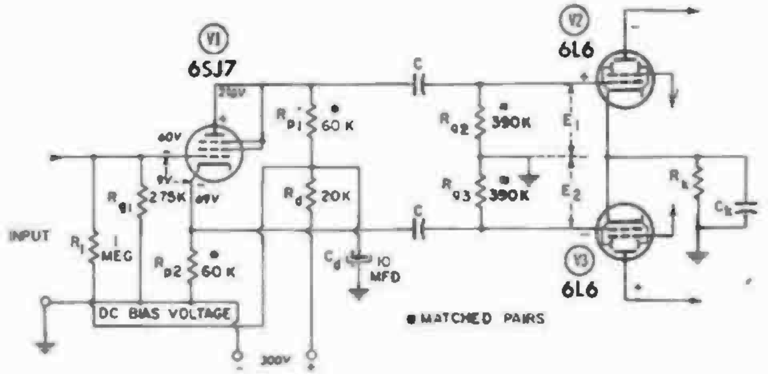


Fig. 12-55A. A split-load phase splitter with fixed bias on the control grid.

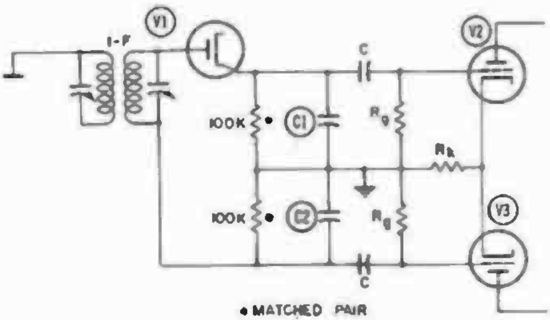


Fig. 12-55B. A phase-splitter circuit for a radio receiver using a diode rectifier.

The latter figure is based on a reactance of at least one-tenth the reactance of either the cathode resistor, screen-grid dropping resistor, or the grid resistor.

12.58 ***What is a single-ended amplifier?***—One having a single tube at the output.

12.59 ***What is a double-ended amplifier?***—It is an amplifier having two tubes at the output; a push-pull amplifier.

12.60 ***What is a parallel amplifier?***—One having two tubes of similar characteristics connected in parallel, such as two triodes with their control grids and plates connected in parallel, as shown in Fig. 12-60. This subject is further discussed in Question 12-111.

12.61 ***How are amplifiers classified?***—The classification depends primarily on the fraction of input cycle during which plate current is permitted to flow under full load conditions. The various classifications have been standardized characteristics and are known as: class-A, AB, B, and C. The dynamic operating of these amplifiers is explained in Questions 12.64 to 12.69 and 12.226 to 12.228.

12.62 ***What class amplifiers are employed for recording and broadcast purposes?***—For tube-type voltage amplifiers, class-A; for power amplifiers, class-A and class-AB. Solid-state voltage amplifiers are generally class-A, with power amplifiers operating class-B.

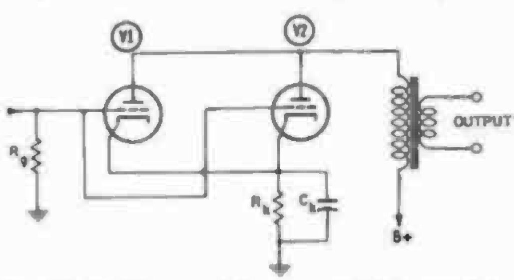


Fig. 12-60. Vacuum tubes connected in parallel.

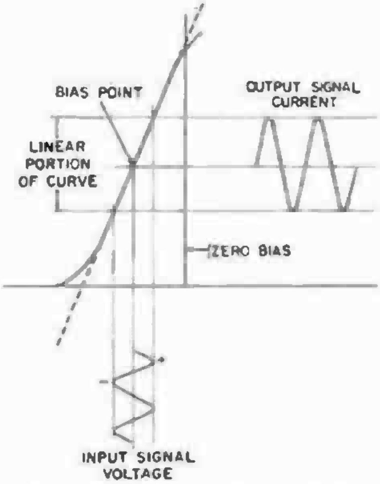


Fig. 12-63. Dynamic characteristics of a vacuum tube biased to operate class-A.

Power amplifiers of both types are always operated in push-pull.

12.63 What is a class-A amplifier? —An amplifier in which the grid-bias voltage is set to approximately one-half the cutoff voltage to obtain linear operation. (See Fig. 12-63.) The peak signal voltage at the control grid is limited to a value that will not exceed the dc grid-bias voltage. The control grid is never permitted to go positive and plate current flows at all times. Amplification is high and the harmonic distortion low. The efficiency is approximately 20 percent. Class-A amplifiers are used in high quality recording and reproducing systems. (See Question 12.226.)

12.64 What is a class-AB amplifier? —Class-A operation falls midway between class-A and class-B. Plate current flows for more than 180 degrees but less than 360 degrees of the signal at the control grid. Operating under these conditions, greater power may be obtained than with class-A and less distortion than with class-B. The efficiency is approximately 40 to 75 percent, depending on the bias voltage employed. (See Question 12.227.)

12.65 What is a class-AB$_1$ amplifier?—It is similar to the class-AB amplifier discussed in Question 12.64. The subscript indicates that grid current does not flow during any part of the cycle. The efficiency is approximately 40 percent. A typical characteristic is shown in Fig. 12-65.

12.66 What is a class-AB$_2$ amplifier?—It is a class-AB amplifier in which grid current is permitted to flow during some part of the input cycle. The efficiency is approximately 40 percent. A typical characteristic is shown in Fig. 12-66. Tubes biased for class-AB$_2$ are generally operated push-pull.

12.67 What is a class-B amplifier? —An amplifier that is biased at or near cutoff. Plate current flows during the positive half of the input grid signal and stops flowing during the negative half-cycle. The efficiency is from 40 to 60 percent. A typical characteristic is shown in Fig. 12-67. (See Question 12.228.) Tubes biased for class-B operation are operated push-pull.

12.68 Are any of the foregoing amplifiers operated as single-ended amplifiers?—Only the class-A and class-AB. Class-B amplifiers are always operated in push-pull.

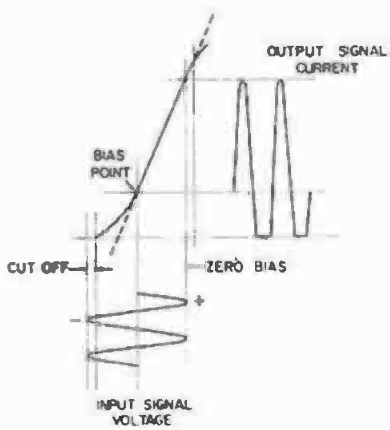


Fig. 12-65. Dynamic characteristics of a vacuum tube biased to operate class-AB.

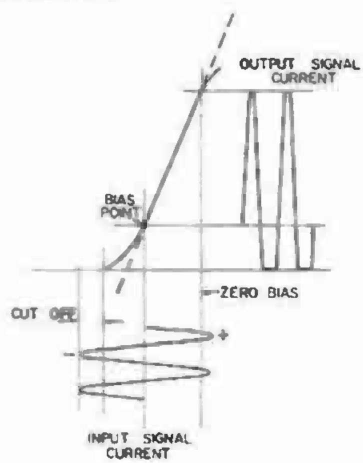


Fig. 12-66. Dynamic characteristics of a vacuum tube biased to operate class-AB$_2$.

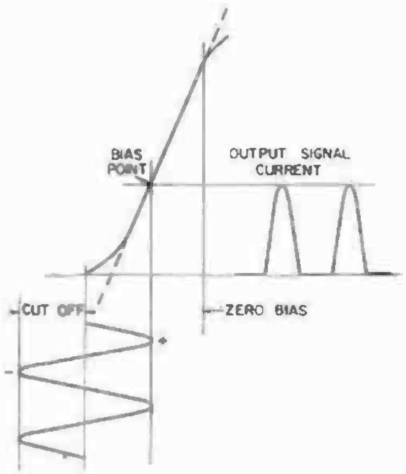


Fig. 12-67. Dynamic characteristics of a vacuum tube biased to operate class-B.

12.69 What is a class-C amplifier? —One whose operating point is located well beyond the plate-current cutoff point so that plate current flows for appreciably less than one-half cycle. (See Fig. 12-69.) Class-C amplifiers are used only for radio-frequency amplifiers where large power output and high efficiency are required. The efficiency of a class-C amplifier is in the order of 60 to 80 percent. The high distortion of a class-C amplifiers is overcome by the flywheel effect of the tuned circuits used.

12.70 What is a voltage amplifier? —An amplifier employed under conditions where voltage output is more important than power. Of course, all amplifiers produce power, although it may be rather small. Voltage amplifiers seldom produce more than one watt of power.

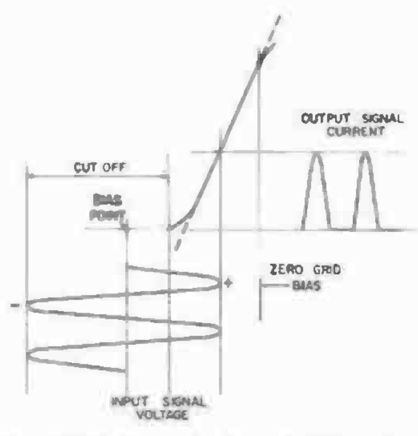


Fig. 12-69. Dynamic characteristics of a vacuum tube biased for class-C operation.

12.71 What is a power amplifier? —One which delivers a considerable amount of current to the load. The output impedance of a power amplifier is low, as a rule (16 ohms or less). However, many power amplifiers delivering a considerable amount of power have 600-ohm outputs for driving cutting heads and similar devices. Power amplifiers may be designed to deliver several hundred watts and are sometimes referred to as current amplifiers.

12.72 What is a preamplifier?—A small voltage amplifier used with photocells, pickups, and microphones, or any place where 30-50 dB of gain is required with a low-level output. Preamplifiers are designed to have extremely low noise and distortion, with or without equalization. A typical circuit for a microphone preamplifier suitable for recording purposes is shown in Fig. 12-72.

High-frequency equalization of up to 8 dB at 10,000 Hz may be introduced by varying the size of capacitor C1 and resistor R1 in the feedback loop in the cathode circuit of V1. The value of the capacitor C1 establishes the takeoff point of the equalization, and resistor R1 the magnitude of the equalization.

The amplifier shown has approximately 44 dB of gain and a frequency characteristic flat within 1 dB from 40 to 10,000 Hz in a nonequalized condition. The noise level is approximately 80 dBm. The distortion at plus 10 dBm is less than one percent. Preamplifiers are sometimes called preliminary amplifiers. The high-frequency response may also be varied by the value of capacitor C2 in the cathode circuit of V2. The 5-mH choke is used for eliminating rf interference.

12.73 What is a booster amplifier? —A small voltage amplifier similar to a preamplifier. It is used to compensate for the insertion loss of equalizers, or when a piece of equipment with a high insertion loss is inserted into a circuit. Also, it may be used in split-mixer circuits to compensate for the insertion loss of master and submaster gain controls. The frequency response is generally flat. High-power bridging amplifiers used with public address systems are sometimes referred to as booster amplifiers.

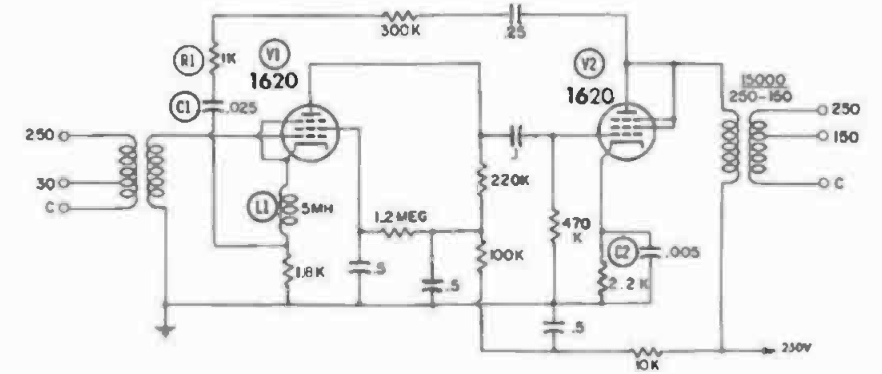


Fig. 12-72. A microphone preamplifier. Resistor R1 and capacitor C1 in the feedback loop control the high-frequency rise. The rf choke L1 is used to eliminate radio-frequency interference.

12.74 What is a line amplifier?—An amplifier following the output of a mixer network which, in turn, may feed a line or bridging bus. Line amplifiers generally are of medium gain, low distortion and noise with an output level of +20 dBm. In recording circuits they may be called recording amplifiers.

12.75 What is an isolation amplifier?—An amplifier of medium gain, with a variable gain control, consisting of several large and small steps so that the gain may be set for no-gain operation. It is used to isolate terminal equipment from a line, or to separate two devices to prevent interaction.

12.76 What is a monitor amplifier?—A high-quality amplifier used for listening to broadcast or recording circuits. Generally of medium gain with considerable power output, it must be of low distortion and noise. Input impedance may be matching or bridging.

12.77 What is a mixer-amplifier?—A portable amplifier used for remote broadcast purposes. It generally consists of three to four preamplifiers, mixer controls, line amplifier, volume indicator, and provision for monitoring the program. The schematic diagram for a typical portable broadcast mixer-amplifier is shown in Fig. 12-77. The amplifier is capable of producing a plus 22 dBm signal at the input of the line pad. Total gain is approximately 82 dBm. The frequency response is plus or minus 2 dB from 300 to 10,000 Hz with a slow rolloff below 300 Hz; at 100 Hz minus 4 dB and at 50 Hz minus 8 dB.

Center-tapped *balanced - to - ground* repeat coils are connected in each input circuit to isolate ground effects from external lines or equipment. (See Question 23.14.)

12.78 What is a direct-current amplifier?—Although a direct-current amplifier may be designed to amplify over a wide range of frequencies, its greatest use is in medical equipment where frequencies of 3 Hz or less and small pulses of direct current must be amplified.

Direct-current amplifiers are slow acting and rather unstable. Such amplifiers are not normally used in audio-frequency applications.

12.79 What is a compressor amplifier?—A special amplifier used for recording. Adjustments are available to adjust the characteristics for limiting the signal to a given output, or to compress the signal into a given volume range. Either characteristic may be used without excessive distortion.

When set for limiting, the amplifier acts as a ceiling control. When set for compressing, the output is held to a given maximum output level although the input signal increases in level.

When used for film recording, an additional circuit called a de-esser is switched in to remove or reduce the effect of sibilance in dialogue. Compressors are treated in detail in Questions 18.86 to 18.102.

12.80 What is an automatic level control compressor amplifier?—An automatic volume control amplifier designed to function both as an automatic level setting and compression amplifier. Such amplifiers are used with public address and background music systems. When used for background music application, it is set to maintain an *average level relative to the ambient noise level.* When the level has been established, it requires no further attention. The schematic diagram for an Altec-Lansing Model 436B compressor amplifier is shown in Fig. 12-80A, and its compression characteristics are given in Fig. 12-80B. Compressor amplifiers used for recording purposes are discussed in Questions 18.84 to 18.102.

12.81 *What is a side amplifier?*—The control amplifier used with a compressor-limiter or expander amplifier.

12.82 *Define the term off-set voltage or current.*—The term off-set is generally associated with dc or differential amplifier. It is the potential difference existing between the output with no signal at the input, and is caused by unbalance in the flow of current from two differential transistors or vacuum tubes to ground. Under these conditions, two voltages exist that are *unequal to zero.* Also, an off-set voltage is present if the two voltages are unequal to each other.

Assume two transistors are used in a differential amplifier; the off-set volt-

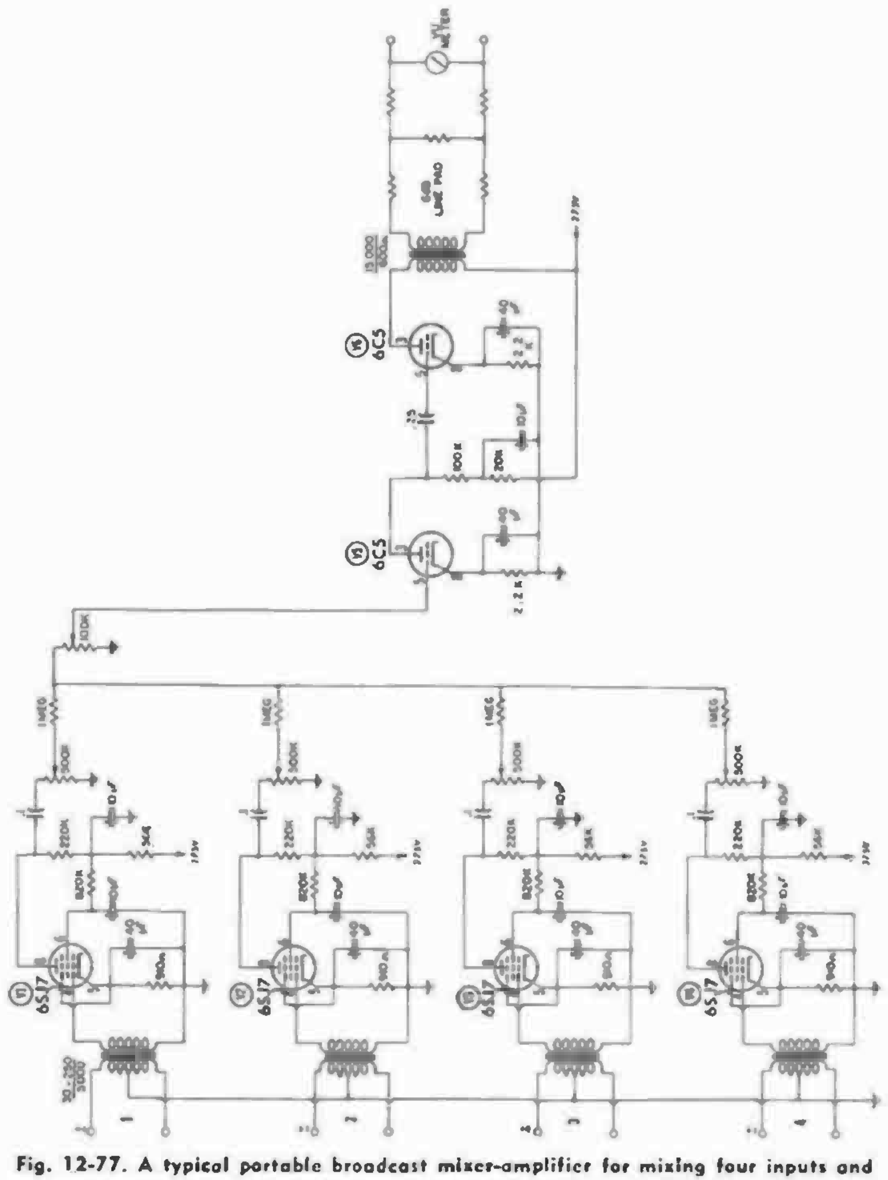


Fig. 12-77. A typical portable broadcast mixer-amplifier for mixing four inputs and feeding a 600-ohm line.

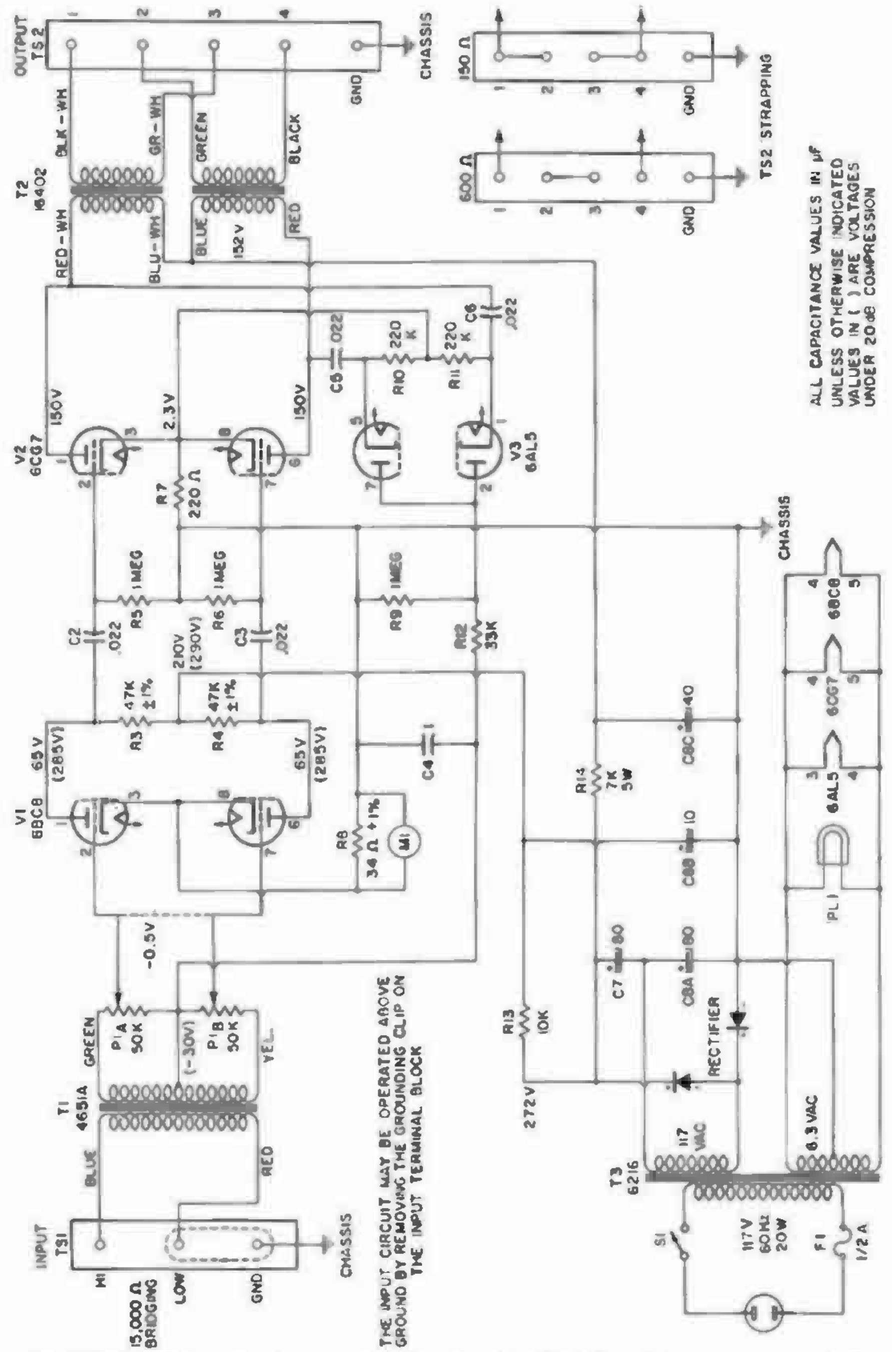


Fig. 12-80A. Schematic diagram for Altec-Lansing Model 436B compressor amplifier.

age, referred to the input, is simply the differential output off-set voltage divided by the double-ended voltage gain of the circuit. Using this method, a signal is applied to one input of the differential amplifier and varied until the differential output voltage is reduced to zero. The magnitude of the input voltage is then the off-set voltage *referred* to the input. Another method used is to reduce the base resistors to zero by shortcircuiting the input circuits to ground. Output off-set voltages measured in this manner are restricted in that unbalances caused by unequal base currents (beta) or by unequal base resistances are excluded. However, other unbalances are included in the measurement.

Unbalance is due to small variations in the geometry of the transistors, and also to differences in beta, internal resistance, thermal resistance, heat-flow paths, external resistances, and the leads within the transistor. (See Question 12.196.)

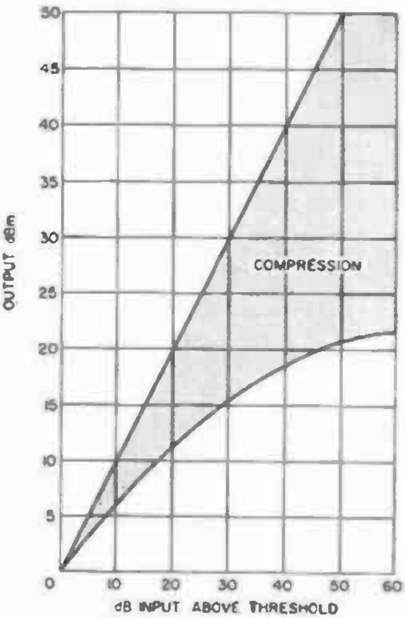


Fig. 12-80B. Compression characteristics for the Altec-Lansing Model 436B compressor amplifier.

12.83 What Is a noise-reduction (NRA) amplifier?—A special amplifier used for recording optical film to reduce the area of exposure on the film during periods of low or no modulation. It is also called a ground-noise reduction amplifier. (See Questions 18.55 to 18.76.)

12.84 Describe a program-leveling amplifier using electroluminescence control.—In broadcasting program material, some type of control system is required to prevent overmodulation of the transmitter. The control device is generally set for as near 100-percent modulation of the transmitter as possible, without overmodulation. To do this requires a control amplifier that limits the program peaks to a predetermined value so full advantage may be taken of the transmitter output. The use of conventional limiters or compressors in stereophonic broadcasting has posed problems such as unequal limiting between the left and right sides, causing an apparent loss of stereo balance. The Model LA-2A leveling amplifier (Fig. 12-84A) to be described is manufactured by Teletronix Engineering Co., and is designed especially for stereophonic and monophonic broadcasting and recording systems.

The description to follow is based on its monophonic use. The amplifier has essentially instantaneous gain reduction over a range of 40 dB, with no increase in harmonic distortion. The compression characteristics are shown in Fig.

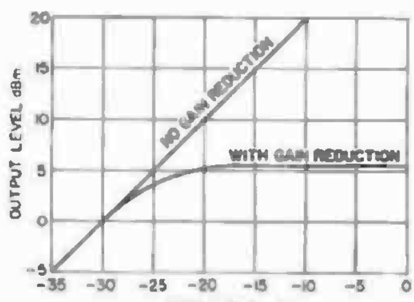


Fig. 12-84B. Gain-reduction characteristics of Teletronix Engineering Co. Model LA-2A leveling amplifier.

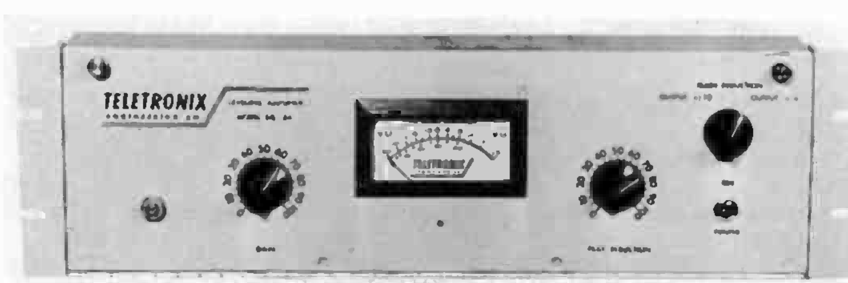


Fig. 12-84A. Teletronix Engineering Co. Model LA-2A leveling amplifier.

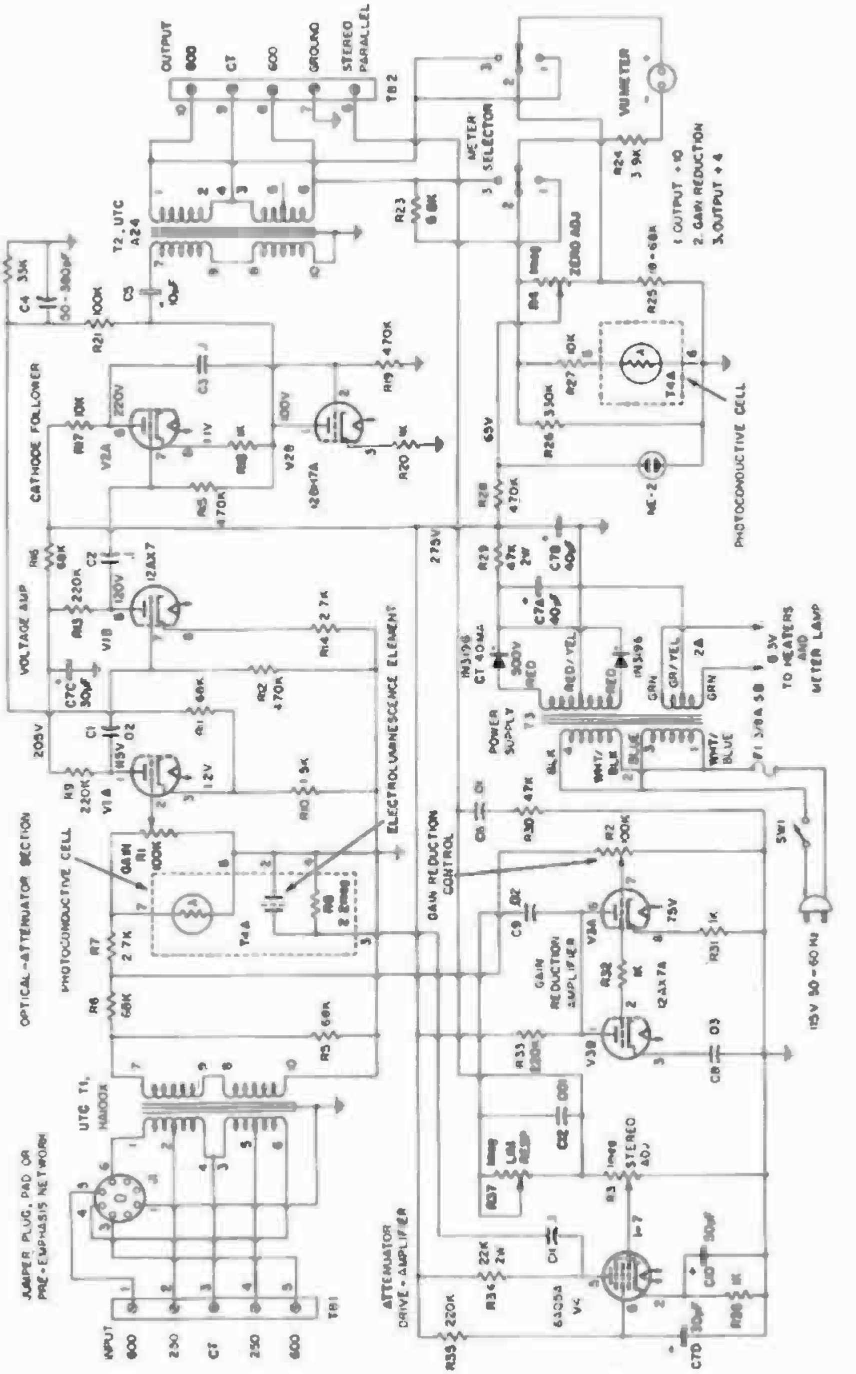


Fig. 12-84C. Schematic diagram for Teletronix Engineering Co. leveling amplifier Model LA-2A.

12-84B. Compression action starts at the breakaway point at minus 30-dB input and continues up to minus 20 dB, at which point the curve becomes horizontal or limiting in its action. When increasing the input signal another 20 dB, the output is held to within 1 dB. Thus, both compression and limiting are combined in its action. The heart of the leveling amplifier is an *electro-optical attenuator system* which is connected ahead of the voltage-amplifier stages.

The advantage of this type attenuation system is that the voltage-amplifier gain is fixed and not varied as in the conventional compressor-limiter. (See Question 18.84.)

The optical attenuator consists of a photoconductive cell which is optically coupled to an electroluminescence light source. This device provides a light intensity which is proportional to the audio voltage applied to its terminals. The gain or level-controlling element is a photoconductive cell. The internal resistance of the cell decreases with an increase of light intensity impinging on the luminous surface. Since light is produced by the audio-signal voltage, the response of the amplifier is instantaneous. Because no electrical filters are required in the control circuits, the speed of operation is limited only by the characteristics of the photoconductive cell. The cell will provide extremely fast attack and release times, the release time being about 60 milliseconds for 50-percent gain reduction, then a gradual release over a period of 1 to 15 seconds to a point of complete release. This device is further discussed in Question 5.97.

Referring to the schematic diagram of Fig. 12-84C, the input signal is applied directly to the optical attenuator from input transformer T1. At this juncture the signal is sent in two directions, to the input of voltage amplifier V1A through gain control R1 and through R2 at V3A from the junction of R6 and R7. The signal passes through V3A and V3B and is capacity-coupled through the limiting-response control R37 and stereo control R3 to the control grid of V4, the electroluminescence attenuator drive amplifier, and back to the lamp of the optical attenuator which controls the amount of attenuation and signal level to voltage amplifier V1A and V1B. The signal level to the voltage amplifier is also controlled by gain control R1. The voltage-amplifier stages have a net gain of 40 dB, with a feedback loop from the plate of V2A supplying 20 dB of negative-feedback voltage back to the cathode of V1A. The output stage consists of a single 12BH7A tube connected as a double cathode follower (see Question 12.45), providing a very low output impedance, and flat frequency response with low distortion.

For stereophonic broadcasting, an output signal is taken from the junction of limiting control R37 and stereo control R3 and brought out to terminal 6 on the right hand terminal strip. When two leveling amplifiers are used for stereo broadcasting, this terminal is connected to a similar terminal on the second amplifier; thus, the control voltage becomes common to both amplifiers. A gain-reduction control voltage generated at either amplifier will cause equal gain reduction in both amplifiers. The VU meter serves two functions; it indicates the output level and the gain reduction in dB.

During periods of no gain reduction, the meter reads zero. The scale is calibrated in terms of a standard VU meter. The VU meter circuitry consists of a zero-set control, and a second photoconductive cell connected across the balance circuit. The cell decreases its resistance in accordance with the intensity of the optical attenuator. Thus, the VU meter is made to follow the variations in gain reduction with the attenuation of the voltage amplifier. The NE-2 neon lamp acts as a voltage regulator to stabilize the 65 volts used for energizing the balance circuit.

The device is capable of 30-dB gain reduction or limiting, with less than 0.5-percent harmonic distortion. The gain reduction is a function of input level and is independent of frequency.

12.85 ***Describe a program-distribution amplifier?***—An amplifier or amplifier system for distributing a program signal to a number of feeds or services simultaneously. Figure 12-85A shows such an amplifier with a simple resistive combining network in its output for supplying six individual services. The combining network is of the conventional balanced type commonly used for sound mixers. The design of such networks is described in Questions 9.25 and 9.44. Disadvantage of the circuit shown is that little isolation between circuits is afforded, and if trouble should develop on any one feed circuit, the whole transmission may be lost.

Combining amplifiers have also been designed by making the output impedance of the amplifier very low, on the order of 1 to 2 ohms. The feed lines are then connected across this low output impedance using bridging impedances of 1000 ohms. If one of the feed circuits is shorted using this arrangement, little effect will be noted unless grounding troubles develop. This type circuit should be used only as a last resort.

The most desirable type program-distribution system is that shown in Fig. 12-85B, where each feed circuit has its own individual amplifier, affording complete isolation from the other circuits. Now if trouble develops, only the circuit concerned will be affected.

The system shown in Fig. 12-85B was developed for feeding radio and television networks, recording equipment, and public address systems simultaneously. The program signal is fed from the output of a mixer at a level of plus 4 dBm over a 600-ohm line to the bridging bus of the distribution amplifier system. Twenty-two plug-in amplifiers of the type shown in Fig. 12-85C are bridged across the bus. Normally, the amplifiers are designed to operate from a 600-ohm source; however, for this particular application a bridging resistor of 30,000 ohms is connected in series with one side of the 600-ohm input transformer. A 600-ohm resistor is connected internally in parallel with the 600-ohm input in the amplifier. These two resistors form a 30,000-ohm bridging input pad as shown in Fig. 12-85D. Installing the 600-ohm resistor in the amplifier and the 30,000-ohm series resistor at the bus will permit the replacement of any amplifier without resorting to the use of a soldering iron. Amplifiers 19 and 20 employ a 10,000-ohm bridging pad. This was done to obtain an output level of plus 8 dBm for feeding broadcast lines. All other outputs are zero dBm.

For feeds requiring a lower level, external pads are used. Connection of the pad at the input of the source being fed, rather than at the output of the program distribution amplifier (if the line is long), increases the signal-to-noise ratio as described in Question 23.133.

The method in which this amplifier was used on one remote installation to feed 22 different sources is shown in Fig. 12-85E. The circuit diagram of a Langevin Type 5116 plug-in amplifier is shown in Fig. 12-85F. Two views of the completed program-distribution amplifier appear in Figs. 12-85G and H. The microammeter in the rear view (Fig. 12-85H) is used for measuring the plate currents of the tubes and is actuated by small push buttons on each amplifier chassis.

Two power supplies may be seen on the lower shelf of Fig. 12-85G. The outputs of all amplifiers are terminated on binding posts to facilitate the connection to lines and the connection of attenuator networks. Three Cannon XL connectors are provided for the mixer output and two output circuits for feeding magnetic recorders.

The overall noise level of the amplifiers is minus 79 dBm. Distortion is 0.11 percent at 400 Hz. The frequency response is flat within plus or minus 1 dB from 30 to 16,000 Hz.

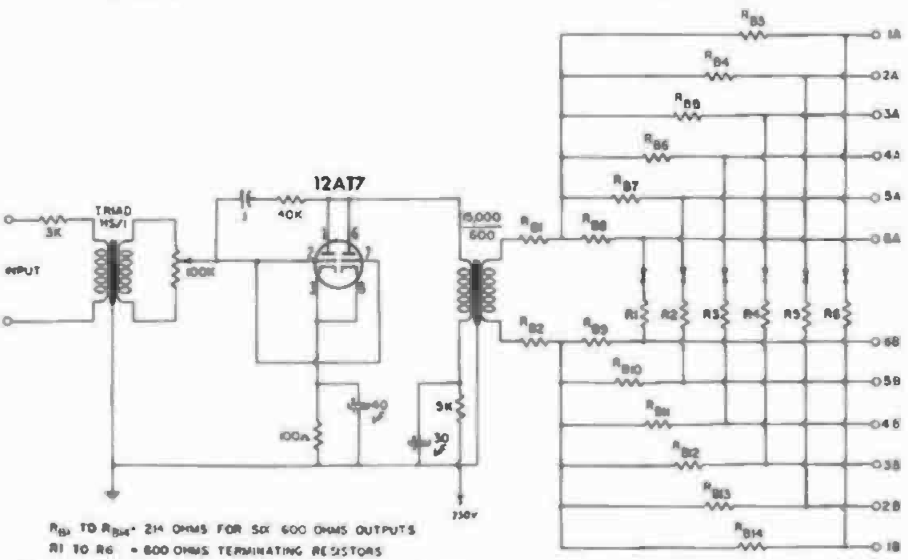


Fig. 12-85A. Program-distribution amplifier using a resistive network in the output.

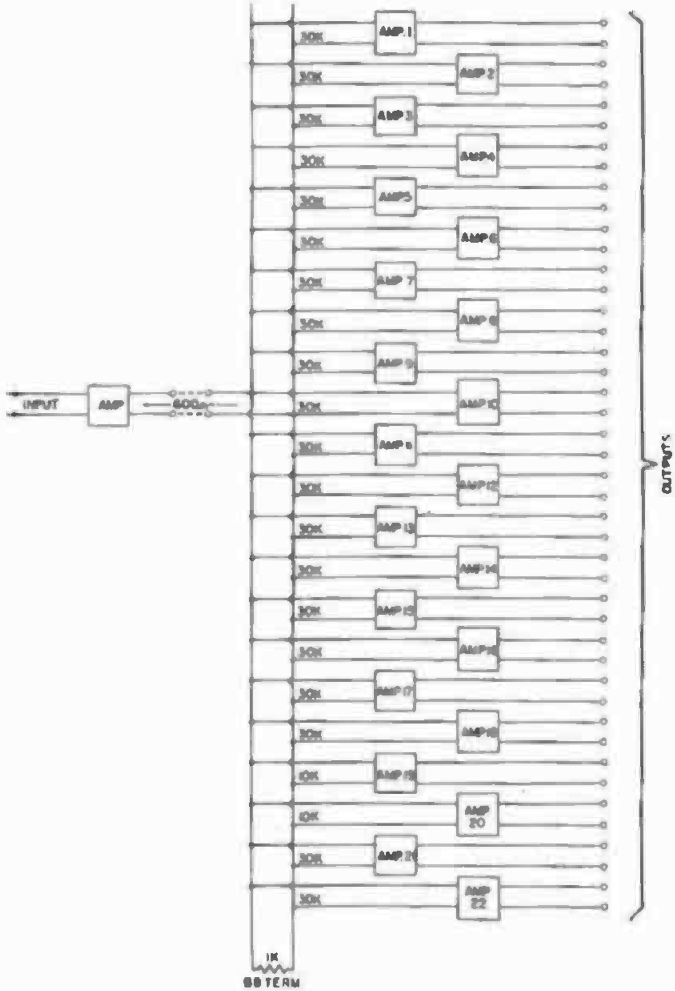


Fig. 12-85B. Program distributing amplifier system employing 22 separate amplifiers, one for each feed circuit. Amplifiers 19 and 20 employ 10,000-ohm bridging pads to raise the output level to plus 8 dBm.

Fig. 12-85C. Langevin type 5116 plug-in amplifier.

12.86 Describe a differential amplifier.—The term differential associated with an amplifier may be rather misleading; in reality, it is a difference amplifier. Such amplifiers are used in oscilloscopes for displaying a voltage difference that exists at every instant between the two signals applied to its inputs. When such an amplifier is used with an oscilloscope and two signals

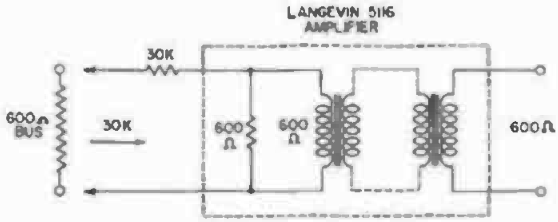


Fig. 12-85D. Bridging pad configuration.

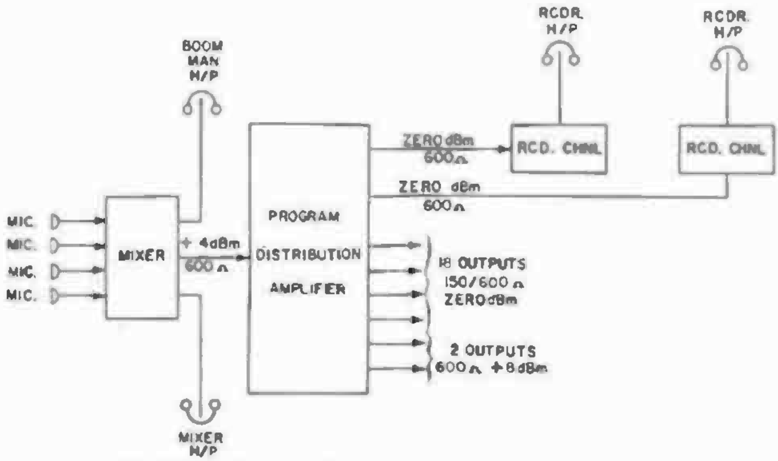


Fig. 12-85E. Block diagram for program-distribution amplifier system.

differing in both amplitude and coincidence are applied, the oscilloscope will display a complex waveform that represents the instantaneous difference between the two signals. On the other hand, when the two signals are identical in every respect, no display will appear on the oscilloscope cathode-ray screen.

The term differential implies a rejection of equal amplitude and coincident signals. However, the degree of rejection depends on the two amplifiers being identical. As this cannot be obtained in practice, the amount of rejection for a particular design is stated as a ratio between the two amplifiers, and is called the common-mode rejection ratio. Ratios of 20,000:1 are not uncommon. The design of such amplifiers is discussed in Question 22.72.

12.87 What is a bridging amplifier? —An amplifier having an input impedance of 10,000 ohms or greater and employing a special input transformer. Such amplifiers are designed to operate

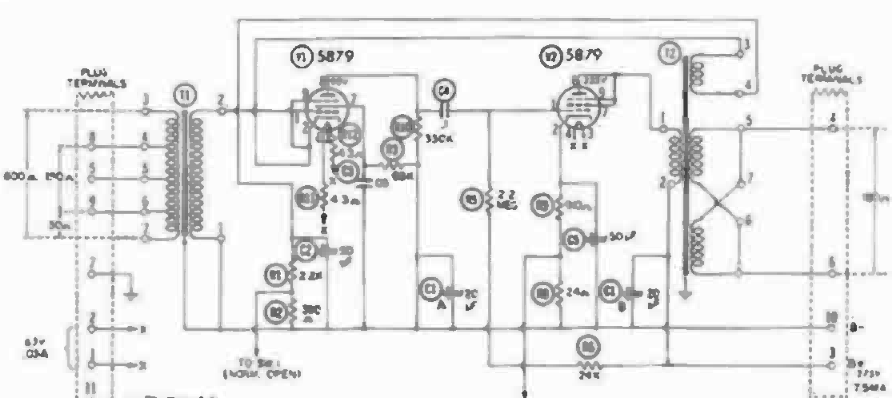


Fig. 12-85F. Circuit diagram of Langevin type 5116 plug-in amplifiers.

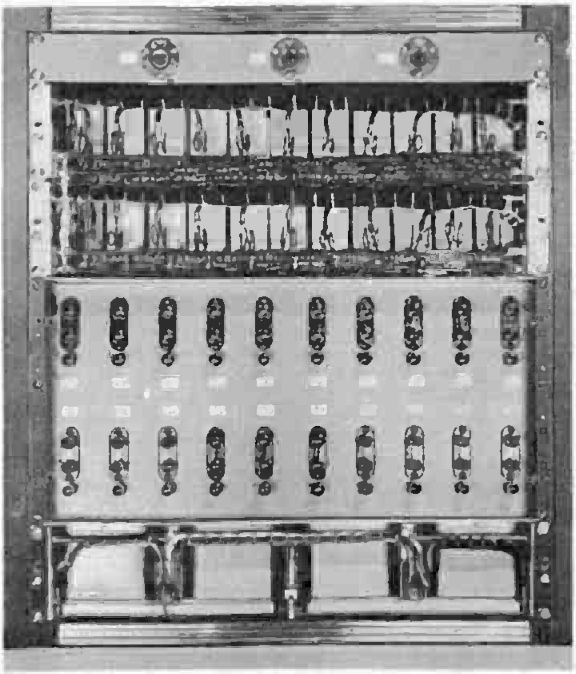


Fig. 12-85G. Front view of program-distribution amplifier showing the bridging bus and output terminals. (Courtesy, Lookout Mountain Air Force Station.)

from a low-impedance source, such as 600 ohms or less. Standard bridging impedance are: 10,000, 20,000, 25,000, and 30,000 ohms.

12.88 What is a cascoded amplifier?—An amplifier tube or tubes connected in such a manner that unusually low internal noise is obtained with high gain. A typical cascoded amplifier, as might be used in a piece of test equipment, is shown in Fig. 12-88A. The amplifiers consist of three tubes: a 12AT7,

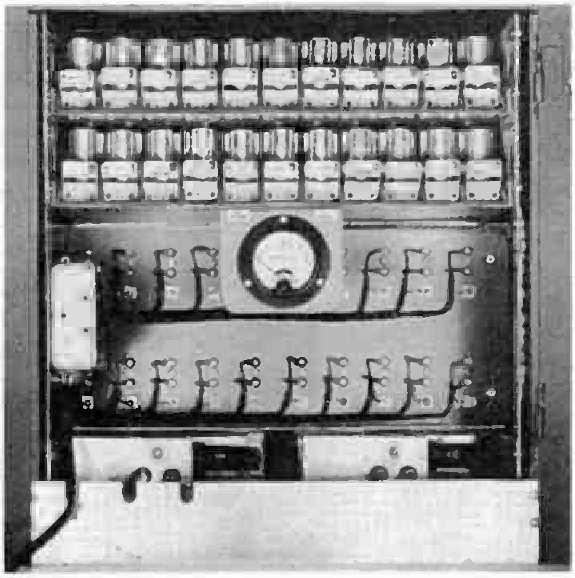


Fig. 12-85H. Rear view of program-distribution amplifier showing the 22 plug-in amplifiers. The power supplies are at the bottom of the case. (Courtesy, Lookout Mountain Air Force Station.)

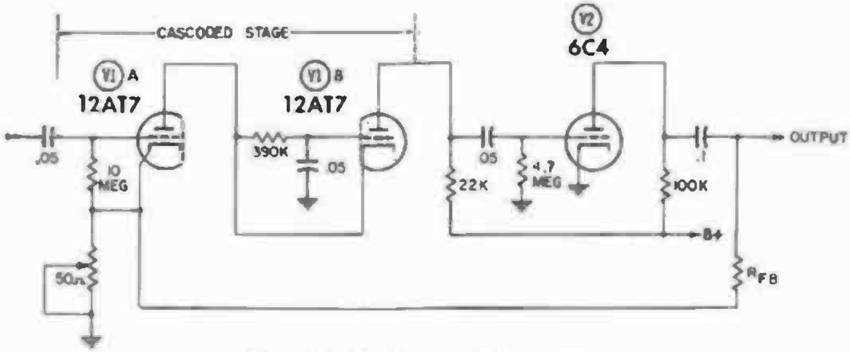


Fig. 12-88A. A cascoded amplifier.

V1A and V1B, an output tube V2, and a 6C4 medium-mu triode.

Tube V1A acts as a conventional voltage-amplifier stage. The plate load for this stage is supplied by the plate resistance of V1B and the 22,000-ohm resistor in the plate circuit of V1B. The signal voltage is amplified by V1A and direct-coupled to the cathode of V1B.

The gain in V1B is obtained by connecting its control grid through an 0.05-μF capacitor to ground and isolating the control grid from the cathode of V1B by means of a 390,000-ohm resistor. Thus, the control grid of V1B remains at a fixed potential while the cathode voltage is varied. This causes V1B to act as a grounded-grid amplifier.

Plate loading for V1B is supplied by a 22,000-ohm resistor in the plate circuit. The output signal voltage is coupled from the plate of V1B through an 0.05-μF capacitor and applied to the control grid of V2. The output signal is coupled through a 0.1-μF capacitor and returned to the cathode of V1A, creating a negative-feedback loop.

Current changes in the cathode circuit of V1A cause a change in the feedback voltage; consequently, the gain is changed. In this manner, gain stability is achieved. The variable resistor in the cathode circuit of V1A may be used to vary the gain over a limited range. A vacuum-tube voltmeter using a cascoded amplifier is described in Question 22.101.

A cascode preamplifier circuit developed by H. Wallman which has unusually low internal noise is shown in Fig. 12-88B.

12.89 What is an expander amplifier?—A reproducer amplifier with a nonlinear characteristic designed to increase its output level during loud passages of recorded music for the purpose of increasing the dynamic range. The amount of expansion is controlled by a rectifier in the side amplifier which utilizes the input signal to control the rate of expansion.

Volume-expansion amplifiers are, in reality, variable gain amplifiers that amplify the louder passages louder but have little effect on the low-level passages. Thus, an approach is made to the original volume range of the program material. Volume expansion is not used in professional recording and reproducing equipment because of the difficulty in maintaining the correct ratio between the recording compression and the reproducing expansion. Volume expansion has seen some use in home reproducing equipment; however, unless it is carefully controlled,

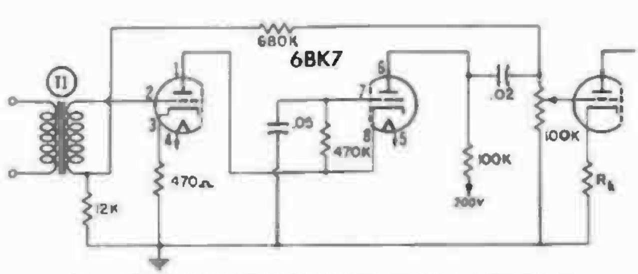


Fig. 12-88B. The Wallman cascode preamplifier.

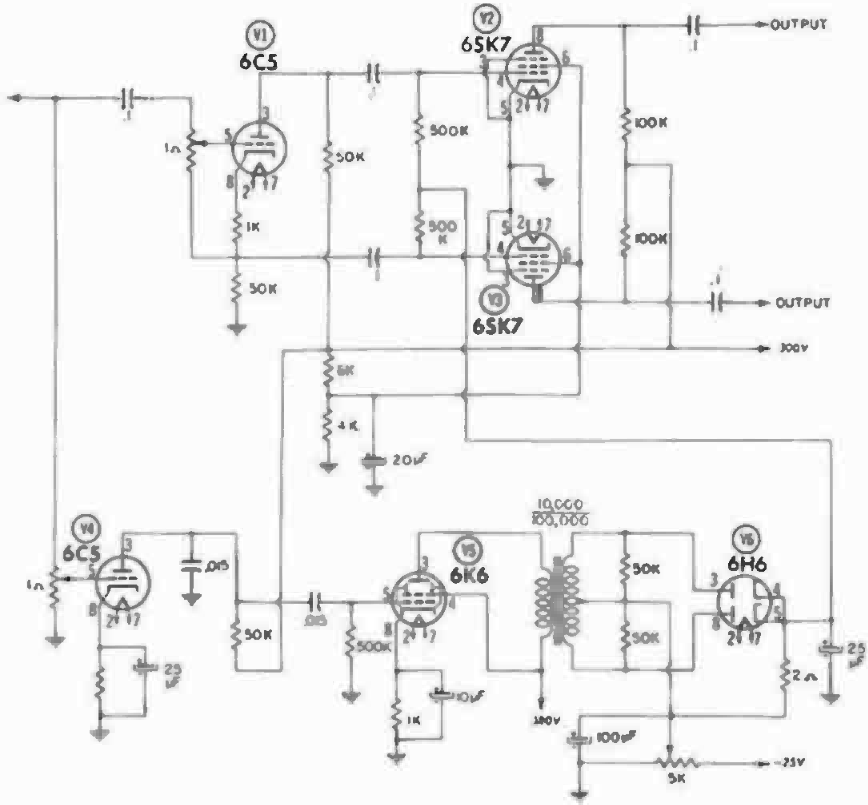


Fig. 12-89. A volume-expander amplifier.

volume expansion introduces distortion and will produce listening fatigue. Also, with the modern methods of recording, volume expansion is not required because of the wide dynamic range recorded in the original masters.

The circuit for a typical volume expander is shown in Fig. 12-89. The input signal is applied to the control grid of V1, a phase inverter, and then to the control grids of the push-pull stage, V2 and V3. Simultaneously, the input signal is applied to the control grid of a side amplifier V4 and then to V5 which is transformer-connected to a full-wave rectifier V6. The output from V5 is rectified and applied to the control grids of the 6SK7 variable-mu tubes in the output stage. An initial bias voltage is set by pot P1 in the rectifier circuit which establishes the amount of volume expansion. The changing value of the rectified dc signal at the output of V6 controls the amount of amplification of the variable-mu tubes V2 and V3. Thus, volume expansion controlled by the amplitude of the program signal is achieved.

12.90 What is a compander?—An amplifier similar to an expander, except for this important difference—the music in the original recording is compressed within a given range. When reproduced by the compander amplifier, the program material is expanded in the same ratio as recorded. Because of the difficulty in controlling the recording and reproducing characteristics, this system has not been used to any great extent.

12.91 What is a tuned amplifier?—An amplifier designed to pass only a given band of frequencies. It is used in audio test equipment and radio-frequency amplifiers.

12.92 What is a transformer-coupled amplifier?—An amplifier in which the stages are coupled by transformers, as shown in Fig. 12-92.

12.93 What is an impedance-coupled amplifier?—An amplifier coupled between stages by means of large choke coils in both the plate and grid circuits, as shown in Fig. 12-93.

The disadvantage of this type coupling is that the choke in the plate circuit must have a reactance at least

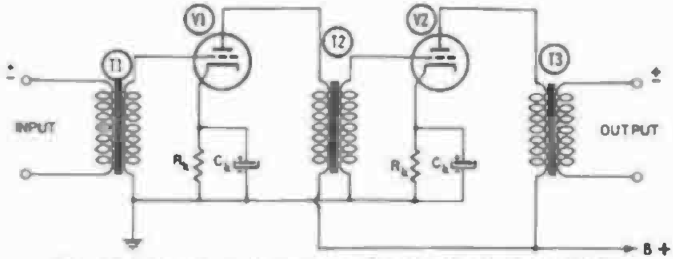


Fig. 12-92. A two-stage, transformer-coupled amplifier.

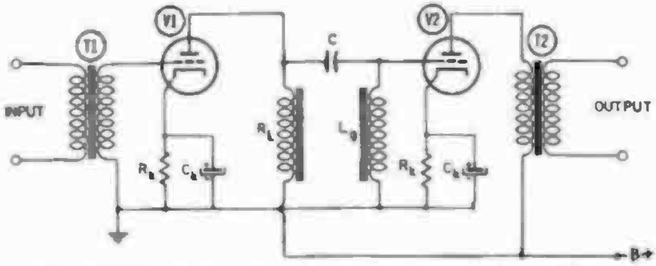


Fig. 12-93. A two-stage, impedance-coupled amplifier.

twice that of the plate resistance of the tube at the lowest frequency to be amplified. The choke in the grid circuit must also be of large proportions, which complicates the design, because the distributed capacity in the windings affects the high-frequency response.

12.94 What is a video amplifier?—A wideband, resistance-coupled amplifier with a frequency response from a few hertz up to several megahertz. They are used as picture amplifiers in television sets, radar equipment, oscilloscopes, and many similar types of equipment requiring extremely wide frequency characteristics.

12.95 What are in-phase amplifiers?—Amplifiers using cascode, grounded-grid, and inverted-input circuits. The input and output signals are in phase with each other.

12.96 What does the term rolloff mean?—The frequency response is slowly tapered off at either end of the frequency response curve for the particular device.

12.97 How many times does the phase of a signal reverse in an amplifier?—The number of phase reversals will depend on the number of tubes, transformers, etc. A typical two-stage amplifier is shown in Fig. 12-97 with the phase reversals shown by the instantaneous sines above the circuits. The total phase reversal for the amplifier shown is 720 electrical degrees.

For a single-stage *transistor* amplifier, the only configuration that will provide 180 degrees phase reversal between the input and output circuit is the grounded-emitter circuit. (See Question 11.60.)

12.98 What are considered good engineering practices in the design of audio amplifiers?—When practical, the power-output capabilities of an amplifier should be 6 to 10 dB higher than the working output level. The total rms

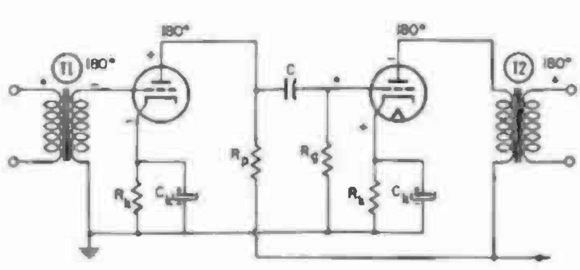
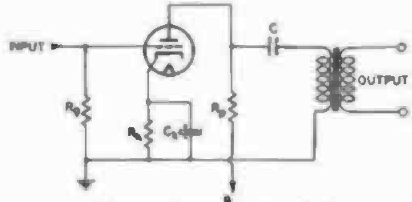


Fig. 12-97. Phase reversals in a two-stage, transformer resistance-coupled amplifier. Total phase reversals equal 720 electrical degrees.

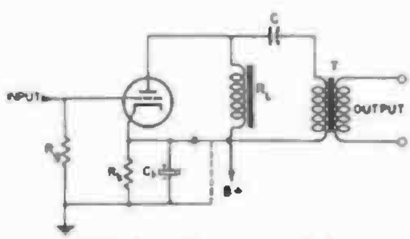
harmonic distortion should not exceed two percent at the maximum power output level, and one percent in the operating level. The internal noise should not be less than 40 dB below the maximum power output for monitor and sound reinforcement systems. For recording systems, it should be at least 60 dB.

12.99 Is it practical to couple a tetrode or pentode by means of a transformer?—No; the plate impedance is too high. Such tubes are generally resistance-coupled. This statement should not be applied to a beam-power tube which is a power-output tube and may be classed as a tetrode or a pentode, depending on its internal structure. (See Question 11.25.)

12.100 What is parallel-plate coupling a transformer?—Many times the manufacturer of a high quality transformer will specify that the transformer is to be operated with zero direct current in the primary winding to prevent saturation of the core. Referring to Fig. 12-100, the primary of the transformer is fed from the plate of the driver tube through a large coupling capacitor and the lower end of the transformer is returned either to the cathode of the driver tube or to ground. A resistor or an inductance is connected in the plate circuit of the tube to carry the plate current. Thus, only the ac signal current flows through the primary of the transformer. The coupling capacitor C must be of such a value that it does not



(a) *Resistance coupled.*



(b) *Impedance coupled.*

Fig. 12-100. Transformers parallel-coupled to the output of a triode.

resonate with the primary of the transformer at the low frequencies; otherwise, a severe resonant peak may result. For triode operation the value of the resistance (R_L) generally falls between 25,000 and 50,000 ohms. If an inductance is used, it must have an inductive reactance twice that of the plate resistance of the tube at the lowest frequency to be amplified.

12.101 If a dc milliammeter is connected in the plate circuit of a class-A amplifier, how does it react when a signal is applied?—If the tube is correctly biased, the meter will indicate a steady current, either with or without a signal at the input. As the signal amplitude is increased above the normal operating range, the plate current will start to fluctuate and, at the point of overload, will fluctuate quite rapidly over a wide range of current. If the tube is over-biased, the current will drop considerably and, if under-biased, the plate current swing will be the greatest in the positive direction.

Operating with the correct bias voltage and not overloaded but near the maximum signal voltage, the plate current will swing through a small arc of equal amounts of current, plus and minus the steady current (no signal).

12.102 What happens when the control grid of a class-A amplifier is driven into the positive region of grid voltage?—Grid current is caused to flow. The tube is driven into saturation causing it to draw excessive plate current generating harmonic distortion.

12.103 What is a zero-biased operated tube?—A circuit in which the control grid and cathode circuits of a tube are both returned to ground. Such circuits are discussed in Question 11.62.

12.104 What is a push-pull amplifier?—An amplifier circuit employing two tubes connected as shown in Fig. 12-104. The tubes may be operated either class-A, -AB, or -B (class-C is used for radio-frequency amplifiers only). Push-pull amplifiers afford greater output power with lower distortion and are used extensively in high-quality sound recording and reproducing systems. Referring to the diagram, it will be noted the upper half of the circuit is similar to the lower half; therefore, the tubes must be of the same type with the same static and dynamic characteristics. A common bias resistor may be used for both tubes. Transformer T2 supplies the load impedance to the plates of tubes V1 and V2 and couples them to the external load circuit. The primary of the output transformer is continuously wound and tapped in the exact impedance center of the winding and then connected to the plate supply voltage E_{bb}. The symbols for the various signal currents and voltages are indicated by the subscripts, which indicate which tube it is associated with.

It will be noted the control grid voltages E_{sig-1} and E_{sig-2} are of the same amplitude but 180 degrees out of phase with respect to each other. This same condition prevails for the signal output voltages E_{o1} and E_{o2} but, in addition, they are 180 degrees out of phase with the signal voltages on the control grids. The plate currents I_{p1} and I_{p2} are also of equal amplitude and 180 degrees out of phase. The instantaneous voltage across the entire primary winding of the output transformer is E_{prim}. The instantaneous current in the output winding is designated I_o and, without a signal at the control grids, is zero.

The dc path of tube V1 is from the cathode to the plate through the upper half of the output transformer T2 (points B to A) and back through the power supply to the cathode. For V2 the path is from the cathode to the plate through the lower half of T2 (point C to point A) and back through the power supply to the cathode. Points B and C are equally negative (no signal) with respect to point A, since the steady plate current is equal in both halves of the primary. Thus, total magnetizing force in the primary is zero and dc saturation of the core is avoided.

When sinusoidal signals E_{sig-1} and E_{sig-2} are applied to the control grids, sinusoidal plate currents I_{p1} and I_{p2} flow in the primary of transformer T. Current I_{p1} is 180 degrees out of phase with I_{p2} since the two control grid voltages are 180 degrees out of phase with each other. During positive swings of I_{p1}, point B on the primary becomes more negative with relation to point A. At the same time, the fall in I_{p2} causes point C to become less negative with respect to point A by an equal amount. Therefore, the voltage across the entire primary E_{prim} is twice the value of either E_{o1} or E_{o2}. A half-cycle later all polarities will be reversed.

Again the voltage E_{prim} across the primary is equal to E_{o1} plus E_{o2}. The relationship of E_{prim} in terms of E_{o1} and E_{o2} will hold true for all instantaneous values of plate current.

It is of importance that the signal voltages supplied to the control grids be of equal amplitude and exactly 180 degrees out of phase with respect to each other. These voltages may be obtained by means of a center-tapped transformer or a phase inverter similar to that described in Question 12.50. The dynamic characteristic of a class-A push-pull amplifier is described in Question 12.226. As a rule, for high power class-AB amplifiers, a self-bias resistor is not used but a fixed-bias voltage is employed as described in Question 12.229.

12.105 What are the advantages of operating tubes in push-pull?—Broadly speaking, a greater power output is possible with less harmonic distortion. More specifically, triode tubes which predominate in even harmonic distortion products can be operated at load values which will result in an increased power output and cancellation of the even harmonics. Under the proper load conditions, output powers up to 2.5

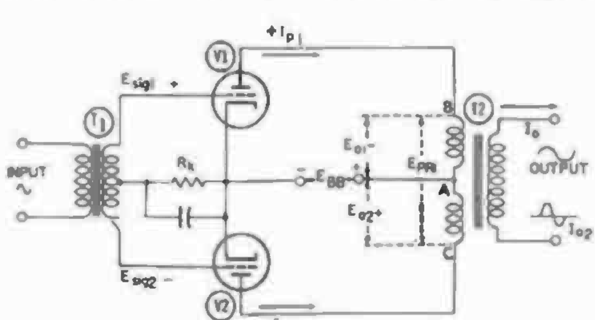


Fig. 12-104. Basic principles of a push-pull amplifier.

to 3 times that possible with a single tube may be obtained, with distortion components less than those of a single tube.

Because the odd harmonics predominate in a tetrode (four element tube) or a pentode (five element tube), they are not cancelled when operated in push-pull. However, if a pentode is operated into a higher than normal load impedance, the power output can be increased (with an attendant increase of second harmonic distortion). Operation of pentodes in push-pull will result in the cancellation of the second harmonic, leaving only odd harmonics. By the use of the proper load impedance and the addition of negative feedback, the push-pull connection may be used to lower odd harmonic distortion to negligible proportions.

Push-pull operation eliminates the dc magnetization of the transformer core material, preventing core saturation and distortion caused by the nonlinear magnetization of the core material. Power-supply ripple is cancelled by the fact that the plate currents in the amplifier are balanced; therefore, any ripple in the plate circuits caused by ripple voltage from the power supply will be cancelled. This reduces the amount of filtering in the power supply compared to that required for a single-ended amplifier. In certain types of push-pull amplifiers, the efficiency is quite high.

Beam-power tubes, although they may be considered to be pentodes, are treated differently; as a rule, the load impedance is roughly one-fifth that of the plate resistance. The value of the load impedance is determined by the class of operation and the type bias (self or fixed). Negative feedback is also recommended for operation with these tubes.

12.106 ***Why is it necessary to bypass the cathode resistor in a push-pull amplifier?***—Because it is next to impossible to maintain a perfect balance at all times between the two halves of the amplifier. If the tubes are perfectly balanced, the current flow through the cathode resistor for each tube is exactly the same. The odd harmonics cancel and only the even harmonics remain. This condition will cause the phase relationships to be such that the even harmonic components create a negative-current feedback, reducing the distortion by degeneration. When the tubes are not matched and the plate currents are out of balance, the phase relationships become such that odd harmonics are generated. This causes positive feedback with the odd harmonics predominating. Therefore, it is advis-

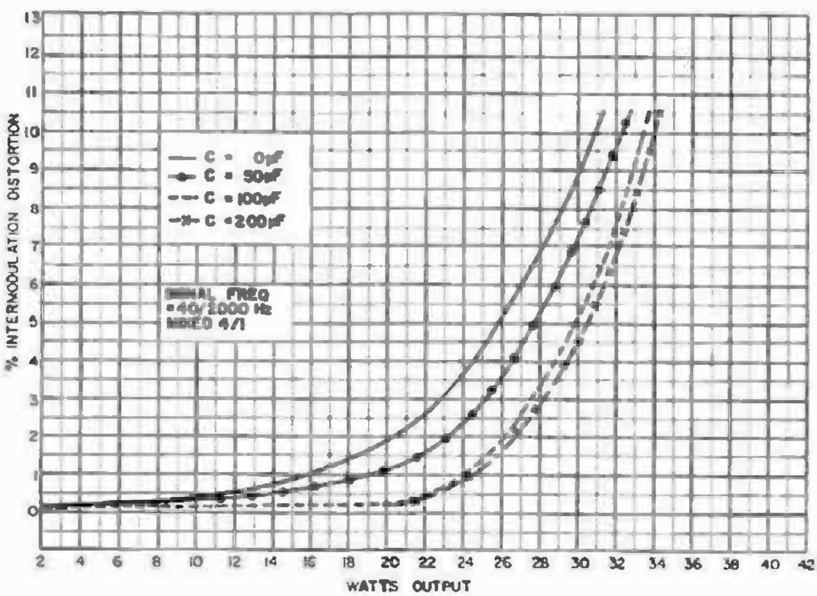


Fig. 12-106. Effect of cathode-bypass capacitor as intermodulation distortion in a Williamson ultralinear amplifier.

able to bypass the bias resistor with a large capacitor of 40 to 100 μF. However, one of 500 to 1000 μF is often used. As the voltage across a cathode resistor rarely exceeds 50 volts, these are not impractical values. The effects of a cathode-bypass capacitor relative to the intermodulation distortion in a Williamson amplifier using a pair of well-balanced tubes is shown graphically in Fig. 12-106. Using one-percent intermodulation as a reference distortion, it will be noted the power output for one percent distortion is 16 watts without a cathode-bypass capacitor. Connecting a capacitor of 50 μF across the bias resistor increases the power output to 19 watts for one-percent distortion. Increasing the capacitance to 100 μF, the power output is increased to 24 watts for the same distortion. Increasing the capacitance to 200 μF increases the output only one-half watt for the same distortion, because the amplifier has reached its maximum power output.

If the amplifier is designed for separate bias resistors, one in each cathode, each resistor is bypassed separately. If this is not done, the output plate impedance will be increased because of degeneration. Mitchell concludes, "In a class-A amplifier, the use of a bypass capacitor across the output cathode generally reduces intermodulation distortion although it may either decrease or increase the harmonic distortion."

12.107 How is the self-bias resistor calculated for a push-pull amplifier?—By the use of Ohms law:

$$R = \frac{E_c}{I_p}$$

where,

E_c is the desired bias voltage,
I_p is the plate current for the two tubes.

The bias voltage is that required for *one tube*. If the tubes are pentodes or beam-power tubes, the screen-grid current must be included in the total current flow through the bias resistor.

For calculating the value of a bias resistor for beam power, tetrode, and pentode tubes, the minimum plate and screen currents as stated by the manufacturer are used. For fixed-bias applications see Question 12.229.

12.108 If a phase-splitter is used for driving a push-pull stage, how critical are the grid resistors and coupling capacitors?—The grid resistors should be within one percent and the coupling capacitors as close to the required values as possible. If the two coupling capacitors vary in capacitance in opposite directions, at frequencies below 50 Hz the capacitive reactance variation with frequency will be considerable. Thus, the two voltages produced by the phase-splitting circuit will not be of the same amplitude at the lower frequencies, resulting in an unbalance and an increase in harmonic distortion. Capacitive reactance balance at higher frequencies is not important, as reactance of the capacitors is small; however, if excessive stray capacitance exists in the wiring, it may cause unbalance and subsequent distortion.

12.109 What are the impedance relations of a push-pull output transformer relative to the plate resistance of the tubes?—Because the tubes in a push-pull amplifier are effectively in series and parallel with the primary of the output transformer, the plate-to-plate impedance of the transformer must be twice that of the plate resistance of the tubes. (The above statement applies to triodes only.)

It appears that each tube in a push-pull amplifier sees only one-fourth its plate resistance on the basis of turns ratio. However, a push-pull transformer is a three-winding device and the reaction of one-half the primary winding to the other half is such as to cause each tube to see one-half the plate-to-plate impedance. Removing one tube from the circuit will result in the remaining tube seeing one-fourth the plate impedance.

12.110 What is the equation for calculating the plate-to-plate load impedance for a push-pull amplifier?—The two tubes in a push-pull amplifier may be considered to be two generators in series working into a load impedance of equal values. (See Fig. 12-110.) The

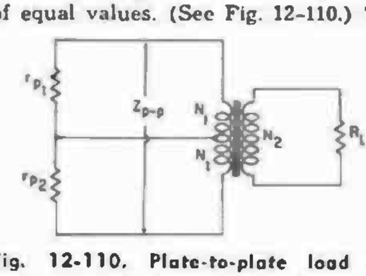


Fig. 12-110. Plate-to-plate load impedance equivalent circuit of a push-pull amplifier stage.

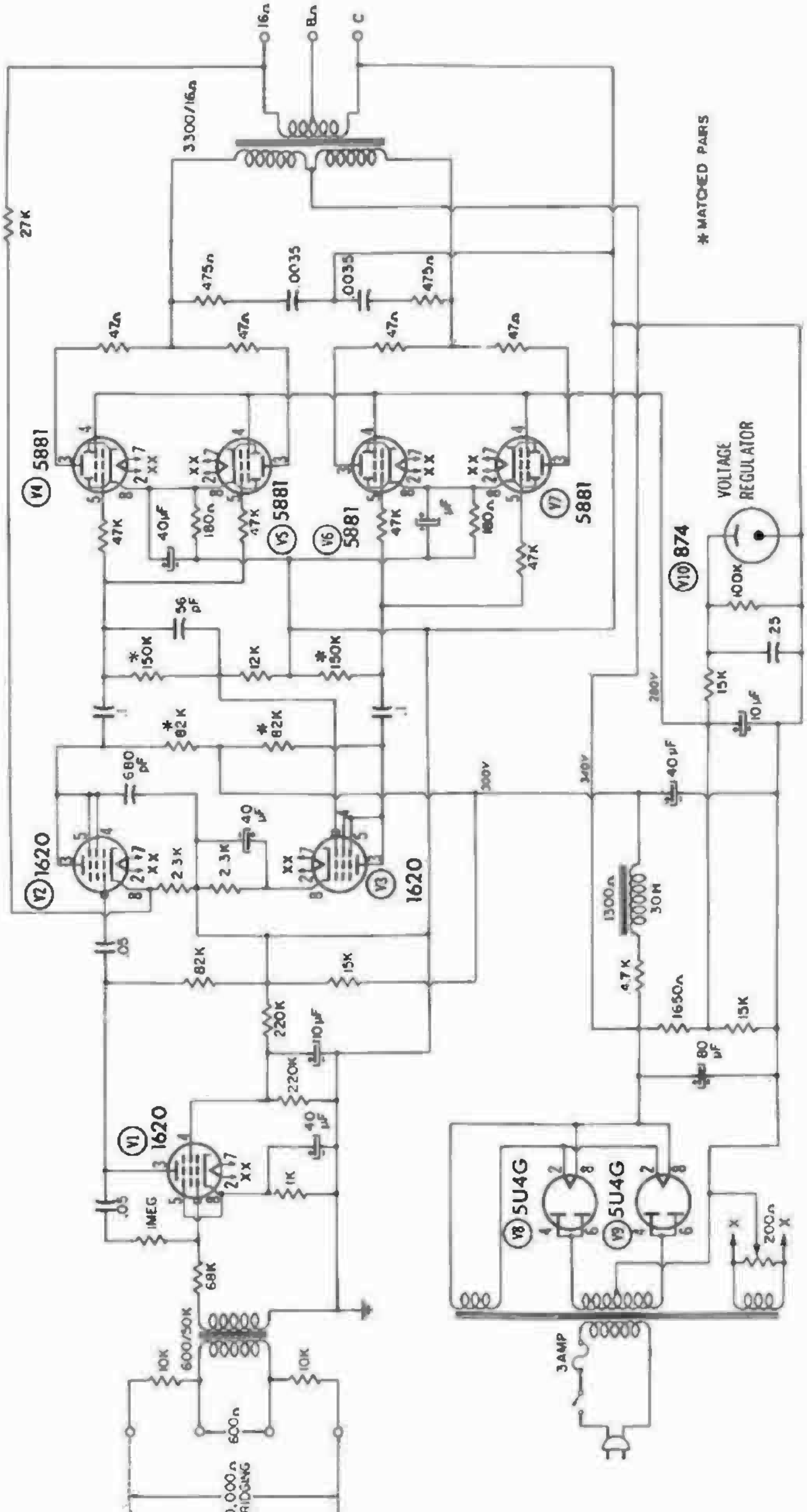


Fig. 12-111. Schematic diagram of an amplifier using push-pull parallel output.

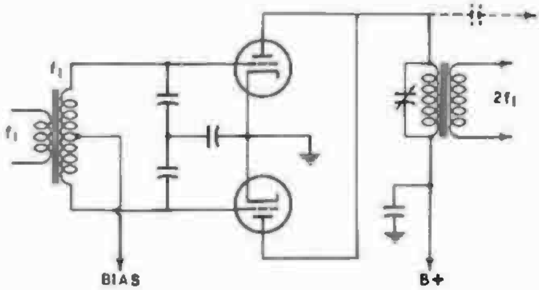


Fig. 12-112. A push-push frequency-doubling circuit. The plate circuit is resonated to the second harmonic of the fundamental frequency.

plate-to-plate impedance is equal to the reflected secondary load impedance. The primary winding is continuous and tapped at its exact electrical center. The plate-to-plate impedance may be calculated:

$$Z_{p/p} = \left(\frac{2N_1}{N_2}\right)^2 R_L$$

or

$$\left(\frac{N_1}{N_2}\right)^2 R_L = \frac{r_p}{2}$$

where,

$Z_{p/p}$ is the reflected load impedance seen by the plates of the tubes,
R_L is the external load impedance,
N_1 is the number of turns in one-half the primary winding,
N_2 is the number of turns in the secondary winding,
r_p is the plate resistance of each tube.

12.111 What is a push-pull parallel amplifier?—A push-pull amplifier in which two tubes are connected in parallel in each side of the circuit as shown in Fig. 12-111. For this type operation, the impedance of the output transformer primary, plate-to-plate, is one-half that normally used for two tubes. The power output is increased approximately six times that of a single tube.

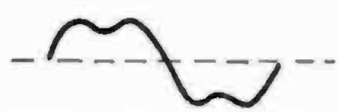
To prevent parasitic oscillation, a 47-ohm resistor is connected in the plate circuit of each tube and a 47,000-ohm resistor is placed in series with each control grid. The plate currents of each tube should be approximately the same to secure the lowest distortion. The amplifier shown will produce 30 watts of power with approximately 0.75 percent rms distortion.

12.112 What is a push-push amplifier?—An amplifier with an input circuit similar to a push-pull amplifier. The control grids are operated 180 degrees out of phase. The plates are connected in parallel. It is used for frequency doubling.

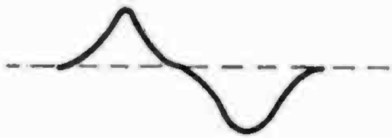
The amplifier is operated class-C. The plate circuit is tuned to the second harmonic of the frequency applied to the control grids. The bias is adjusted for a maximum output. (See Fig. 12-112.)

12.113 What are the phase shift characteristics of an amplifier?—Phase shift is a problem that exists in almost any equipment that must pass complex waveforms, and amplifiers are no exception. In modern wide-range high-quality amplifiers, phase shift is a very important characteristic. Phase shift as discussed here should not be confused with phase reversal discussed in Question 12.97 and Question 11.60.

If one frequency component of a complex waveform takes longer to pass through an amplifier than another, a time delay or displacement occurs and the output waveform is not an identical reproduction of the input waveform. This is phase distortion. In general, phase distortion causes a loss of articulation and a decrease in the intelligibility of speech.



(a) *Fundamental and third harmonic.*



(b) *Distorted waveform.*

Fig. 12-113A. The appearance of a fundamental frequency and its third harmonic after being distorted by passing through an amplifier with phase shift.

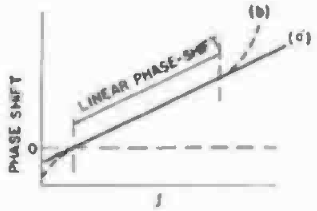


Fig. 12-113B. Phase-shift characteristics of an amplifier. (a) Constant phase shift (ideal). (b) With phase shift.

A signal consisting of a fundamental frequency and its third harmonic is shown at part (a) in Fig. 12-113A. Although the amplitudes of both components are increased in identical ratios as they pass through the amplifier, the resultant waveform at the output, shown at (b) is considerably different from that of the input, because the phase of the third harmonic has been shifted with respect to the fundamental frequency. Phase distortion is introduced in amplifiers because of the reactance in the circuits, and the type coupling used between the stages. A typical phase-shift characteristic is shown in Fig. 12-113B.

In an amplifier using negative feedback, unwanted phase shift may be caused by the components in the feedback loop or the output transformer. The measurement of phase shift is discussed in Question 23.111.

12.114 What are the effects of nonlinearity in an amplifier?—Nonlinearity in an amplifier is generally the result of operating the amplifier tubes in the nonlinear portion of the characteristic curve (Fig. 12-114A). Operating vacuum tubes in the nonlinear portion of the characteristic curve causes generation of unwanted harmonic distortion and intermodulation products, and as the harmonics are not a part of original program material, they cause distortion of the output signal waveform.

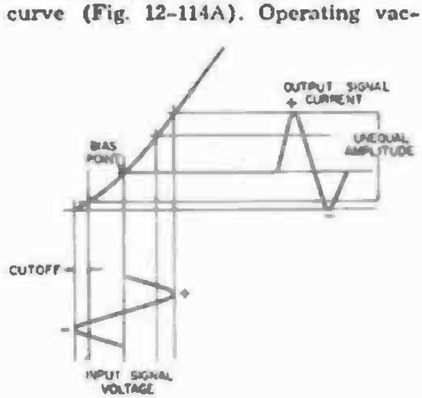


Fig. 12-114A. Dynamic characteristics of a vacuum tube incorrectly biased and operating on the nonlinear portion of its dynamic transfer characteristic.

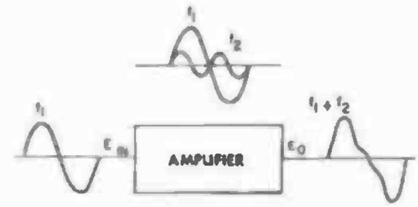


Fig. 12-114B. An amplifier with a fundemental frequency f_1 applied to the input, and a second harmonic f_2 generated within the amplifier because of internal distortion. The output waveform is distorted because it contains both the frequencies f_1 and f_2.

As an example: If a fundamental frequency f_1 (Fig. 12-114B) is applied to the input of an amplifier, and because of its nonlinearity characteristic it generates a strong second harmonic f_2 and the output waveform will be distorted. This is caused by the combining of the generated second harmonic with the fundamental frequency.

Fig. 12-114C is a plot showing the characteristics of a linear and nonlinear amplifier. It will be noted for the linear plot (a) the output voltage rises directly in proportion to the amplitude of the input signal, resulting in a straight line between input and output voltages.

Plotting the input signal amplitude versus the output voltage for a nonlinear amplifier results in a characteristic as shown by the dashed line (b). As may be seen, the output voltage does

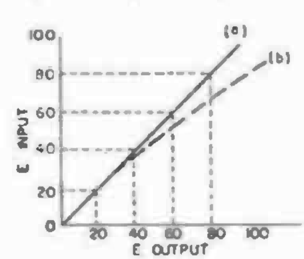


Fig. 12-114C. Input and output characteristics of an amplifier with a linear response (a) and one with a nonlinear response (b).

not rise in the same proportion as the input signal. Although the nonlinear characteristic is shown as a straight line, it will generally bend near the upper limits.

The measurement of nonlinearity in amplifiers is discussed in Question 23.94.

12.115 ***What is frequency distortion in an amplifier?***—If the gain frequency characteristics of an amplifier are not uniform, certain frequencies will be amplified to a greater or lesser degree causing the ratio between frequencies to be changed from the original waveform ratios. Thus, when frequencies passed through an amplifier are not amplified with their original amplitude ratios, this is called frequency distortion. If the high frequencies are amplified excessively, the harmonic distortion will be increased. (The foregoing statement does not apply to an equalized amplifier.)

12.116 ***Define the term intermodulation distortion.***—In amplifiers and other electrical devices designed to pass a multiple frequencies, sum and difference frequencies may be generated inducing intermodulation distortion. Intermodulation distortion may be defined as the production in a nonlinear circuit element of frequencies corresponding to the sum and differences of the fundamentals and harmonics of two or more frequencies which are transmitted through that element.

Designating two fundamental frequencies f_1 and f_2, the intermodulation products will consist of second and third order terms. In general none of these terms is harmonically related to either of the fundamental frequencies.

Intermodulation distortion will manifest itself as a harsh or rough unpleasant reproduction, or by a buzz at the higher frequencies. This type of distortion is not directly related to harmonic distortion and cannot be measured using a conventional harmonic analyzer or meter. Therefore, special intermodulation analyzers have been developed for this express purpose. (See Questions 22.129 and 23.117.) To date, no significant psychological tests have been made to determine just how much intermodulation distortion can be tolerated by the human ear. (See Question 1.118.) However, it is sufficient to say that amplifier systems having a low order of intermodulation distortion sound much cleaner than an amplifier of comparable harmonic distortion. It is generally agreed that sound recording and reproducing systems must have less than 1-percent intermodulation distortion.

Other forms of distortion aside from that just mentioned are harmonic, phase-shift, frequency, and noise. Distortion may also be caused by magnetic fields from power transformers and chokes, and the overloading of the output transformer which causes saturation of the core material.

12.117 ***What is random noise?***—Spurious frequency signals which are amplified with the program material. Random noise is caused by thermal agitation produced by the movement of electrons in a material. This movement of electrons causes minute pulses which are intermixed with the signal voltage and amplified. Random noise covers the entire frequency band and is the limiting factor in the design of a low-level, high-gain amplifier.

12.118 ***What is the shot effect?***—Noise in a vacuum tube caused by the movement of electrons between the cathode and plate elements. This effect causes unsteadiness in the plate current and is amplified with the signal voltage.

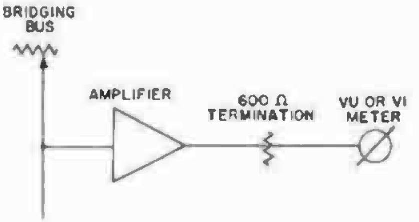


Fig. 12-119. VU or VI meter with an amplifier connected ahead of the meter to increase the voltage level.

12.119 ***Describe the use of a VU or VI meter amplifier.***—When the voltage level of a sound circuit is too low for the use of a standard VU meter, an amplifier may be connected ahead of the meter, and the signal amplified before it is applied to the meter (Fig. 12-119).

The disadvantage of this method is *gain variations* in the amplifier caused by tube variations and operating voltages. However this may be overcome to a great extent by the use of a semiconductor amplifier highly stabilized by the use of a generous amount of negative feedback and regulated operating voltages. Even then, the calibration should be frequently checked if the meter is connected in a critical part of the system. Whenever possible, the meter (without amplifier) should be connected across a circuit impedance of 600 ohms and of the correct voltage level. The frequency response of the amplifier must be uniform across the frequency band of interest. (See Question 12.196.)

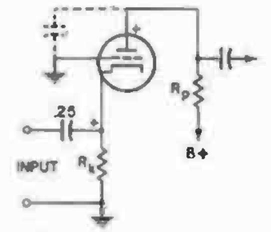
To calibrate the meter with the amplifier, a plus 4-dBm signal is applied to the amplifier input and the amplifier gain adjusted for a 100-percent reading on the meter (100 percent on a standard meter is plus 4 dBm), with the meter attenuator set to plus 4 dBm. (See Questions 10.17 and 10.24.)

12.120 ***What is a buffer amplifier?***—A stage of amplification between an oscillator or other device to isolate it from succeeding stages.

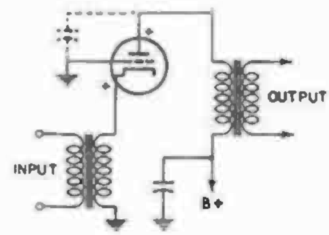
12.121 ***What is a frequency-doubler stage?***—An amplifier stage having a resonant circuit in its plate circuit tuned to a harmonic of the fundamental frequency at the input. They are used in radio transmitting equipment and in test equipment. A push-push amplifier is a frequency doubler. (See Fig. 12-112.)

12.122 ***What is a grounded-grid amplifier?***—A vacuum tube connected as shown in Fig. 12-122. The incoming signal is applied to the cathode across a low resistance while the output signal is taken across a high-resistance load in the plate circuit. The control grid is tied directly to ground. The input and output signals are in phase.

Grounded-grid amplifiers are often used to couple a low-impedance source to a high-impedance load, thus providing gain as well as an impedance match.



(a) *Resistance-coupled.*



(b) *Transformer-coupled.*

Fig. 12-122. Grounded-grid amplifiers.

The theoretical gain of a grounded-grid amplifier may be computed:

$$V_g = (1 + \mu)\frac{Z_2}{r_p + Z_2}$$

where,

r_p is the plate resistance,
Z_2 is the load impedance including the capacity between the grid and plate, neglecting the capacity between plate and cathode.
μ is the amplification factor of the tube.

In some respects it may be said a grounded-grid is the reverse of a cathode-follower amplifier.

12.123 ***What is a grounded-plate amplifier?***—This is another name of a cathode-follower amplifier as described in Question 12-41.

12.124 ***What is a gas-tube coupled amplifier?***—An amplifier using a gas-filled tube as a coupling medium be-

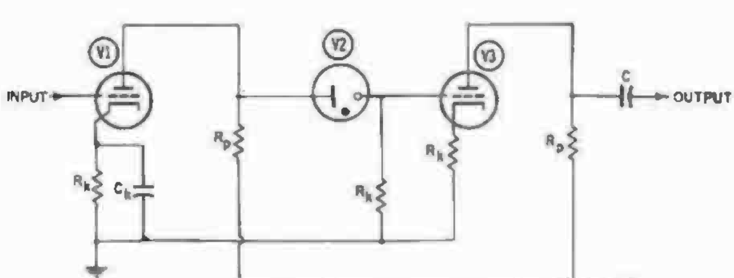


Fig. 12-124. A two-stage, RC amplifier using a gas-filled tube in place of the usual coupling capacitor.

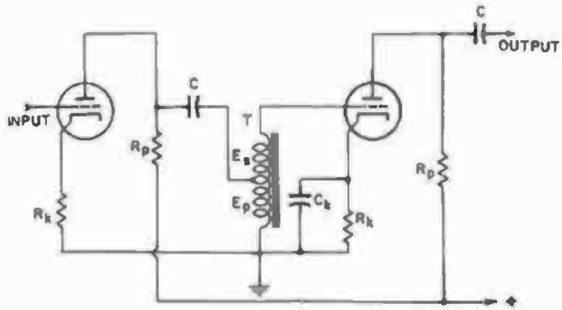


Fig. 12-125. A two-stage, autotransformer-coupled amplifier.

tween stages. A typical circuit is shown in Fig. 12-124. The first stage is a conventional amplifier with a resistive plate load. Instead of the usual coupling capacitor, the control grid of the second stage is fed from the plate of the first tube through a small neon tube. Voltage-regulator tubes, as a rule, are not satisfactory because of their high operating potentials. The purpose of such coupled amplifiers is to reduce the phase shift generally associated with resistance-coupled amplifiers at the extremely low frequencies. Gas coupling is only practical at fairly high signal levels because of the high internal noise level of a gas tube. The frequency response and stability may be improved by the use of negative feedback by eliminating the usual cathode-bypass capacitors.

12.125 What is an autotransformer-coupled amplifier?—An amplifier as shown in Fig. 12-125 which uses an autotransformer rather than the conventional two-winding coil to step up the voltage between stages. The autotransformer is parallel-coupled to the plate of the first tube to eliminate the flow of direct current through the winding. There is no particular advantage to autocoupling, therefore it is seldom used.

12.126 What is a series or single-ended push-pull amplifier?—An output stage in which two tubes are connected in series across the direct-current plate-voltage supply, as shown in Fig. 12-126. The output tubes are driven from a phase-inverter stage which produces two voltages of equal amplitude but 180 degrees out of phase.

The unique feature of this phase-inverter circuit is that it drives each tube from its own control grid to its own cathode circuit, driving the tubes in a balanced fashion.

The advantages of this circuit are that it has a single-ended output and is capable of driving a two-terminal load such as a loudspeaker voice coil, and requires no output transformer. The advantages of eliminating the output transformer are manifold; also, large amounts of negative feedback may be applied with a high degree of stability, enabling very low distortion characteristics to be obtained. With modifications this circuit is used extensively in transistor output stages.

12.127 What is a cathamplifier?—A push-pull amplifier circuit in which

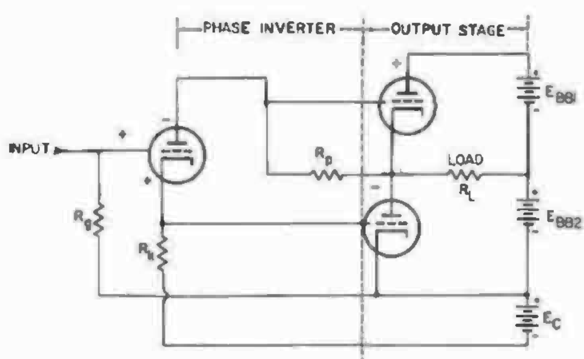


Fig. 12-126. Basic circuit of a single-ended push-pull amplifier.

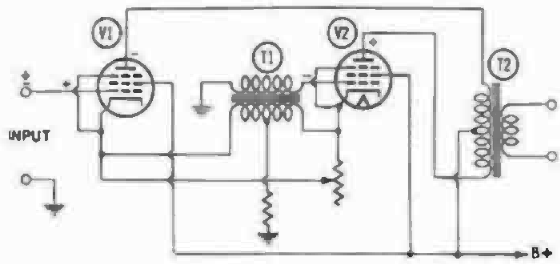


Fig. 12-127. A push-pull cathamplifier.

the push-pull input transformer is connected to the cathodes, as shown in Fig. 12-127. The input signal is applied directly to the control grid of tube V1 and to the secondary of transformer T1. Transformer T2 is a conventional push-pull output transformer.

12.128 What is an ultralinear amplifier?—An amplifier circuit designed for high quality sound reproduction which was developed by Hafler and Keroes and employs a special output transformer. (See Fig. 12-128A.)

When a pentode tube is operated with the screen grid and the plate elements at the same positive potential, pentode operation is obtained. If the screen grid is connected to the plate element and fed from the same end of the output transformer, triode operation is obtained.

In the ultralinear method of operation, the screen-grid element is returned to a tap on the primary of the output transformer at a point about 18.5 percent of the primary impedance (measured from the center tap of the primary winding). This provides an intermediate operating point between a pentode and a triode, with the power output of a pentode and the low internal output impedance of a triode. Under the above operating conditions, the tube functions as a pentode with negative feedback applied to the screen grid, resulting in lower distortion products at the higher levels of operation. When the

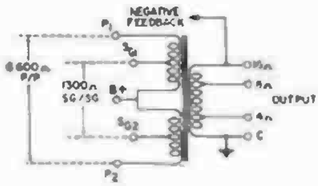


Fig. 12-128A. An ultralinear output transformer designed for use with 5881 or 6L6 beam-power tubes.

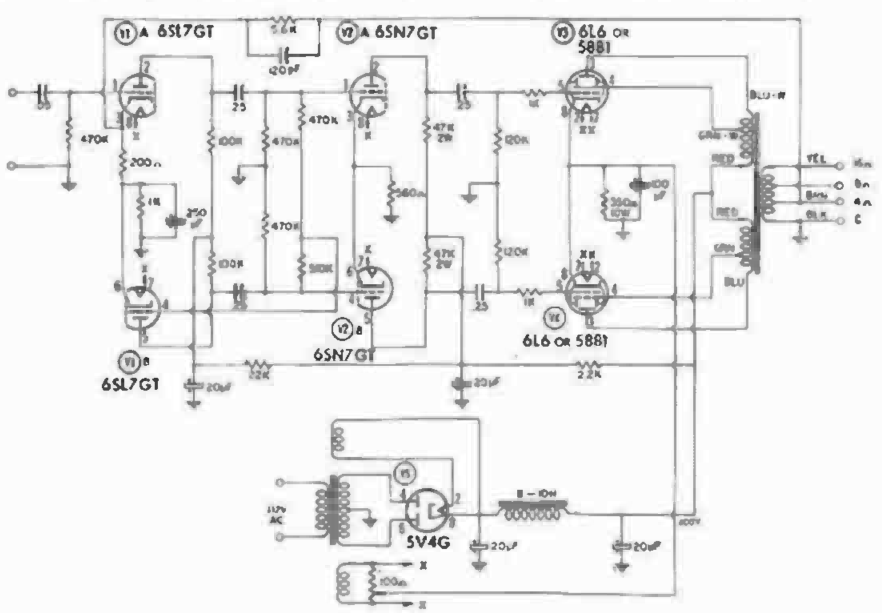


Fig. 12-128B. An ultralinear amplifier using 6L6 or 5881 tubes.

ultralinear circuit is used in conjunction with the well-known Williamson amplifier circuit (Fig. 12-232A), a power output of over 20 watts is possible from a pair of 5881 tubes, with less than one percent intermodulation distortion.

The impedance of the output transformer for ultralinear operation is shown in Fig. 12-128A. The impedance from plate-to-plate is 6600 ohms and from screen grid to screen grid is approximately 1300 ohms.

Referring to Fig. 12-128B, negative-feedback voltage is taken from the secondary winding of the output transformer and carried back to the cathode circuit of the first tube, through a 5600-ohm resistor shunted by a 120 pF capacitor. The capacitor assists in maintaining a smooth frequency response above 100 kHz and also helps maintain a constant 180-degree phase shift at the higher frequencies. The total feedback is 20 dB. The frequency response has been extended to more than two octaves above the usual 20 to 20,000 Hz. This wide frequency response is necessary to maintain a constant phase shift and improve transient response.

The weakness of the amplifier lies in the fact that its *negative feedback becomes a positive feedback at ultrasonic and subsonic frequencies*, leading to a considerable amount of distortion at the lower frequencies, unless it is very carefully designed. The amount of negative feedback applied to the first stage should not exceed 20 dB.

Tests for oscillation at inaudible frequencies may be made by the application of a square wave to the input while observing the output waveform on an oscilloscope.

In some circuit designs of the Williamson amplifier using the ultralinear output stage, high-frequency oscillation will occur when a capacitive load such as an electrostatic loudspeaker is connected across the output, or the leads of the loudspeaker have considerable distributed capacity. The measurement of capacitive load effects is discussed in Question 23.125.

12.129 What is a Circlotron amplifier?—An amplifier circuit developed by Wiggins, and manufactured by Electro-Voice Inc., using a specially designed output stage employing two 6V6 beam power tubes which develop an unusually large amount of power with low distortion. One of the fundamental requirements of a high-quality amplifier is that it have an output transformer with negligible leakage reactance. This leakage reactance must be low to avoid the transient distortion resulting from collapsing currents such as those encountered in class-AB or -B push-pull amplifiers when the tubes are driven beyond cutoff. Transient distortion appears as a parasitic oscillation in the waveform at the instant of cutoff. A high value of leakage reactance will also cause the output transformer of a conventional amplifier to lose efficiency at the high frequencies; therefore, the distributed capacitance of the output transformer should be very low, in order to minimize high-frequency attenuation and phase shift. It is claimed for the circuit of Fig. 12-129A, a Wiggins Circlotron amplifier, that this objectionable feature of the conventional amplifier is overcome.

Fig. 12-129B is a simplified version of the above amplifier. Two power supplies are indicated as batteries. Each power supply is connected from the plate of one tube to the cathode of the other. The plate current of each tube circulates through both power supplies without traversing the windings of the output transformer. Because any pair of opposite points in this configuration is equipotential, the circuit is a balanced bridge under no-signal conditions.

The total primary impedance of the output transformer presents a load to each of the two output tubes. One-half of this load is in the cathode circuit and the other half is in the plate circuit. The plate load of one tube is the cathode load of the other. Because each tube looks into the same load as the other, the result is unity coupling between the tubes. Despite the residual leakage reactance in the transformer, no switching transients can occur during the operation of the amplifier, for both halves of the transformer primary winding have the same signal current flowing in them; thus, switching transients are eliminated

The impedance of the primary winding is one-fourth that of the conventional amplifier; therefore, the distributed capacitance is less. Also, the leakage reactance makes a wide frequency range more easily attained. The quiescent current being low results in higher efficiency and produces more power without exceeding the dissipation rating of the tubes.

The gain of the Circlotron output stage is approximately unity; thus, it requires a high signal driving voltage. This higher signal voltage is obtained by means of a technique called "bootstrapping." By this method, the B-plus supply voltage to the driver stage is dynamically changed as the signal volt-

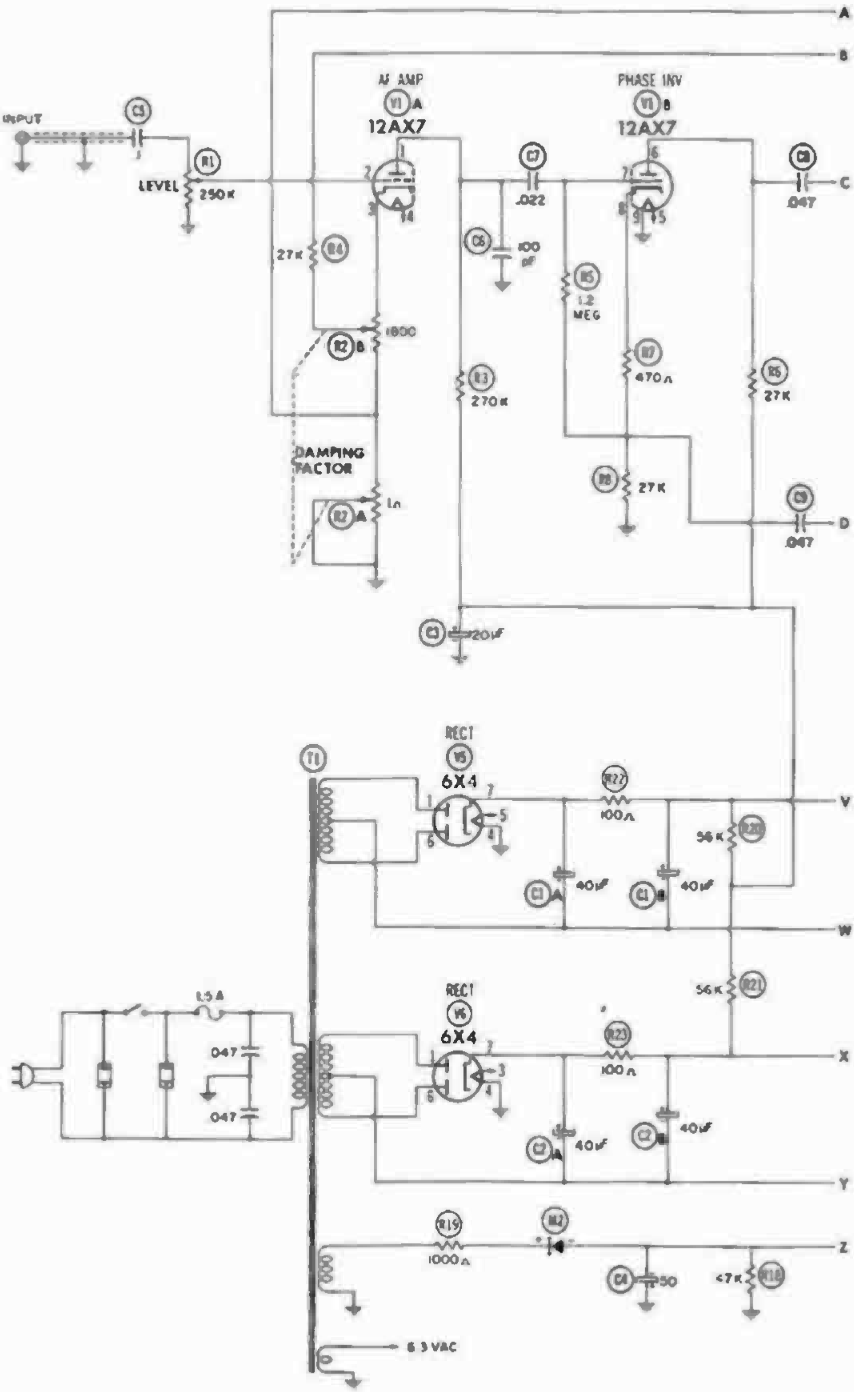
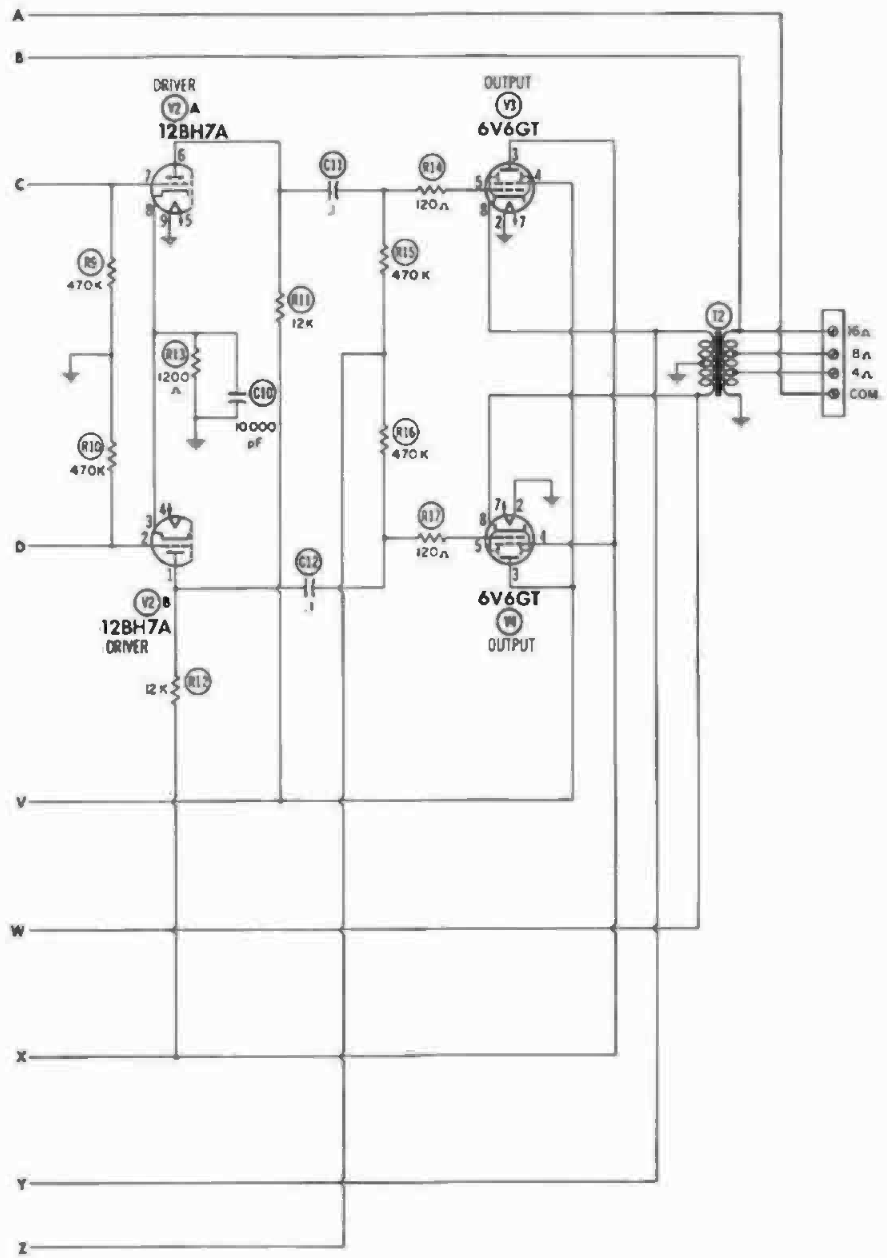


Fig. 12-129A. Wiggins Circlotron

age changes. This permits linear operation over a wide range.

Variable damping is also provided in the output to present a means of matching the impedance characteristics of the loudspeaker system to the amplifier at the lower frequencies. The values of the critical damping resistance vary widely with different type speakers and are dependent upon the flux density, type of enclosure in which the speaker is mounted, the length of the conductor



20-watt power amplifier.

in the air gap, and, to some extent, the position of the speaker enclosure in the room.

The variable-damping control circuit is so designed that varying amounts of reactive voltage and current feedback are combined to match the effective impedance while maintaining the total feedback at a constant value. (Variable damping is discussed in Question 12.242.) The maximum power output available is independent of the damping factor and remains constant for all settings of the damping control. Referring to Fig. 12-129A, the incoming signal is applied to the control grid of V1A which is coupled to a split-load phase splitter V1B which, in turn, drives a push-pull driver stage V2A and V2B. The output stage V3 and V4 is cathode coupled to the external load circuit by means of transformer T2. The primary of the output transformer is connected to the cathodes of the output tubes and center tapped to ground. A fixed bias voltage is applied to the control grid of the output tubes. This is supplied by a half-wave rectifier M2. (Fixed bias circuits are discussed in Question 12.229.) Negative-feedback voltage is obtained by taking voltage from the 16-ohm tap of the output winding and returning it to the potentiometer R2A in the cathode circuit of V1A. Negative-current feedback is obtained by connecting the ground side of the output winding to the junction of R2A and R2B in the cathode circuit of V1A. These two controls are mechanically connected and operate as a unit.

The power output is rated at 20 watts (40 watts on peaks), at less than 0.5-percent harmonic distortion, or less than one-percent intermodulation. Hum and noise are rated 85 dB below 20 watts or minus 42 dBm. The feedback is 19 dB in the output circuit, 16 dB in the feedback loop, and 2 dB positive feedback in the driver circuits, making a total of 33 dB total negative feedback. The frequency response is plus or minus 1 dB from 20 to 60,000 Hz at 20 watts output. The damping factor is adjustable from 0.1 to 15. The input impedance is 250,000 ohms with a 0.1-μF isolation capacitor in the input.

12.130 What is an extended class-A amplifier?—A push-pull amplifier employing a triode and a beam-power tube in parallel on each side of a push-pull circuit. (See Fig. 12-130.) At low levels the amplifier operates as a triode amplifier with the beam-power tubes at cutoff. At the higher levels, the output is almost entirely from the beam-power tubes. The plate dissipation is about one-third that of a normal class-A amplifier for the same power output.

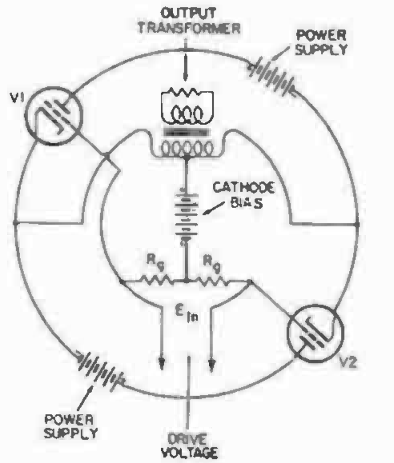


Fig. 12-129B. Basic circuit for Wiggins Circlotron power amplifier.

In the circuit shown, a power output of 47 watts may be obtained with low distortion. The input signal may be supplied from either a transformer or a phase splitter.

12.131 Are the frequency characteristics of a power amplifier the same at all power levels?—Unless the output transformer is of unusually good design, frequency response will show a loss at both the high and low ends at higher output levels. Typical power curves are shown in Fig. 12-131.

12.132 Describe an output transformerless amplifier (OTA).—It is an amplifier that does not use an output transformer. The external load circuit is generally connected to the cathode of the output tubes, through a 100- to 1000-μF capacitor, to the external load circuit, such as a voice coil of a loudspeaker. The capacitor is only required if the external load circuit will cause a current flow between the output point and ground. A typical OTL circuit is shown in Fig. 12-132.

12.133 What is a preliminary amplifier?—This is another name for a preamplifier. (See Question 12.72.)

12.134 What is a dielectric amplifier?—An amplifier requiring no vac-

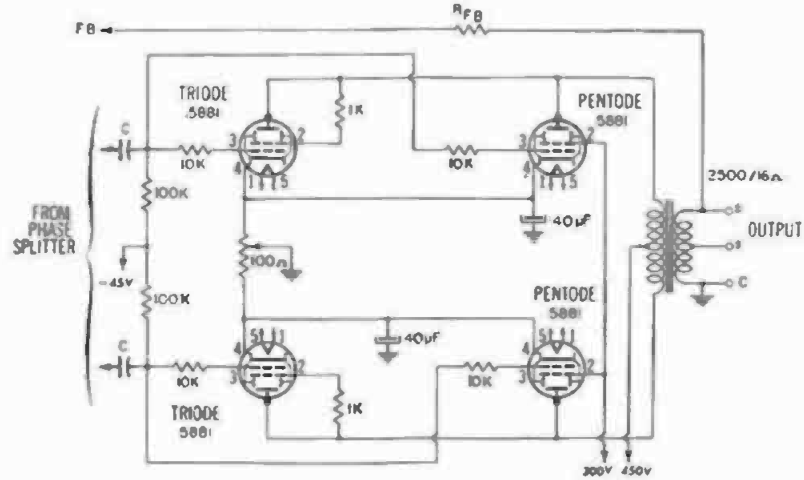


Fig. 12-130. A push-pull parallel, extended class-A amplifier using a combination triode and pentode connection.

uum tubes but employing a device similar to a capacitor using polycrystalline dielectric which exhibits a ferromagnetic effect. Ceramics using barium titanate, barium zirconate, strontium titanate, and lead zirconate are among the materials used. The device has a frequency range running into the megahertz. Its characteristics are nonlinear. It is caused to operate on the linear portion of its characteristic curve by the use of a dc bias voltage in conjunction with a high-frequency power supply. This type amplifier is used in computers and similar devices.

12.135 What is a program amplifier?—An amplifier connected in the main program channel of a broadcast station speech-input system. It must be capable of developing an output level of at least plus 18 dBm. It is also called a line amplifier.

12.136 What is a negative-feedback or degenerative amplifier?—An amplifier in which a portion of the output signal voltage is fed back 180 degrees out of phase to the input. Two types of negative feedback are in common use—voltage and current.

12.137 Describe a nyquist curve.—It is a graphical method, devised by H. Nyquist of the Bell Telephone Laboratories, for plotting the relationship between amplification and feedback in an amplifier employing either negative or positive feedback. Such curves are generally plotted in polar form and are used as a criteria for determining the stability of an amplifier to which feed-

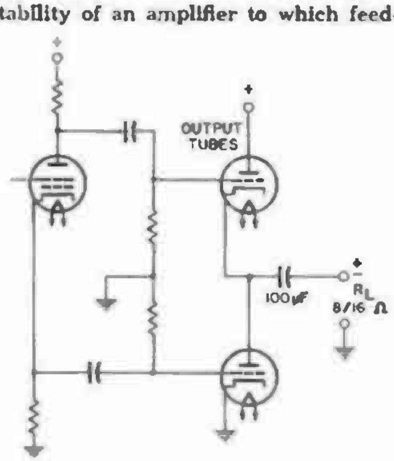


Fig. 12-132. Output stage for transformerless (OTL) amplifier.

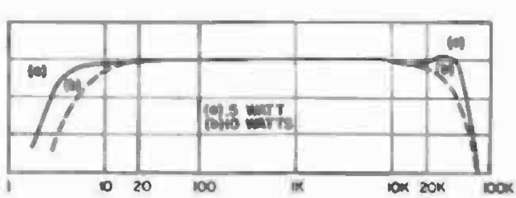


Fig. 12-131. Frequency response of a wide range high-quality audio amplifier at 0.5 and 10 watts output.

back is applied. Nyquist curves can also be used for plotting oscillators and other circuit data critical to stability. A typical Nyquist plot is shown in Fig. 12-137. The reader is referred to the references at the end of this section.

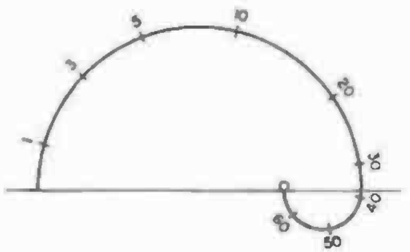


Fig. 12-137. Typical Nyquist plot used for determining the stability of an amplifier or other device.

12.138 What are the advantages of using negative feedback in an amplifier? —Negative feedback lowers the harmonic and intermodulation distortion components(assuming the amplifier does not have excessive distortion before the negative feedback is added); the frequency response may be made quite uniform over a very wide range; the stability may be increased and is less subject to variation in tubes and operating voltages; lower hum and internal noise; constant gain for a given response; lower internal output impedance providing a high factor of internal output damping; and less phase shift.

12.139 Explain how a negative-feedback amplifier functions.—In the conventional audio amplifier, the internal resistance of the output tube shunts the external load impedance. When the plate resistance is less than the load impedance, variations in the load impedance have very little effect on the output voltage, because the load-impedance variations are shunted by the low resistance of the output tube. This indicates that a tube of low plate resistance is desirable in the output stage.

When the plate resistance is high (as is the case of a pentode or beam-power tube), the effect of a variable load impedance is to increase the distortion components, producing strong peaks, and thus impairing the quality of reproduction. Amplifiers employing pentodes or beam-power tubes in the output stage may make use of a resistance-capacitance network in the plate circuit as described in Question 12.169 to compensate for the effects of the variable load impedance. However, using this method, some loss of power is to be expected, particularly at the higher frequencies.

Amplifiers employing pentodes or beam-power tubes may be designed to have the characteristics of a low plate resistance tube, yet have the advantages of the increased power sensitivity afforded by the pentode. This advantage is obtained by the use of negative feedback in the output stage. The addition of uegative feedback is accomplished at the expense of sacrificing a portion of the amplifier gain. However, the benefits gained by the use of negative feedback offset the loss of gain in the improved performance of the amplifier. The loss of gain may be compensated for by increasing the gain in the voltage-amplifier section. A typical two-stage RC amplifier employing negative feedback is shown in Fig. 12-139A. The feedback voltage is obtained by feeding back to the input of the amplifier a portion of the output voltage in such a

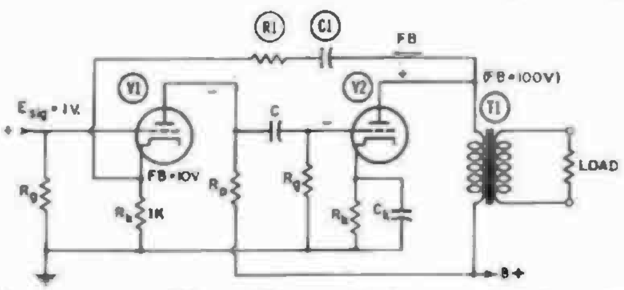


Fig. 12-139A. A two-stage RC amplifier employing negative-voltage feedback. In this circuit the characteristics of the output transformer are not included. To include the output transformer, the feedback voltage is taken from the top of the transformer secondary. The lower end is returned to ground (feedback voltage must be of the correct polarity.)

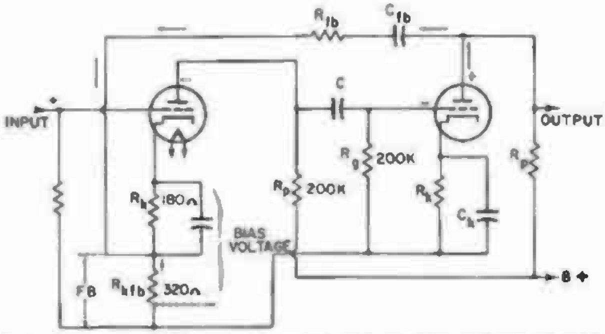


Fig. 12-139B. A two-stage RC amplifier using negative-voltage feedback from the plate of the output tube to a portion of the cathode resistance of the first stage.

phase relationship as to oppose the input signal. Assume an instantaneous positive pulse of 1 volt is applied to the control grid of tube V1. At the plate will appear a negative pulse because of the 180-degree phase reversal within the tube as described in Question 11.60. This negative pulse is transferred to the control grid of V2 through the coupling capacitor C, causing a decrease of plate current in V2. The signal is again amplified, reversed 180 degrees, and appears as a positive pulse at the plate of V2. This pulse is transmitted through capacitor C1 and resistor R1 in the negative-feedback loop to the upper end of the cathode resistor R_k of V1, applying a positive pulse to the cathode. This effect is the same as biasing the control grid more negatively. When the feedback voltage increases, the gain of V1 is decreased. As an example: with 100,000 ohms in the feedback loop and 1000 ohms in the cathode circuit of V1, a voltage divider is formed with the cathode of V1 at the junction of the two resistors (for this illustration it will be assumed the reactance of capacitor C1 in the negative-feedback loop is negligible).

With a signal voltage of 100 volts at the plate of V2, 10 volts will be fed back to the cathode of V1. Now to obtain 100 volts of signal at the plate of V2 will require the signal voltage at the input of V1 be increased to 11 volts. In other words, the signal voltage must be increased the amount of the feedback voltage in the cathode circuit plus the original signal voltage at the control grid of V1. By definition, the foregoing illustration is called 10-percent negative feedback.

Negative feedback may also be applied as shown in Fig. 12-139B. Here the feedback voltage taken from the plate of V2 is applied to only a portion of the cathode resistor R_{KFB} of V1. Under these conditions the dc bias voltage for the tube is the voltage drop across R_K and R_{KFB}, while the negative-feedback voltage is formed across only the lower portion of the resistor. As the feedback voltage is confined to the lower portion of the cathode resistor, the upper portion is used only for bias voltage and may be bypassed to prevent current feedback, thus increasing the actual gain of the stage. This system of applying negative feedback permits the correct value of feedback voltage to be applied to the tube, while maintaining the correct value of resistance for the cathode circuit.

Gain controls cannot be connected in the area encompassed by the feedback loop, but must be removed to a position outside the feedback loop, as shown in Fig. 12-217B. Negative feedback may be utilized to increase the output impedance (negative-current feedback) or to reduce the impedance by the use of negative-voltage feedback. This subject is discussed in Questions 12.142 and 12.154.

12.140 How is harmonic distortion reduced by means of negative feedback? —All amplifiers, regardless of design, have some distortion and will have some component in the output signal that was not present in the input signal. The feedback voltage taken from the output of the amplifier contains both the original signal and the distortion component added by the amplifier. That part of the feedback voltage consisting only of the original signal will be cancelled at the input by the applied signal voltage. However, any distortion originating within the amplifier, when fed

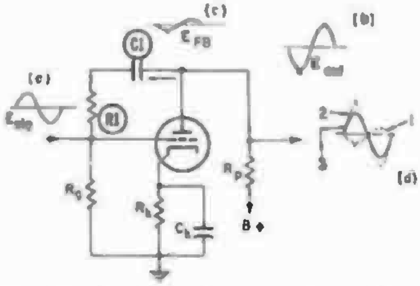


Fig. 12-140. Waveforms indicating how negative feedback tends to cancel distortion generated within an amplifier.

back will not be cancelled by anything present in the input, since the distortion is not present in the input but is developed within the amplifier. Therefore, the distortion voltage is passed through the amplifier from the input and arrives at the output with a polarity opposite to the distortion produced within the amplifier. The result is that the total distortion is reduced because it cancels itself. To better understand how the distortion is reduced by the application of negative feedback, Fig. 12-140 shows a single-stage, resistance-coupled amplifier with negative feedback, with the various voltage waveforms as they appear at different points in the circuit. At point (a) is shown the signal voltage which is of sine-wave character. It will be assumed that the signal in passing through the tube is distorted and this distortion appears as a small pip on the negative half of the output signal waveform as shown at point (b). A portion of this voltage (c) is fed back to the grid circuit. This voltage is 180 degrees out of phase with the grid voltage and has the same wave shape as the voltage at (b) but at a fraction of the output amplitude. Applied to the control grid, it acts to cancel the original distortion. At the output (d) is shown the relationship of the various waveforms. Curve (1) of waveform (d) is the plate-current component due to the feedback voltage applied to the control grid. Curve (2) of (d) is the plate-current component developed by the signal voltage at the control grid. Curve (3) of (d) is the algebraic sum of curves (1) and (2) and is the resultant plate current, clearly indicating a considerable reduction in the original distortion shown at point (b). The same principles apply to any amplifier in which negative feedback is used, including push-pull amplifiers class-A, class-B, and class-AB_1.

As a rule, negative feedback is not used with the power-output triodes such as the 2A3, because of its low plate impedance (800 ohms) and the fact that it is not too greatly affected by changes in the load impedance, such as might be encountered with a loudspeaker. However, negative feedback is highly desirable with pentodes and beam-power tubes.

12.141 How is noise reduced by negative feedback?—Any noise induced by the amplifier is fed back to the input and appears at the output with reversed polarity, effecting an overall reduction of noise. (See Fig. 12-140.)

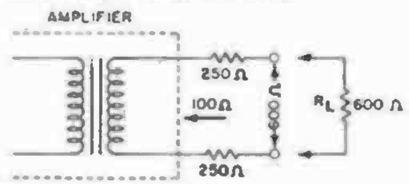


Fig. 12-142. Building-out resistors in the output of a negative-feedback amplifier, with a 100-ohm internal output impedance. The resistors build-out the impedance to 600 ohms.

12.142 Describe the effect of negative feedback on the internal impedances of an amplifier, and the use of building-out resistors.—As a majority of present day amplifiers employ a considerable amount of negative feedback to reduce distortion and noise, this lowers the internal output impedance. The effect on input impedance is not determined by the type of feedback (voltage or current), but by the method it is applied. If it is applied in series with the input-signal voltage, the input impedance is increased. Applied in shunt with the input voltage, the input resistance is decreased. *Negative-voltage* feedback reduces the internal output impedance, while *negative-current* feedback increases the internal output impedance. (See Question 12.139.)

Although the amplifier may be rated to operate with a given load impedance, the internal output impedance of the amplifier may be on the order of 100 ohms or less. This lower impedance can seriously affect the operating characteristics of equipment following the amplifier, particularly if it is a filter or an equalizer. A typical example would be the case where an amplifier is rated to operate with a 600-ohm load impedance, but has an internal output impedance of 100 ohms (in some instances it could be as low as 4 ohms). To effect a proper impedance match, series resistors are connected in each side of the output circuit (Fig. 12-142). The value of the resistors are such that they supply a match between the amplifier and the load impedance, which in the instance of a 100-ohm internal output impedance, wonld be 500 ohms divided equally between the two sides. It should be remembered, the insertion of the resistors decreases the output in the ratio of the internal output impedance to load impedance. Using the example of the 100-ohm internal output impedance terminated in 600 ohms, the ratio is 6:1. Building-out resistors are used extensively in preamplifier circuits when used in sound mixers. Referring to the graph of Fig. 5-50, the incurred loss is 3dB; therefore, if the amplifier is capable of producing plus 26 dBm for a given amount of distortion, the maximum output level for the same value of distortion, using one 250-ohm resistor in each side of the output, would be reduced to 23 dBm.

12.143 ***How does negative feedback affect the frequency characteristics of an amplifier?***—Negative feedback tends to maintain a uniform output resembling a constant-voltage generator. When the output voltage increases, the feedback voltage is increased, which has the effect of reducing the output. If the output level is reduced, the reverse takes place; thus, the output is held constant. Therefore, for a constant input of any frequency within the frequency range of the amplifier, output voltage will be constant.

12.144 ***What does the term open-loop gain mean?***—It is the gain characteristics of an amplifier before the application of negative feedback. Thus, an amplifier using 20-dB feedback could have a closed loop gain of 40 dB, but with the loop open, a 60-dB gain.

12.145 ***What are the advantages of a low internal output impedance?***—If an amplifier is used to drive an electromechanical transducer, a reduction in the internal output impedance is desirable, because a mechanical vibratory system continues to vibrate even after the exciting force is removed. In a loudspeaker, cutting head, or light modulator, this is an undesirable condition as it causes hangover effects, resulting in a fuzziness and lack of definition.

If the internal output impedance is lower than the load impedance, a damping effect is obtained and the unwanted vibrations are damped out. It is not uncommon for an amplifier having a 600-ohm output impedance to have an internal output impedance of 100 ohms or less, and for one rated 16 ohms, less than 1 ohm. In the case of a voltage amplifier (preamplifier) driving a filter or equalizer, building-out resistors, discussed in Question 9.24, may be required. (See Questions 12.142 and 20.103.)

12.146 ***What is the gain-reduction factor of a negative-feedback amplifier?***—A factor used in the design of negative-feedback amplifiers. This factor may be calculated:

$$dB = 20 \operatorname{Log}_{10} (1 - A\beta)$$

where,

A is the gain without feedback,
Beta (β) is the fraction of the output voltage fed back to the input.

Because of the gain reduction, the input-signal level must be increased by the amount of the feedback voltage to obtain the same level signal at the output obtained without feedback. (See Question 12.139.)

12.147 ***Does negative feedback affect the power output of an amplifier?***—No. However, the power sensitivity is reduced and a greater input signal is required to drive the amplifier to produce the same amount of power for a given input signal without feedback. As a rule, greater power is obtained for a given amplifier when using negative feedback because the distortion is reduced; therefore, more useful power is available.

12.148 ***What does the term decibels of feedback mean?***—It designates the amount of negative feedback induced in an amplifier, and is equivalent to the gain reduction caused by the application of the feedback. (See Question 12.139.)

12.149 ***Show an equivalent circuit for a simple negative-feedback amplifier.***—A simple resistance-coupled amplifier employing negative feedback is

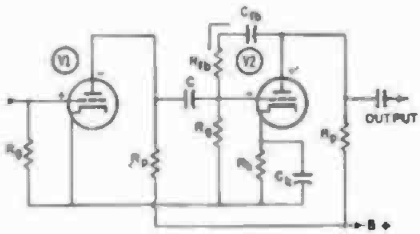


Fig. 12-149A. A two-stage RC coupled amplifier with negative feedback applied to the output stage only.

shown in Fig. 12-149A. In this circuit negative feedback is used only in the second tube; however, certain factors of the first tube enter the configuration of the equivalent circuit. Examining the equivalent circuit in Fig. 12-149B, it will be noted the grid resistor R_g is paralleled by the plate-load resistor R_p of tube V1. This, in turn, is paralleled by the plate resistance r_p of V1. The total impedance seen by the control grid of V2 is these three impedances in parallel and is designated R_{eq} (equivalent resistance).

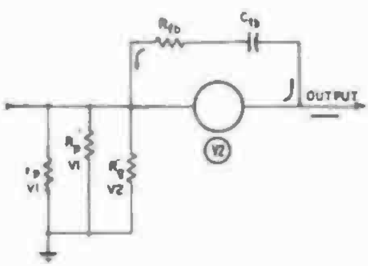


Fig. 12-149B. Equivalent circuit for the plate load of the negative-feedback amplifier shown in Fig. 12-149A.

12.150 Is it practical to include several amplifier stages in the feedback loop?—Yes, because negative feedback reduces the distortion only within the feedback loop; therefore, it is desirable to include as many stages as can be operated with good stability. If the amplifier contains several stages and has a tendency to become unstable, individual feedback loops may be used covering a single or several stages.

If the amplifier stages include coupling transformers, the proper polarities must be observed or positive feedback may be the result. If several feedback loops are used, the total feedback is the sum of the feedback in each individual loop.

12.151 How are the polarities determined for an amplifier in which the feedback loop encompasses several stages?—For an odd number of stages, the feedback loop is connected from the plate of the output tube to the control grid of the first stage. If taken from an output transformer secondary, the feedback loop is connected to the cathode of the first stage. For an even number of stages, the feedback loop is connected from the plate of the last tube to the cathode of the first stage.

12.152 Are special output transformers required for use with negative-feedback amplifiers?—No. Standard high-quality transformers may be used; however, they must have the correct impedance ratios as specified for a particular circuit operation. Output transformers should have a high degree of coupling between output winding and primary winding to reduce the effects of leakage reactance and insertion loss.

12.153 Does the method used for interstage coupling in an amplifier affect the phase-shift characteristics?—Yes, the interstage coupling has a pronounced effect on the phase-shift characteristics. This is particularly true of resistance-coupled amplifiers at very low frequencies.

12.154 What is negative-current feedback? — Negative-current feedback is obtained by connecting a resistor in series with the output load of the amplifier, as shown in Fig. 12-154. The feedback voltage is taken from across this resistor and returned to an earlier part of the circuit. Negative feedback is proportional to the output current and may be likened to a constant-current generator with infinite internal impedance, because it delivers a constant current regardless of the load impedance. If the output current increases, the feedback voltage is increased, tending to reduce the output current to its

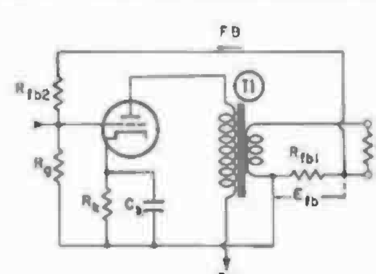


Fig. 12-154. Negative-current feedback.

former value. Negative-current feedback is not generally used in audio amplifiers except in the case where variable damping is used.

Negative-current feedback can also be obtained by omitting the cathode-bypass capacitor C_k across the cathode resistor R_k. This method decreases the gain and the distortion, but increases the internal output impedance. Consequently the output voltage rises, and if a loudspeaker is being driven, the resonant frequency of the speaker system and hangover effects are accentuated. Current feedback increases the internal output impedance, while negative-feedback voltage decreases the internal output impedance.

12.155 What is the procedure for designing an amplifier using negative feedback?—Referring to the amplifier schematic in Fig. 12-155, assume the gain of each stage is 20 or a total voltage gain of 400 (20×20). To determine the amount of negative feedback that will be required, assume the overall harmonic distortion of the two stages is 10 percent, and that it is desired to reduce this distortion to a value of 2 percent. This is a gain reduction of five, therefore:

$$(1 - A\beta) = 5$$
$$-A\beta = 4$$
$$-\beta = \frac{4.0}{A}$$

where,

A is the voltage gain of two stages before the application of negative feedback.

Substituting the voltage gain:

$$-\beta = \frac{4.0}{400}$$
$$= 0.01$$

To obtain the value of the feedback resistor R_{fb}, beta is used as a positive number.

$$\beta = \frac{R_k}{R_{fb} + R_k}$$

All values are known except R_{fb}; therefore, solving for R_{fb}:

$$R_{fb} = \frac{R_k}{\beta} - R_k$$
$$= \frac{500}{0.01} - 500$$
$$= 49{,}500 \text{ ohms.}$$

The feedback capacitor C_{fb} is selected for a value that will have a reactance equal to one-tenth of 49,500 ohms or 4950 ohms at the lowest frequency to be amplified. The capacitor may be calculated:

$$C = \frac{1}{2\pi f X_c}$$

where,

f is the lowest frequency to be amplified,

X_c is its reactance at the lowest frequency.

To express the feedback voltage in decibels use the formula:

$$dB = 20 \text{ Log}_{10} (1 - A\beta)$$
$$= 20 \text{ Log}_{10} 5$$
$$= 14 \text{ dB.}$$

The foregoing does not include the current feedback applied to the first stage because of the unbypassed cathode resistor. To determine beta for the current feedback use the formula:

$$\beta = \frac{R_k}{R_{eq}}$$

where,

R_k is the cathode resistor,

R_{eq} is the load impedance of the tube. (The load impedance is the parallel combination of R_p and the following grid resistor R_g.)

Therefore:

$$\beta = \frac{500}{100 \times 10^{-3}}$$
$$= 5 \times 10^{-3}$$
$$= 0.005$$

To express the current feedback in decibels, use the formula:

$$dB = 20 \text{ Log}_{10} (1 - A\beta)$$

where,

A is the gain of the stage to which current feedback is applied (for this example, 20).

$$dB = 20 \text{ Log}_{10} 1 - (20)\ (-0.005)$$
$$= 20 \text{ Log}_{10} 1 - (-0.1)$$
$$= 20 \text{ Log}_{10} 1.1$$
$$= 0.84 \text{ dB.}$$

In performing calculations involving beta for negative-feedback amplifiers, beta is always negative. However, when calculating the feedback resistor R_{fb}, it is used as a positive number.

12.156 What is the effect of reducing the size of the feedback-loop capacitor C_{fb}?—Less feedback is obtained at

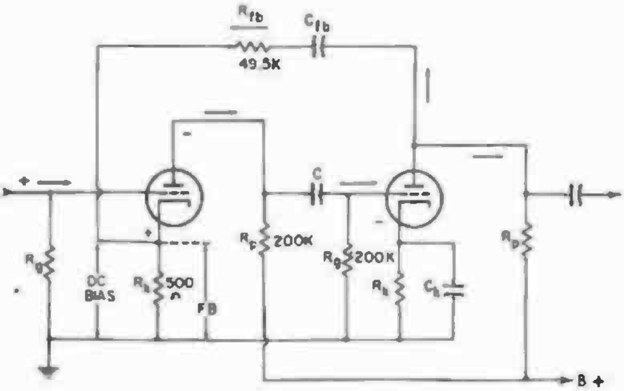


Fig. 12-155. A two-stage RC negative-feedback amplifier for design problems.

the lower frequencies, resulting in more gain at the lower frequencies. This results in a rising frequency characteristic and is sometimes used to increase the low-frequency response in preamplifiers.

12.157 Can the feedback-loop circuit be used for equalization?—Yes, if precautions are taken to prevent excessive phase shift and the amplifier is not overloaded at the maximum frequency of equalization. Fig. 12-157 shows how the equalization (equalizer components) may be connected in a feedback loop to secure a particular frequency response. Resonant circuits may be inserted in the feedback loop; however, the coils should be of fairly high Q to permit the shunting of resistors across them, and thus control the shape of the resonant curve. Resistors may be connected in series or shunt with the capacitors to obtain a particular shape of the resonant curve. For the initial adjustment, the coils are resonated to their midfrequency; then resistors are adjusted in value for desired response.

The use of tuned circuits in a feedback loop is rather risky, as oscillation may be induced because of phase shift. After final adjustment, the amplifier should be tested for oscillation by connecting an oscilloscope across the output and applying short pulses or tone bursts to the input. Also, the frequency response should be limited to about 8000 Hz. If bridged-T equalizer configurations are used in the feedback loop, it is less likely to develop oscillation. (See Question 6.96.)

12.158 Is it permissible to equalize an amplifier stage within the feedback loop?—No. Equalization is always connected in the stages outside the feedback loop because the feedback attempts to flatten out the frequency response of the stages within the feedback loop. Fig. 6-96 shows the various positions where equalization may be induced in an amplifier circuit.

12.159 What are the disadvantages of taking the feedback voltage from the primary or top of the output transformer?—Although phase shift is re-

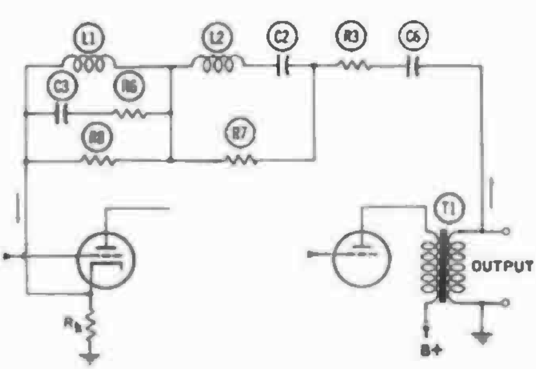


Fig. 12-157. Equalizer components in a negative-feedback loop. A parallel-resonant circuit is formed by L1 and C3. A series-resonant circuit is formed by L2 and C2.

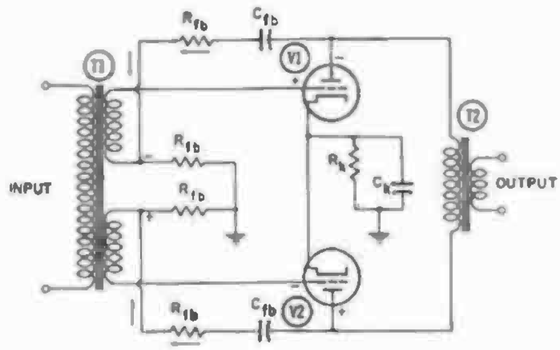


Fig. 12-161. A push-pull amplifier with negative feedback.

duced to a minimum, some loss of high frequencies may result due to the shunting effect of the feedback loop. Also, the characteristics of the output transformer are not included in the feedback loop.

12.160 What are the advantages of taking the feedback voltage from the output winding of an output transformer?—The characteristics of the output transformer are included in the feedback loop.

12.161 Is it practical to use feedback in both sides of a push-pull amplifier?—Yes; it may be done as shown in Fig. 12-161. However, the feedback loops must be carefully balanced to secure exactly the same amount of feedback in each side of the circuit. The transformer characteristics are not included in the feedback loop.

12.162 Describe a tertiary winding and its use.—It is the practice of some manufacturers to use a tertiary winding (third-winding) in the negative-feedback loop of an amplifier to eliminate the ground connection required to complete the feedback loop circuit when the characteristics of the output transformer are to be included in the feedback loop. Fig. 12-162 is a typical example of such design.

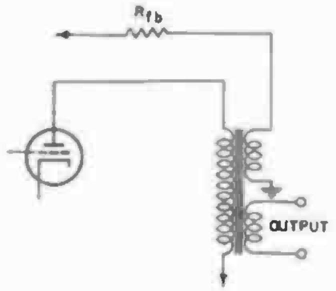


Fig. 12-162. Negative feedback using a tertiary or third coil on the output transformer.

Tertiary windings are also used on interstage transformers for driving auxiliary equipment and on output transformers for the same purposes. In designing such transformers, specified loads are given for each winding to prevent upsetting the impedance relationship of the windings relative to other portions of the circuitry.

12.163 How can a negative-feedback amplifier be used with 20 dB of feedback if it becomes unstable with 18 dB of feedback?—By introducing phase-shift networks or by staggering the time constants in the feedback loop. The time constants must also take into consideration the output transformer. Because of manufacturing variations, the output transformer is often the cause of instability and the duplication of a given circuit may be difficult.

The time constant of the output transformer will also vary with the dc balance of the tubes, core material, windings, and the excitation level of the output stage. All these factors will cause instability, particularly at the low-frequency end of the spectrum.

In a well designed negative-feedback amplifier, it should be possible to increase the feedback at least 6 dB before the amplifier becomes unstable over a frequency range one octave below the lowest frequency, and two octaves above the highest frequency.

12.164 What type amplifier is recommended for driving a cutting head, loudspeaker, light modulator, and similar devices?—One using negative feedback in the output stage, because of the damping offered by its use. (See Question 12.145.) The internal output impedance of the amplifier should be approximately one-tenth that of the load impedance. This would be a damping factor of 10:1. Damping factors below 1.0 are generally hard to obtain, because of the dc resistance of the output winding. The measurement of internal output impedance is discussed in Question 23.138.

12.165 Show several methods of connecting the negative-feedback loop

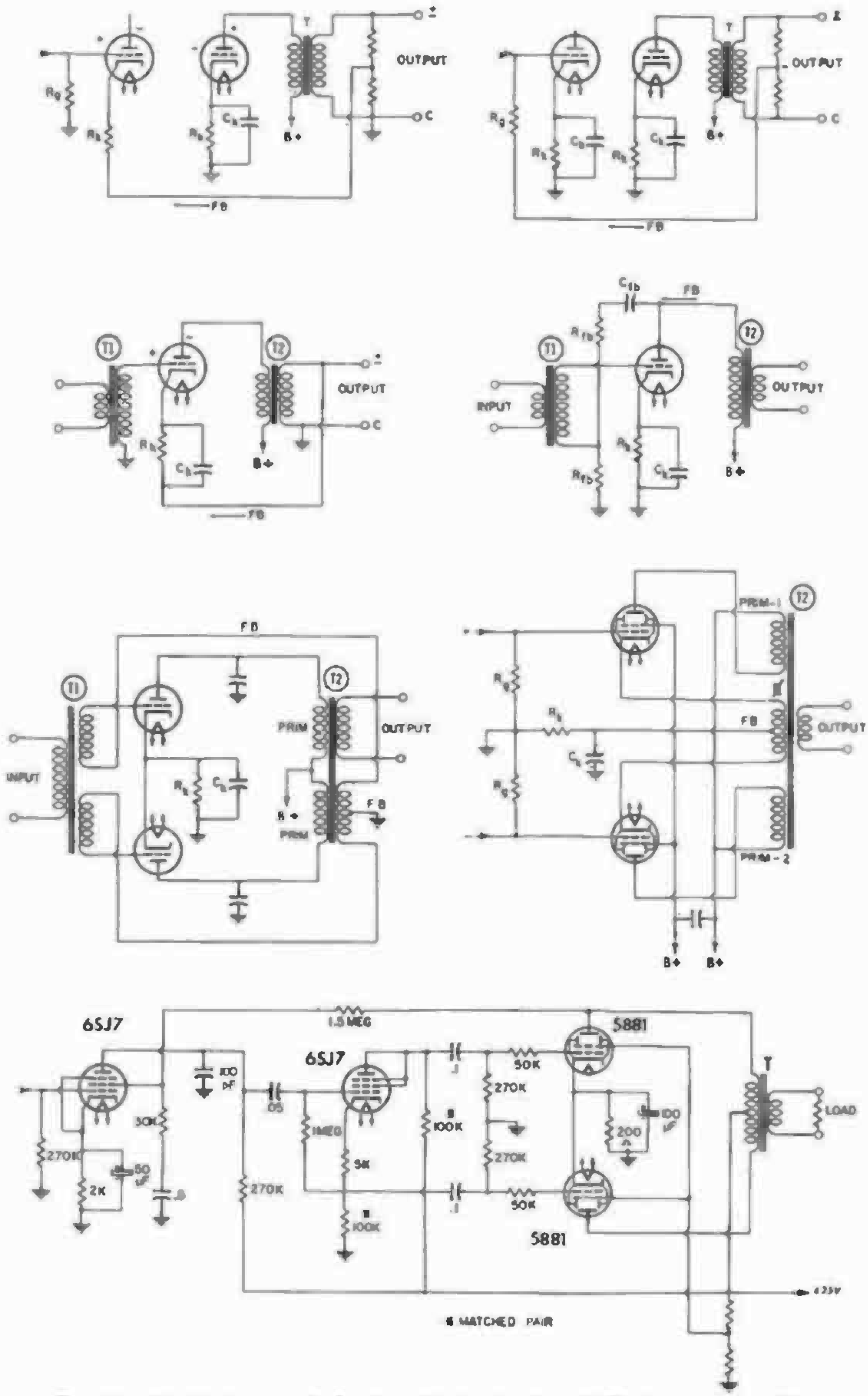


Fig. 12-165. Various methods of applying negative feedback to an amplifier.

in an amplifier.—Various methods used for applying negative feedback to an amplifier are shown in Fig. 12-165. In so doing, care must be taken that the voltage fed back from the output to the input is always 180 degrees out of phase with the voltage at the input circuit where the feedback loop is connected. The simplest way to determine the polarity of the voltage is to assume an instantaneous positive voltage at the input and then follow the phase reversals through the amplifier to the point from which the negative-feedback voltage is to be taken.

12.166 What is a balanced negative-feedback amplifier? — An amplifier employing both negative and positive feedback as shown in Fig. 12-166. The advantage of such circuitry is the wide frequency range possible without the loss of gain suffered with the conventional negative-feedback circuits. The amplifier is first designed for use with conventional negative feedback and then the positive feedback loop is added to bring the gain back to its original value before the negative-feedback loop was added. Although this type amplifier is called a balanced amplifier it is not necessary that the two feedback loops balance each other. The disadvantage of this circuit is that if the negative loop should open, the amplifier will go into violent oscillation.

12.167 How may a negative-feedback amplifier be tested for stability?— By connecting a capacitor across the output winding in parallel with a resistive load and observing with an oscilloscope the effect on a square wave applied to the input, as described in Question 23.125.

Negative-feedback amplifiers are generally thought of as being quite stable and little attention is given to the idea that they may oscillate at an inaudible frequency when a signal is applied to the input. Although the stability of an amplifier may appear to be satisfactory using a resistive termination, when a reactive load is substituted for the resistive termination the story may be quite different. This is particularly true of some types of amplifiers when a capacitive load, such as an electrostatic loudspeaker, is connected across the output. Such devices reflect a high capacitive reactance to the plates of the output tubes, equivalent to several microfarads; the exact amount will depend on the impedance ratio of the output transformer and the internal capacitance of the loudspeaker. The transformation of capacitance is discussed in Question 8.34.

As an example, assume an electrostatic loudspeaker with an internal capacitance of 0.0025 μF is connected across the output of a push-pull transformer with a ratio of 6600/16 ohms. This is an impedance ratio of 412:1. The equivalent reflected capacitance to the plates of the output tubes will be in the order of 1.03 μF. Unless the amplifier is of very stable design, it will oscillate violently. In some instances, if the capacity of the cable connecting the loudspeaker is high, it may reflect enough capacitive reactance to cause the amplifier to oscillate. When making tests for stability, capacitors ranging from 0.0025 to 1.0 μF should be connected in parallel with a resistive termination across the output and the effect observed as described in the foregoing.

Negative-feedback amplifiers in which the feedback voltage is taken from the top of the output transformer

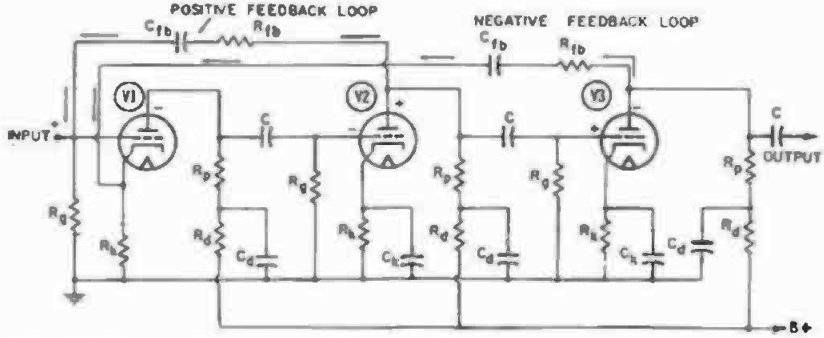


Fig. 12-166. A balanced amplifier using both negative and positive feedback.

primary are, as a rule, quite stable. Those supplying the feedback voltage from an output winding are more prone to oscillation at inaudible frequencies, because of the phase shift induced by the output windings at the high frequencies.

The stability of a negative-feedback amplifier may be increased by staggering the cutoff frequency of the various stages to vary the time constant between stages. The calculation of time constants is discussed in Question 25.106.

To assure the maximum stability in an amplifier, the design should be such that the feedback voltage can be increased at least 6 dB and preferably 10 dB over that normally used in the circuit over a frequency range of two to three octaves above and below the normal operating frequency range. This means that for an amplifier of 20 to 20,000 Hz, the frequency response must be extended from 10 to 60,000 Hz.

When an amplifier is oscillating at an inaudible frequency, the high frequencies become hard and rasping and will quickly cause listener fatigue. If the amplifier is near the point of oscillation and is shock-excited by the input signal, it will break into oscillation at subharmonic frequencies, causing the low frequencies to be distorted and have a lack of definition.

12.168 What are the phase-shift characteristics of a high-quality negative-feedback amplifier?—To obtain a flat frequency response in an amplifier from 20 to 20,000 Hz, with low distortion characteristics and phase-shift, requires that the bandwidth be 7 to 10 times the usable frequency response. Phase shift becomes apparent and measurable at values of one-seventh to one-tenth the frequency where the response falls off 2 dB, with respect to the reference frequency. Small changes in the frequency response of only 0.1 dB can cause phase shift up to 10 degrees, and a change of 3 dB can cause a shift of 45 degrees. Therefore, the negative feedback is reduced in proportion to the cosine of the phase-shift angles. This indicates the necessity for such a wide bandwidth. The phase-shift characteristics for an amplifier having a frequency response of plus or minus 1 dB, at 20 Hz to 100,000 Hz, are shown in Fig. 12-168.

12.169 What is the purpose of connecting a resistor and capacitor from plate to plate in a push-pull amplifier employing beam-power tubes in the output?—To improve the output impedance characteristics and frequency response when the tubes are terminated in a load such as a loudspeaker, cutting head, light modulator, or similar device in which the reflected impedance is not constant.

A network connected in the plate circuits of a push-pull amplifier using beam-power tubes without negative feedback is shown at part (a) in Fig. 12-169. It will be noted the network consisting of resistor R and capacitor C is connected in parallel with the reflected impedance to the tubes.

The equivalent circuit for the network shown at part (a) is given at part (b). When the reactance of the transformer primary Z_{pri}, and the resistor, R, are equal, the combined impedance seen by the plates of the tnbes will equal R. As the frequency is increased, the reactance of capacitor C decreases; thus, at the higher frequencies, only the resistor R is left in the circuit. At the lower frequencies, the reactance of C is quite high which is equivalent to removing resistor R from the circuit.

A network more suitable for push-pull amplifier operation is shown at part (c) in Fig. 12-169. Resistors R1 and R2 are about 1.3 times the recommended plate-to-plate impedance and are connected in series with capacitors C1 and C2 to ground. The capacitor is selected for a value that will result in the same voltage gain at 3000 Hz as that at 400 Hz. Generally, a capacitor of 0.03 to 0.06 μF will suffice for the average amplifier. If the components of the above networks are selected properly, a fairly uniform impedance is reflected to the plates of the output tubes resulting in

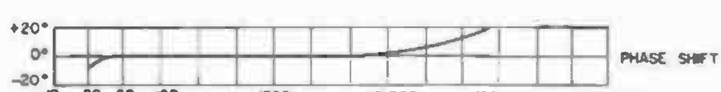
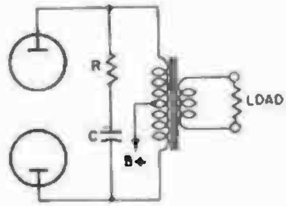
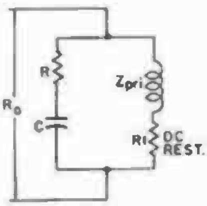


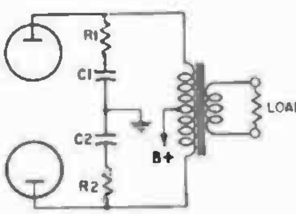
Fig. 12-168. Phase-shift characteristics for an amplifier with a frequency response of 20 to 20,000 Hz.



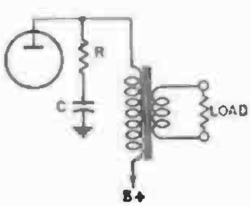
(a) *Push-pull amplifier (without negative feedback).*



(b) *Equivalent circuit for (a).*



(c) *Push-pull amplifier with network returned to ground.*



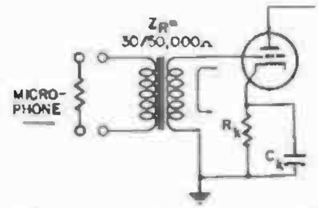
(d) *Single-ended amplifier.*

Fig. 12-169. Using a resistor and capacitor in the plate circuit to improve the output impedance characteristic and frequency response.

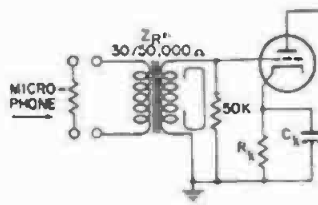
a considerable improvement in performance.

The same type of network may also be used in a single-ended amplifier as shown at part (d) in Fig. 12-169. In this instance, the resistance is also 1.3 times the recommended plate-load impedance. As a rule, the foregoing networks are not used in amplifiers employing negative feedback, although they have been used to some extent in the past.

12.170 Is it advisable to terminate the secondary of an input transformer in a microphone preamplifier?—No, because of the reduced signal voltage at the control grid. In the circuits of Fig. 12-170 are shown an unterminated input, also called an open-circuit input in part (a) and a terminated input in part (b).



(a) *Unterminated input (maximum voltage).*



(b) *Terminated input (signal voltage 50% lower).*

Fig. 12-170. Preamplifier input circuits.

To illustrate the difference between the two circuits, assume the input transformer has an impedance ratio of 30 to 50,000 ohms (1666/1). Using the unterminated circuit in part (a), the primary does not terminate the output of the microphone because the secondary winding is unloaded. In this instance, it may be said the transformer is terminated by the microphone on the primary side. Therefore, the secondary presents to the control grid of the tube a 50,000-ohm impedance and the maximum signal voltage of the secondary is applied to the control grid. (Actually, the input impedance of the tube terminates the secondary of the transformer; however, unless the input capacitance is considerable and the secondary very high, the effect of the tube capacitance may be neglected.) For the terminated secondary in part (b), only half the signal voltage is applied to the control grid. This is a 6-dB reduction of signal voltage, which also reduces the signal-to-noise ratio by 6 dB. This may be explained as follows: Connecting a 50,000-ohm resistor across the secondary of the input transformer is the same as having a 50,000-ohm generator working into a 50,000-ohm load. Therefore, the voltage at the secondary of the transformer is divided, with one-half across the internal impedance of the generator (50,000-ohm secondary) and the other half across the load (50,000-ohm resistor). Because the control grid is connected to the top of the secondary and the other end of the secondary is grounded, the control grid sees only one-half of the signal voltage—that across the 50,000-ohm resistance.

As a rule, microphone input transformers are designed to be operated open circuit (unterminated) and, because the grid circuit of a vacuum tube may be considered to be an open circuit, the secondary is said to be unterminated or working into an open circuit.

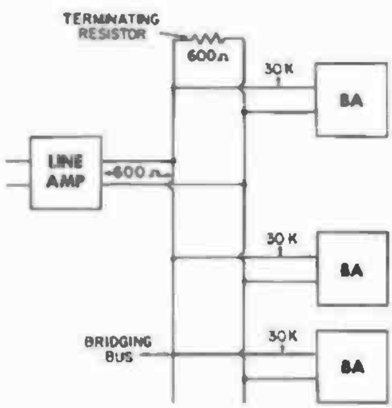


Fig. 12-171. Line amplifier with a 600-ohm output bridged by three 30,000-ohm inputs.

12.171 What is a bridging input?—A device with a high input impedance, designed for connection across an impedance of low value. Bridging inputs are used with amplifiers, volume indicator meters, and similar devices that may be required to bridge another circuit without drawing appreciable power from the circuit bridged. Standard bridging impedances used in the broadcasting and recording industries are: 7500, 10,000, 25,000, and 30,000 ohms. A 600-ohm circuit being bridged by three 30,000-ohm inputs is shown in Fig. 12-171. Bridging transformers are discussed in Question 8.62.

12.172 How can a low-impedance input transformer be used as a bridging coil?—This problem is discussed in Question 8.63.

12.173 How is the loss for a low-impedance transformer computed when it is used as a bridging transformer?

$$\text{dB loss} = 10 \operatorname{Log}_{10} \frac{B_r}{R}$$

where,

B_r is the impedance of the bridging transformer,

R is the impedance of the circuit bridged.

This equation should not be used for an impedance ratio less than 10.

12.174 What is the purpose of connecting a resistor in series with each plate of a push-pull parallel amplifier?—To balance the plate currents and to prevent parasitic oscillation. The value of these resistors is generally on the order of 47 ohms.

12.175 What is the purpose of connecting a resistor in series with each control grid of a push-pull amplifier?—The resistor serves two purposes: (1) if a sudden overload occurs, grid current will be caused to flow and the series resistor tends to reduce distortion by causing the bias on the control grid to be increased by the flow of grid current through it, and (2) the resistor also tends to prevent parasitic oscillation by acting as a damper in the grid circuit.

The use of grid-limiting resistors is not entirely confined to use in push-pull amplifiers but they are used in many vacuum-tube circuits. Generally, the value of such resistors will vary between 100 and 10,000 ohms. There are no set rules as to the exact value; however, a rule of thumb indicates that tubes with high values of G_m will have a tendency to become unstable, particularly tubes such as 6L6, 1614, 5881, 6146, 6550 and similar type tubes, unless grid-stopping resistors are employed. This instability is due to wiring and socket capacitance, causing the tubes to oscillate at very high frequencies. Stopping resistors of 500 ohms are used in screen-grid circuits, and sometimes a 47-ohm resistor is connected in the plate circuit, which helps to balance the plate currents. The resistor must be connected at the socket terminal to be effective. The use of stopping resistors is illustrated in Fig. 12-130.

12.176 What does the term plate-to-plate mean?—It is the total impedance seen by the plates of the two sides of a push-pull amplifier. (See Question 12.110.)

12.177 How is the damping factor of an amplifier calculated?—By the use of the equation given below:

$$D_f = \frac{Z_L}{Z_{out} + R_{vc}}$$

where,

Z_L is the rated load impedance,
Z_{out} the internal output impedance of the amplifier,
R_{vc} the dc resistance of the voice coil as measured by an ohmmeter.

The subject of damping is discussed in detail in Questions 12.183, 20.103, and 23.138.

12.178 What is a driver stage?—The amplifier stage preceding the output—the power-amplifier stage.

12.179 What does the term "looking into" mean?—An expression used to designate the point from which a circuit is to be considered. Example: The impedance seen when looking into the output of an amplifier.

12.180 What is source impedance?—The impedance of a device or line that is the source of a signal voltage.

12.181 What is input impedance?—The rated input impedance of a device.

12.182 What is internal input impedance?—The actual impedance seen when looking into the input terminals of a device. The actual internal impedance may be different from that specified for the source impedance. As an example, an amplifier with a bridging input impedance of 10,000 ohms may be specified to work from a 600-ohm source. Or, an amplifier designed to work from a 250-ohm source may have an actual input impedance of 1000 ohms.

12.183 What is internal output impedance?—The actual impedance seen when looking into the output terminals of a device. The internal impedance may be only a fraction of the specified load impedance. This is particularly true of negative-feedback amplifiers. A typical case is an amplifier using negative feedback which is specified to work into a 16-ohm load, although the internal outpnt impedance may only be 1.6 ohms. (See Question 12.145.)

12.184 What is load impedance?—The specified load impedance a device is designed to work into. Many times the load impedance is higher than the internal output impedance. (See Question 12.183.)

12.185 What is nominal impedance?—The impedance specified hy the manufacturer which will produce a given set of results. Amplifiers may be designed to operate from several different values of source impedance; thus, the overall gain will vary with different values of source impedance. It is customary for the manufacturer to specify the gain and other characteristics for a given source and load impedance.

12.186 What is the maximum undistorted power output of an amplifier?—The maximum power output obtainable with the lowest possible harmonic distortion.

12.187 Under what conditions is the maximum power transmitted from a generator to an external load?—When the external load is equal to the internal impedance of the generator. This is shown graphically in Fig. 12-187.

As the load impedance is increased from a low value the transmitted power is increased until a point is reached where the internal impedance of the generator and the external impedances are the same. At this point, a maximum transfer of power is effected. The voltage across the load and the generator will be equal. If the load impedance is increased to a value greater than the generator impedance, the power will drop off as shown by the right-hand portion of the curve.

Although maximum power is transmitted when the load is equal to the generator, this does not always hold true for amplifier design. It is not uncommon to find the internal output impedance (generator) of an amplifier considerably different from that of the specified load impedance.

12.188 What should the load impedance be to obtain the maximum undistorted power output from a vacuum tube?—For triodes, the plate-load impedance should be twice the plate resistance of the tube at the lowest fre-

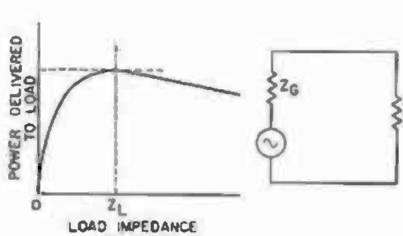


Fig. 12-187. Variation in power transfer for different values of load resistance.

quency to be amplified. For pentodes and beam-power tubes, the plate-load impedance is generally in the order of one-fifth to one-tenth the plate resistance. Under these conditions, the maximum power is obtained with the lowest harmonic distortion. In the latter instance, the manufacturer generally states the load and distortion for a given set of conditions in the tube data sheet.

12.198 What type tubes are generally selected for a voltage amplifier?—Tubes having a high transconductance with low internal noise.

12.190 What type tubes are used in power amplifiers?—Tubes which will deliver a high signal current rather than voltage. Typical power tubes are the 2A3, 6V6, 6L6, 5881, and the KT-66. All of the above, except the 2A3 are of the beam-type construction.

12.191 What is the Miller effect?—It is the effect on the high-frequency response of a vacuum tube caused by the internal capacitance existing between the various elements used in its construction.

Input capacitance exists in all vacuum tubes between the control grid and the plate, screen grid, and cathode elements. The value of this capacitance will vary with the internal construction and the amplification factor. This effect is known as the "Miller effect" and was named for its discoverer.

An interesting fact regarding the control grid-to-plate capacitance is that its apparent value is controlled by the amplification factor of the tube. When an input signal changes the voltage at the control grid, the cathode potential remains stationary and the internal capacity is charged to a given value. However, the capacitance existing between the control grid and the plate elements is not only charged by the control grid signal but also by the plate variation in voltage which is opposite in direction to the control grid change and larger by the voltage gain of the tube involved. These considerations lead to an equation for the input capacitance:

$$C_{in} = C_{cg-k} + C_{cg-p} + (A + 1)$$

where,

A is the voltage gain of the tube.
C is the capacitance between the tube elements.

The static values of interelectrode capacitance may be obtained from the tube manufacturer's data sheets. If the plate load consists entirely of resistance, the input impedance of the tube appears as a capacitance. If the plate load consists of a reactance, such as an inductance, the input of the tube appears as a resistance.

A capacitive input tends to reduce the high-frequency responses of the stage. Pentodes which have smaller input and output capacitance show considerably less Miller effect and are preferred where the frequency response must cover a wide range. (See Question 11.63.)

12.192 At what points are the signal voltages in a vacuum tube measured?—The input voltage is measured between the control grid and ground, and the output signal between the plate and ground. This statement will only hold true for conventional tube circuits, as there are instances where the signal is measured between other points and elements.

12.193 What does the term motorboating mean and what is its cause?—Motorboating is a low-frequency oscillation in an amplifier caused by common coupling between the stages of a multistage amplifier. The result is a low-frequency sound similar to the putt-putt of a motorboat; hence, its name.

Generally, this type of oscillation is caused by the lack of adequate decoupling between the amplifier stages, unwanted coupling being provided by the internal impedance of the power supply. In equipment which has been known to be functioning properly, motorboating may be caused by an open, or a changing in value of a bypass capacitor or grid resistor. Decoupling circuits are discussed in Question 12.36.

12.194 What is the purpose of supplying direct current to the heater circuit in a preamplifier?—Because a preamplifier is generally operated in a low-level circuit such as at the output of a microphone, photocell, or pickup, it is subject to hum pickup created by the ac operation of heater circuits. The use of direct current will eliminate this difficulty.

As a rule, most professional magnetic recorders use dc on the heaters of the record and playback amplifiers. The dc voltage may be obtained from small silicon or selenium rectifiers. This subject is discussed further in Question 21.64.

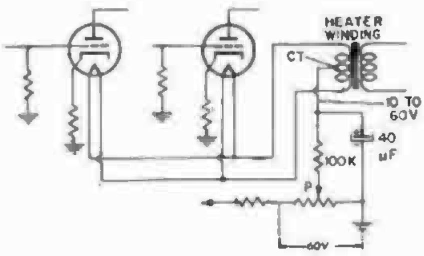


Fig. 12-195. Cathode-leakage reduction by the use of a dc bias connected to the center tap of the heater winding.

12.195 What is the purpose of connecting the center tap of an ac heater winding to a source of dc potential?—To reduce the difference of potential between the cathode and heater circuits. In this manner, leakage between these elements is eliminated resulting in a much lower noise level. A typical circuit for this use is shown in Fig. 12-195.

The dc potential is adjustable and is made positive by the same amount or a slightly higher amount than the cathode bias. Potentiometer P is adjusted while observing the noise level at the output of the amplifier, on a vacuum-tube voltmeter.

12.196 Describe an operational amplifier, its circuitry, and use.—Functionally speaking, an operational amplifier is a device which by means of negative feedback (and other elements), is capable of processing a signal voltage with a high degree of accuracy, limited only by the tolerances of the passive elements used in the amplifier and feedback networks. Electronically, it is a simple high-gain amplifier of high stability with large amounts of negative feedback connected between the input and output circuits. The external feedback elements may be resistive, reactive, nonlinear, or linear.

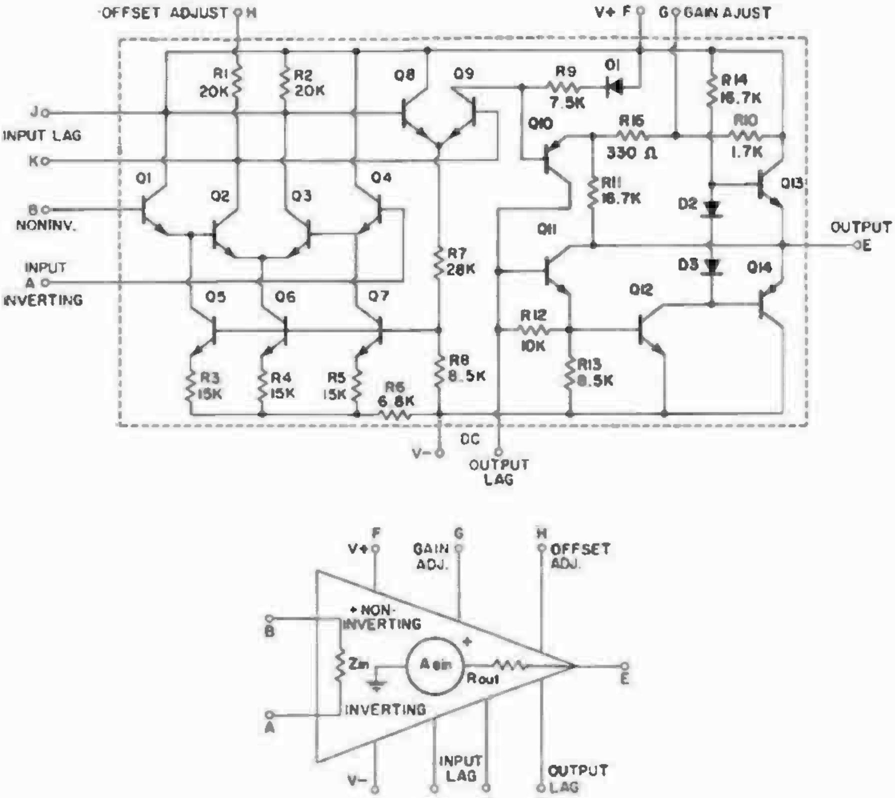


Fig. 12-196A. Circuitry for a monolithic operational amplifier, and its equivalent. (Courtesy, Motorola Semiconductor Products, Inc.)

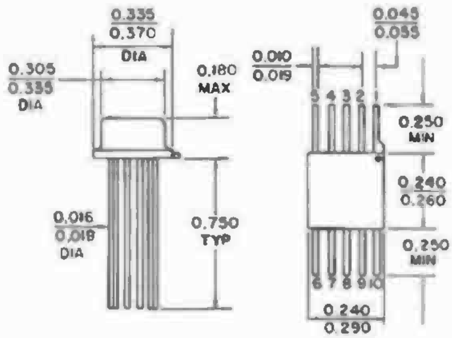


Fig. 12-196B. Physical dimensions of module. (Courtesy, Motorola Semiconductor Products, Inc.)

Operational amplifiers are of such design that an open-loop gain of several million may be reduced to one-tenth the gain by the application of negative feedback. This would be impossible using conventional amplifier circuitry. To achieve these results, the internal amplifier circuits are direct-coupled, the gain rolled-off before unity gain occurs, and before phase shift can cause oscillation. Its frequency characteristics are determined entirely by the external feedback loop and other elements. As a rule, most operational amplifiers are constructed using integrated circuit solid-state devices, although they may consist of vacuum tubes; however, tube amplifiers are not as versatile for this purpose as those employing transistors. (See Question 12.144.)

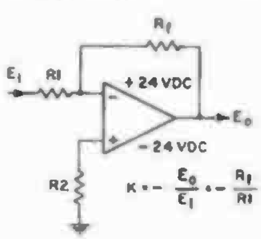
Although originally developed for use in computers, operational amplifiers find many uses in the electronic field, and are generally purchased for a particular application. They are available as booster amplifiers, phase inverters, noninverters, voltage regulators, oscillators, nonlinear amplifiers, and many other types.

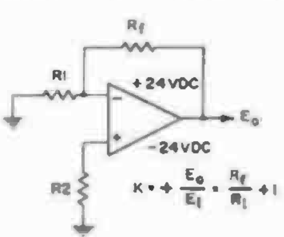
Design requirements for operational amplifiers are: high-gain, low output impedance, zero ground potential (with respect to dc), direct-coupling low dc drift, with the high-frequency response rolled-off before phase shift reaches 180 degrees.

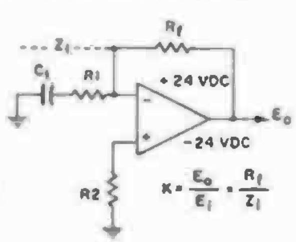
The circuitry for a monolithic-type operational amplifier is given in Fig. 12-196A, with its physical dimensions given in Fig. 12-196B. This particular circuit is designed to be used as a summing amplifier, integrator, or its operating characteristics as a function of the external feedback components. The open-loop gain is on the order of 60,000 (75.5 dB) with an output voltage of 13 volts, using a 15 Vdc power source. The internal output impedance is approximately 100 ohms.

In Figs. 12-196C to R, are given the design considerations for differential operational amplifiers, developed by B. J. Losmandy of Opamps Laboratories, and will be of considerable interest to the audio engineer, as they are similar to the design of solid-state devices used in sound-mixing consoles.

Operational amplifiers may be powered from a conventional single-voltage power supply, or from a bipolar type







—NOTE— ALL AMPLIFIER CIRCUITS ARE OPERATED FROM −24 VDC AND +24 VDC

Fig. 12-196C. Basic design circuit for inverting-gain amplifier.

Fig. 12-196D. Basic design circuit for noninverting-gain amplifier.

Fig. 12-196E. Basic design circuit for an ac amplifier.

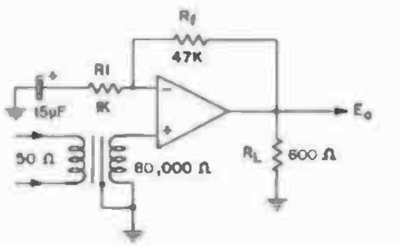


Fig. 12-196F. Microphone preamplifier.

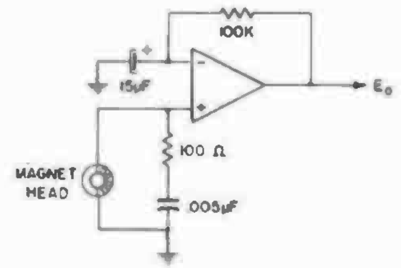


Fig. 12-196G. Magnetic-type playback amplifier.

(Fig. 12-196R). As a rule, a regulated supply is not required since the ripple voltage is a disturbance within the negative-feedback loop, and is reduced by a factor equal to the open-loop gain divided by the closed-loop gain. Thus, if the amplifier has an open-loop gain of 500, and the gain is reduced to ten, the ripple voltage is reduced by a factor of 50.

Referring to Fig. 12-196C, a differential dc gain-inverting amplifier, the closed-loop gain (K) for this circuit is equal to minus feedback resistor R_f value, divided by the value of input resistor R1. Resistor R2 is used to balance the input current to the positive input. Its value is the parallel combination of R1 and R_f. The frequency response for this circuit ranges from dc to the upper limit of the particular amplifier employed. Input voltage E_i is equal to the difference of the positive input voltage and the negative input voltage, times the open-loop voltage gain of the amplifier.

A noninverting-gain amplifier is given in Fig. 12-196D. Here the closed-loop gain (K) is equal to resistor R_f divided by R1 plus one. The additional value of one is required because of certain mathematical relations in the differential-amplifier circuitry. The dc path for the positive input is through ground; thus, bias current is supplied to the positive base of the transistor.

The amplifier of Fig. 12-196E is a typical ac amplifier. At very low frequencies, capacitor C1 approaches infinity at unity gain. Increasing the input frequency reduces the reactance of C1 until a point is reached where R1 equals $1/\pi C1$. At this point (breakaway frequency) the gain is reduced to a value of two, then increases with frequency to full gain (R_f/R1) and is then maintained over the balance of the passband. The low-frequency breakaway point $f_1 = 1/2\pi R1C1$. Capacitor C1 in the input circuit prevents the input off-set voltage from being multiplied by the gain of the amplifier (see Question 12.82) since the gain at dc is unity. This is an advantage, as each stage of a system connected in this manner will always have a minimum of dc off-set voltage in the output. Furthermore, this configuration has no phase reversal of the input signal. Therefore, when used in sound mixing circuits, the signals are not out-of-phase with respect to each other. (See Question 12.82.)

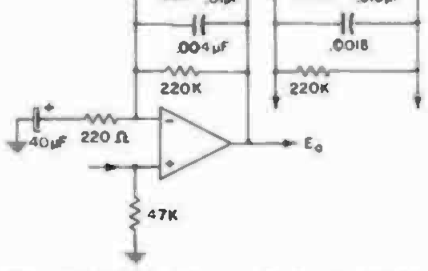


Fig. 12-196H. Phono playback, RIAA equalization.

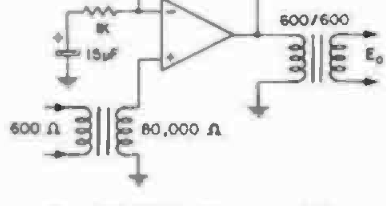


Fig. 12-196I. Line amplifier.

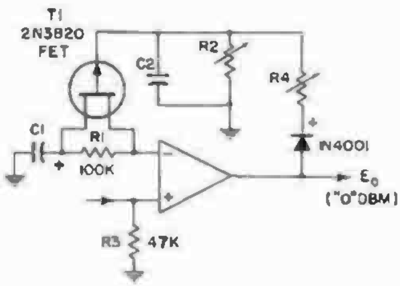


Fig. 12-196J. Compressor amplifier.

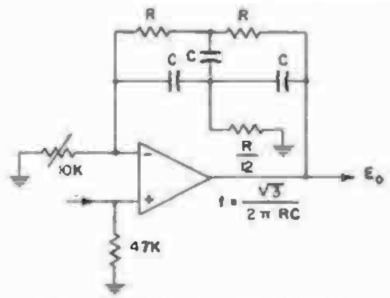


Fig. 12-196K. Bandpass amplifier.

Illustrated in Fig. 12-196F, is a microphone preamplifier with approximately 60-dB (1000) gain, capable of developing plus 22 dBm into 600 ohms. A transformer is used in the input between the output of the microphone and the input of the positive amplifier to step-up the voltage and match impedances. The overall noise is reduced since the total noise is equal to the internal noise of the transformer, plus the noise of the amplifier divided by the step-up ratio of the transformer. With the closed loop adjusted for a gain of 50, the overall gain is 60 dB. For this particular configuration, this is accomplished by adjusting feedback resistor R_f to 47,000 ohms, and the series input resistor R1 to 1000 ohms. The low-frequency minus 3-dB point is set by the combination of the 15-μF capacitor and the 1000-ohm resistor in the input, resulting in a low-frequency cutoff point of approximately 10 Hz. This circuit may also be employed as a playback amplifier for a variable reluctance pickup, by connecting the pickup between the positive input and ground.

For use as a magnetic playback amplifier, the configuration of Fig. 12-196G is employed. Here, the magnetic head is connected between the positive input and ground, with an RC network consisting of a 100-ohm resistor and a 0.005-μF capacitor in parallel to prevent self-oscillation. The network will not affect the frequency response of the head. It will be observed, the input series resistor R1 has been omitted; thus, a full open-loop gain of 54 dB (500) is realized.

An amplifier equalized for phonograph reproduction (RIAA) is given in Fig. 12-196H. The low-frequency breakaway point is set by the 40-μF capacitor and 220-ohm resistor connected in series with the negative input. For frequencies below 20 Hz, the slope is 6 dB per octave downward. The feedback resistor network has a breakaway point of 50 Hz, dropping 6 dB per octave up to 2000 Hz, then continuing out to about 30,000 Hz, then again dropping off at the rate of 6 dB per octave. The gain at 30 Hz is 60 dB, and 22 dB at 10,000 Hz. (See Question 13.95.) This same configuration also may be used for NAB tape equalization (playback) by changing

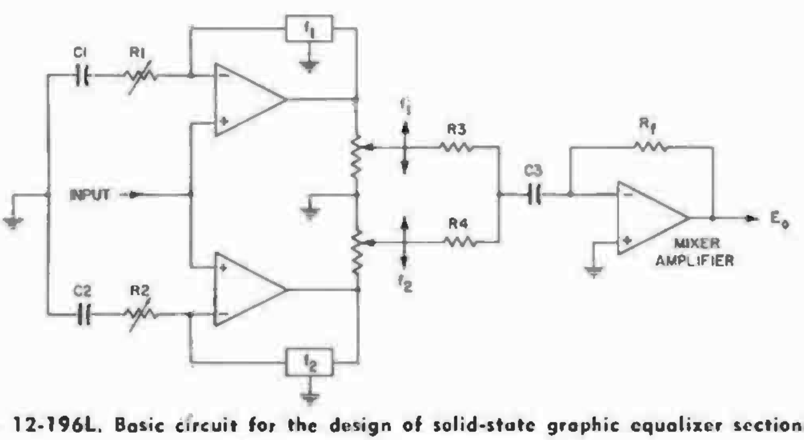


Fig. 12-196L. Basic circuit for the design of solid-state graphic equalizer sections. Only two inputs are shown here.

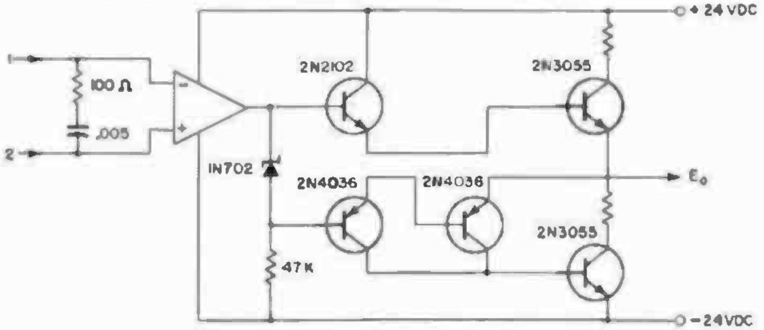


Fig. 12-196M. A 50-watt operational power amplifier.

the component values in the feedback loop given in the same diagram. For this configuration, the low-frequency breakaway point is 20 Hz, and the first high frequency is 50 Hz, then dropping 6 dB per octave to 500 Hz. From this point on, the response remains flat out to 2000 Hz, then again decreasing at a rate of 6 dB per octave. The gain at 30 Hz is 60 dB. (See Question 17.173.)

A line amplifier (Fig. 12-196I) differs from the foregoing circuits, as it makes use of both an input and output transformer. The configuration shown has a gain of 40 dB (100), with a maximum output level of plus 22 dBm. Decreasing the primary impedance of the output transformer from 600 ohms to 150 ohms will increase the output level to plus 30 dBm. The overall frequency response is limited by the frequency response of the transformers.

Compressor amplifiers are an absolute necessity in sound recording. (See Questions 18.84 to 18.101.) A simple configuration is given in Fig. 12-196J. This amplifier has adjustable attack and release-time controls with a maximum compression of 20 dB. The circuit makes use of a p-channel junction field-effect transistor (FET) connected across resistor R1. With the output level set to minus 20 dBm, the FET in the on condition, it presents an equivalent source-to-drain resistance of about 1000 ohms. Therefore, the overall gain is high. Increasing the output level and rectifying the signal through diode IN4001, it appears as a positive voltage at the gate of the FET (see Question 11.145). When the output level reaches about 0.8 V, the FET is then fully off, and the gain drops to a value of two. At low levels, the gain is about 10, resulting in a dynamic range of 10:1 (20 dB). Resistor R4 controls the charging rate of capacitor C2, thus the *attack time*. Variable resistor R2 determines the discharge time of C2, thus the *release time*. It will be noted the FET is connected inside the negative-feedback loop and automatically adjusts the gain of the amplifier as a function of output level.

Quite often a bandpass amplifier is required. Referring to Fig. 12-196K, this may be accomplished by the use of a notch filter connected in the feedback loop. The gain is controlled by the 10,000-ohm variable resistor in the input also serving as a "Q" control for the circuit, permitting the bandwidth of the filter circuit to be varied over a limited range (see Question 7.12).

The basic configuration for developing graphic equalizers is given in Fig. 12-196L. Here, each input (up to 20,

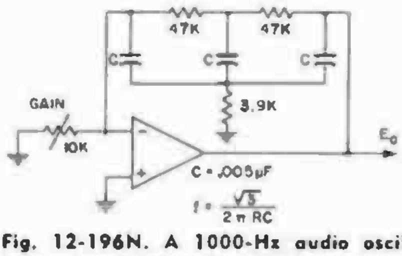


Fig. 12-196N. A 1000-Hz audio oscillator. Amplitude of oscillation is controlled by the variable resistor in the input circuit.

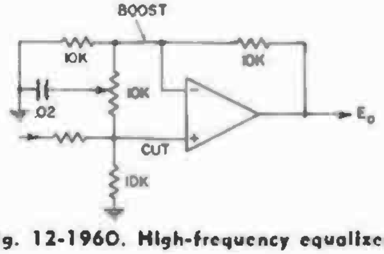


Fig. 12-196O. High-frequency equalizer.

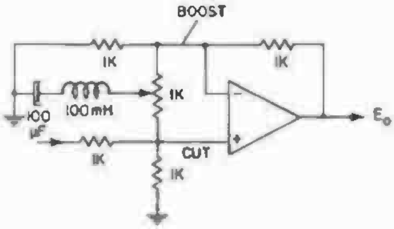


Fig. 12-196P. Low-frequency equalizer.

although only two are shown) has for its feedback elements reactive-tank circuits f_1 and f_2, or a notch filter. Each input is resonated at a selected frequency. Potentiometers R3 and R4 are connected in the output circuits of each channel and set to their midpoints. Controls R1 and R2 are adjusted to result in a broad, flat frequency response after being *summed* into the mixer-amplifier stage. To accentuate or attenuate either one of the selected frequencies (f_1 or f_2) potentiometers R3 and R4 are adjusted for the desired response in a particular stage. With high-Q LC resonant circuits, about 20-dB boost and 20-dB attenuation may be expected. The operation of graphic equalizers is discussed in Questions 6.108 to 6.128.

Configuratiou in Fig. 12-196M is an operational 50-watt amplifier, consisting of two silicon 2N3055 transistors. The upper output transistor is driven from a 2N2102 transistor connected as an emitter follower, resulting in a Darlington circuit. The lower transistor is a combination Darlington quasi-complementary configuration. The four base-to-emitter junctions of the drivers and power stages are forward-biased for class-AB operation by zener diode 1N702. Negative pull-down current is supplied by the 47,000-ohm resistor. The power stages are fed from an operational amplifier, given in Fig. 12-196E. Stability is assured by rolling-off the positive and negative input response at 320 kHz, using a 100-ohm resistor and a 0.005-μF capacitor across the two inputs. (See Questions 12.251 to 12.253.)

Power amplifiers as shown are generally operated with a closed-loop gain ranging between 2 and 20. If necessary, the power output may be increased by paralleling additional 2N3055 transistors across the output stages. In this latter instance, 0.5-ohm resistors are connected in each emitter element of the 2N3055 transistors.

Fixed frequency oscillators may be designed using the configuration of Fig. 12-196N, consisting of a 1000-Hz notch filter (or any other frequency) conuected in the feedback loop. The amplitude of oscillation is controlled by means of a 10,000-ohm variable resistor in the input circuit, and operates similar to the bandpass filter of Fig. 12-196K. Configurations for high- and low-frequency equalization are given in Figs. 12-196O and 12-196P. The high-frequency configuration employs a capacitor connected to a potentiometer across the negative and positive input circuits. The gain of the positive amplifier is only two (feedback loop adjusted for a gain of one). The positive input employs a voltage divider to reduce the gain to unity. A potentiometer is connected across the positive and negative inputs, with a capacitor between the swinger and ground. Moving the swinger from one side to the other provides an adjustable bypass from either one of the input circuits. The range of attenuation or equalization is 20 dB each, at 10,000 Hz.

Setting the potentiometer to the negative input, the upper 10,000-ohm series resistor to ground is completely bypassed and the gain is maximum. When the swinger is placed at the positive in-

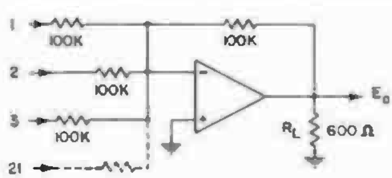


Fig. 12-196Q. Mixer-amplifier (active combining network).

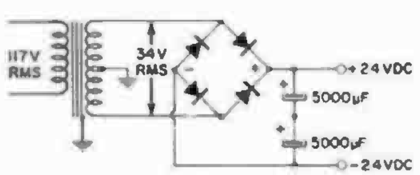


Fig. 12-196R. Bipolar power supply for operational amplifiers.

put, the input signal is also bypassed to ground. Increasing the frequency passes less signal to the amplifier input, thus the signal is attenuated. The parallel combination of the 0.02-μF capacitor and the values of resistance employed determines the frequency of turn-over. For the values given, the turn-over freqnency is 800 Hz.

For control of low frequencies, the configuration in Fig. 12-169P may be used. Here the circuit is much the same, except an LC rather than an RC circuit is used. With the potentiometer at the negative input, the inductance would short-circuit this point through the dc resistance to ground, thus driving the amplifier into saturation. This is avoided by the connection of a 100-μF capacitor in series with the inductor. The LC circuit should be resonated at the low-frequency end of the system in which it is employed. Fnr the values given, the resonant frequency is 50 Hz. The operation of this circuit is similar to the high-frequency configuration. The turnover frequency is 1600 Hz.

A mixing circuit, capable of handling up to 20 inputs, presenting no insertion loss, is given in Fig. 12-196Q. Since this circuit has uniform frequency characteristics, it may be operated as an amplifier with gain, depending on the ratio of R_f to series resistor R1. For minimum cross-talk conditions, the lowest source impedance, highest input series resistor, and highest gain amplifier shonld be used. Cross talk for the values given is on the order of 80 dB, when operating from a source and load impedance of 600 ohms. The gain equals one, with an output level of plus 22 dBm. Such circuits are often referred to as active combining networks. (See Question 5.99.)

Shown in Fig. 12-196S is the internal circuitry for a Motorola Model MC-1533 integrated circuit operational amplifier. The input stage consists of Darlington differential amplifiers Q1, Q2, Q3 and Q4 which provide a high input resistance of 1 megohm. Separate sources of current are used for Q1 and Q2 rather than a single source for the whole stage. Thus, a constant collector current is provided, making the Darlington amplifier characteristics independent of beta. Temperature compensation is provided by the feedback circuit from the second stage Q8 and Q9, through a 28,000-ohm resistor.

Q10 is a pnp low-gain transistor, driving emitter follower Q11. Transistors Q10 and Q11 shift the quiescent dc level applied to driver Q12 to almost zero. Negative feedback taken from the

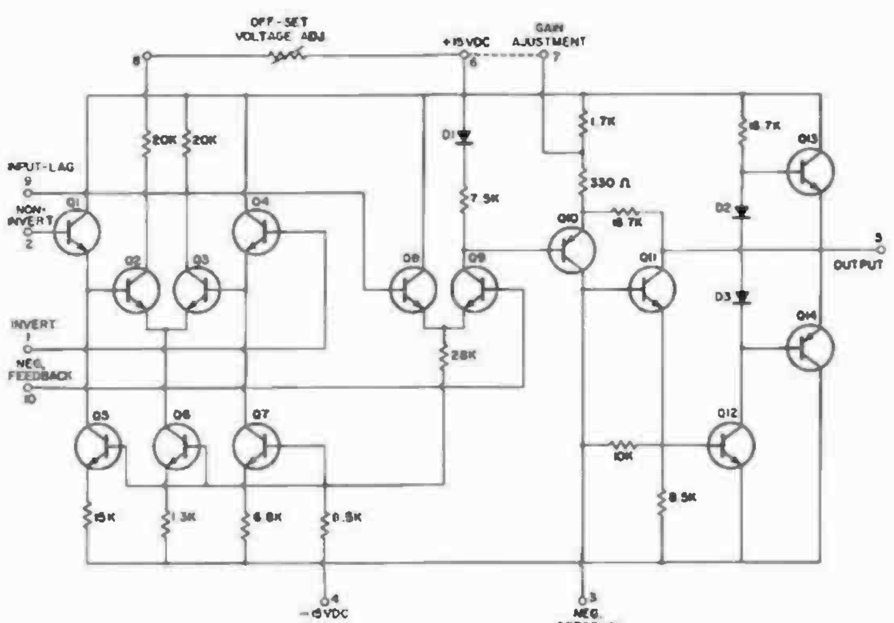


Fig. 12-196S. Internal circuitry of Motorola Model MC1533 IC operational amplifier. Negative feedback is applied between terminals 3 and 9.

output is applied to these stages to reduce the effects of crossover distortion. Output transistors Q13 and Q14 are connected complementary-symmetry biased for class-AB operation, and driven by transistor Q12. Bias voltage is supplied by the voltage drop across diodes D2 and D3, providing a small quiescent forward current to Q13 and Q14. Temperature compensation is provided by diode D1.

This circuit will provide an open-loop gain of 60,000 (75.5 dB). Maximum gain is attained when pin 7 is tied to pin 6. With pin 7 open, the gain is reduced approximately 6 dB because of degeneration in the emitter circuit of Q10. The input offset current is adjusted to zero by adjusting the voltage between pin 8 and 6. With a 15-volt supply, the output signal is 13 volts. Output impedance is 100 ohms. Common-mode rejection is 100 dB.

12.197 How may radio-signal pickup be avoided in a preamplifier stage?—By connecting a 5-millihenry radio-frequency choke coil in the upper end of the cathode in the first stage as shown in Fig. 12-197. The choke coil must *not* be bypassed. As the dc resistance of such choke coils is generally quite low, the bias voltage is not greatly affected; however, if the dc resistance does affect the bias voltage the value of the bias resistor may be decreased to compensate for the dc resistance of the choke. (See Question 24.71.)

12.198 What is a cross-coupled amplifier?—A two-stage, push-pull amplifier as shown in Fig. 12-198 used for driving a push-pull output stage. A balanced signal is applied to the control grids of tubes V1 and V2. The output signal is taken from the plates of V3 and V4. The advantages of the above circuit are low distortion and balanced output voltages.

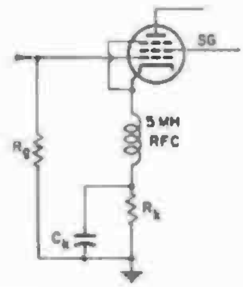


Fig. 12-197. An rf choke connected in the cathode circuit of a preamplifier stage to prevent radio interference.

12.199 If two amplifiers are connected in tandem, what is the overall frequency response?—It is the algebraic sum of the individual frequency-response characteristics of the two amplifiers. This is true regardless of the type of coupling or whether the amplifiers are individual units, as might be found in a recording system.

12.200 If two amplifiers of 20-dB gain each are connected in tandem, what is the overall gain?—Assuming all impedance matches are satisfied, the overall gain is the sum of the individual amplifiers, or 40 dB.

12.201 If two amplifiers of the same power output are connected in parallel, what is the total power output?—Assuming the amplifiers are similar in design and the output impedances are the same, the power output will be double that of a single amplifier, pro-

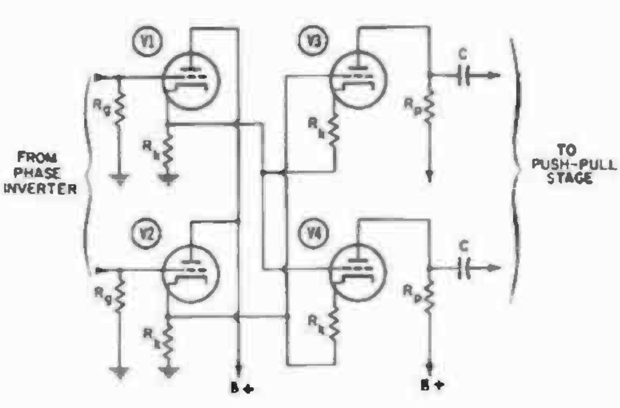


Fig. 12-198. A cross-coupled amplifier designed to take a balanced signal and supply a balanced output for driving a push-pull power amplifier.

vided the outputs are connected in phase. The output impedance will be halved.

12.202 How are the output circuits of parallel-connected amplifiers phased? —Each amplifier is adjusted for the same gain by sending a constant-level signal through both amplifier inputs simultaneously. The output circuits are then connected in parallel. If the output circuits are in phase, the output voltage will be maximum; if they are out of phase, the voltage will drop to a low value. These tests should be conducted at a low power output in an unterminated circuit. When the two output circuits are in phase, the output impedance is one-half of the normal output impedance of one amplifier.

12.203 If two amplifiers of the same power output are connected in parallel, how much is the output level increased?—Three dB and may be computed:

$$10 \operatorname{Log}_{10} 2 = 10 \times 0.301 = 3.01 \text{ dB}.$$

where,

the output impedances are the same, the outputs are in phase.

12.204 What does the term optimum plate load mean?—The ideal plate-load impedance for a given tube and set of operating conditions.

12.205 What type resistors are recommended for high-gain, low-level amplifier construction? — Noninductive wirewound types, as they have extremely low internal noise levels. Wirewound resistors are recommended for grid, plate, screen, and cathode use. Amplifiers using such resistors generally have a much lower internal noise level. The resistors must be noninductive if the amplifier is to cover a very wide band of frequencies. (See Question 5.16.)

12.206 How should the power-output rating of an amplifier be stated?— The power output should be stated for a power at which the harmonic or intermodulation distortion is not objectionable. A typical example would be: Maximum continuous power output 30 watts; total rms harmonic distortion 0.30 percent at 400 Hz, intermodulation distortion 0.5 percent for peaks less than 60 watts, using 40 and 7000 Hz mixed four to one; high frequency 12 dB lower in amplitude than low frequency.

When power amplifiers are rated in *music power output*, they are tested and rated in accordance with the Institute of High Fidelity (IHF) Standard A-201-1966, as discussed in Question 23.208.

12.207 How should the internal noise and hum of an amplifier be stated? —It should be stated in two ways: (1) with reference to the maximum power output, and (2) with reference to a 1-milliwatt reference level.

The mere statement that the noise level is so many dB below the maximum power output is often misleading unless correctly interpreted. If the noise level is stated to be 90 dB below 20 watts output, the actual noise level is, with reference to 1 milliwatt, a minus 47 dBm. This figure is obtained by subtracting the power output level, 20 watts (plus 43 dBm), from the stated noise level of 90 dB.

This interpretation is important because the amplifier is not always operated at its maximum power output. If the internal noise and hum are high, in a small room they can become quite objectionable. For high-quality reproduction, the internal noise of an amplifier should be at least 40 dB below a reference level of 1 milliwatt.

12.208 What causes the internal noise in an amplifier?—The components which make up the amplifier such as the resistors, tubes, capacitors, transformers, gain controls, and ripple voltage from the power supply. Magnetic coupling through the metal chassis also can be a cause of hum. (See Question 23.59.)

12.209 What is a class-D amplifier? —The term class-D applied to an amplifier was proposed by Crowhurst to classify a pulsed-type transistor audio amplifier. The theory behind its design being that the transistors can be operated close to 100-percent efficiency by switching at an ultrasonic frequency. The effects of switching are filtered out of the program signal.

12.210 What is magnetic coupling in an amplifier?—It is hum and noise induced in the low-level stages caused by the close proximity of the power transformer or filter chokes mounted on the same chassis. Hum and noise may also be induced by the wiring and ground loops. The subject of magnetic coupling is discussed further in Question 23.59.

12.211 What precautions should be taken when connecting equalization networks in an amplifier stage?—Measurements should be made to determine that the stage in which the equalization has been connected is not being overloaded when the maximum equalization is used. It should be remembered that an equalizer induces a loss. The amplifier must be capable of developing sufficient output to overcome the loss of the equalizer. If this precaution is not observed, serious distortion will be the result. (See Question 23.13.)

12.212 What is the purpose of extending the frequency range of an audio amplifier into the inaudible frequency range?—To obtain a frequency response covering a range of 20 to 20,000 Hz, with low distortion, flat frequency response, and low phase shift, requires a very wide frequency band for the following reasons: Phase shift becomes apparent and measurable at values of one-seventh to one-tenth the frequency where the amplifier response is down 2 dB with respect to the reference frequency.

To obtain the foregoing frequency response with less than 1-percent total harmonic distortion, the bandwidth must be 7 to 10 times the usable frequency range. This means that for an amplifier to reproduce 20 to 20,000 Hz, the frequency range of the amplifier must extend from 2 to 200,000 Hz. This frequency range is not unusual for modern high-quality amplifiers. (See Question 12.231.) This wide frequency range also permits high frequency sum and difference frequencies generated by the instruments in the original program material to be reproduced in their normal frequency bands.

A wide frequency range must exist without any trace of oscillation either in a static or dynamic condition, as oscillation at the subaudible and inaudible frequencies adds distortion to the reproduction. Oscillation at the extreme ends of the frequency response may be checked by means of a square wave generator and an oscilloscope as described in Question 23.49.

12.213 If an amplifier is oscillating at an inaudible frequency, what is the effect on the reproduction?—Harshness of the high frequencies and a rasping type distortion which induces listening fatigue.

12.214 Is the efficiency of an amplifier output stage important?—No, the efficiency of an output stage is of little importance compared to the overall current consumed by the amplifier. The greatest portion of the power consumed by an amplifier is used to heat the filaments or heaters of the tubes.

The efficiency of an amplifier stage may be calculated:

$$E_{ff} = \frac{P}{E_{av}\, I_{av}}$$

where,

P is the developed power,
E_{av} is the average voltage at the plate,
I_{av} is the average plate current.

However, this is not true of transistor amplifier output stages, as efficiency is important particularly if the amplifier is battery operated.

12.215 What is a starvation amplifier?—An amplifier employing pentode tubes with the screen voltage set 10 percent below the plate voltage and the plate-load resistance increased 10 times that normally used for a particular tube. In this manner the amplification factor is greatly increased, although the transconductance is decreased. A stage gain of 2000 or more is possible with a circuit of this kind.

12.216 How should the gain of a bridging amplifier be stated?—The actual gain when the amplifier is bridging a given impedance. As an example; assume a bridging amplifier is being fed from a 600-ohm bridging bus. The input impedance of the amplifier is 30,000 ohms. The bus level is a plus 4 dBm and the output level of the amplifier is a plus 22 dBm, what is the gain of the amplifier?

The gain is obtained by subtracting the bus level (plus 4 dBm) from the output level (plus 22 dBm) or 18 dB. There is a bridging loss of 17 dB between the input of the amplifier and the bridging bus which is the ratio of 600/30,000 and may be calculated:

$$\frac{30{,}000}{600} = 50$$

$$10 \operatorname{Log}_{10} 50 = 10 \times 1.699$$
$$= 17 \text{ dB.}$$

The actual gain of the amplifier when fed from a 30,000-ohm source is 35 dB.

The gain is stated as being 18 dB when operating from a 600-ohm source impedance. (See Question 8.62.)

12.217 Describe the procedure for designing a small single-ended amplifier with negative feedback.—Assume it is desired to design a small single-ended power amplifier which is capable of producing 6.5 watts with a maximum of 3 percent total rms harmonic distortion. The input and output impedances are to be 600 ohms. The maximum signal-input level is plus 4 dBm.

The first step is to block diagram the amplifier as shown in Fig. 12-217A. The various factors of design will be entered in these blocks as the design progresses.

For an amplifier of the foregoing proportions, a single beam-power tube such as a 6L6 will be satisfactory. Referring to the operational data for this tube, operating class-A, 6.5 watts may be obtained under the following operating conditions:

Plate voltage	250 volts
Screen grid	250 volts
Grid bias	14.5 volts
Cathode resistance	170 ohms
Peak audio-frequency voltage	14.0 volts
Maximum plate-signal current	78 mils
Maximum screen-signal current	7.5 mils
Load resistance	2500 ohms
Total rms harmonic distortion at 6.5 watts output	10 percent
Maximum power output	6.5 watts

It will be noted the distortion is stated to be 10 percent; this is without negative feedback. By the use of feedback, the distortion is to be reduced to a maximum of 3 percent.

The amplifier is first designed without feedback, and then feedback is added as explained later in the procedure. If the distortion is 10 percent without feedback and the final distortion is to be 3 percent, the distortion reduction factor is 3.33. Allowing a margin of safety, a factor of 4 will be used; thus, the gain factor is also 4. With a signal of 14 volts without feedback, 56 volts will be necessary with feedback since to maintain the same signal-to-power output, the signal at the control grid of the 6L6 must be increased by a factor of 4. This value is entered at the input of the output stage on the block diagram. The next step is to determine the gain required to drive the output stage from an input signal level of plus 4 dBm. The power at the input in watts is:

$$P = \text{Antilog}\,\frac{\text{dB}}{10} \times \text{ref. level}$$

where,

dB is the input level of plus 4 dBm,
the reference level is 0.001 watt (1 milliwatt).

The input power may now be calculated:

$$\text{Antilog } 0.4 \times 0.001 = 2.51 \times 0.001$$
$$= 0.00251 \text{ watt}$$

To find the voltage required at the input:

$$E = \sqrt{WZ}$$
$$= \sqrt{0.00251 \times 600}$$
$$= \sqrt{1.51}$$
$$= 1.23 \text{ volts rms}$$

where,

W is the watts as determined from the foregoing,
Z is the input impedance of the input transformer (600 ohms).

It is desirable at this time to convert the above voltage to peak volts, since the signal voltage at the input of the power stage is in peak volts.

$$\text{Peak volts} = 1.23 \times 1.414$$
$$= 1.74 \text{ volts}$$

The required voltage gain between the input of the voltage amplifier and the control grid of the power stage is:

$$VG = \frac{56}{1.74}$$
$$= 32.2$$

The next step is to select an impedance ratio for the input transformer.

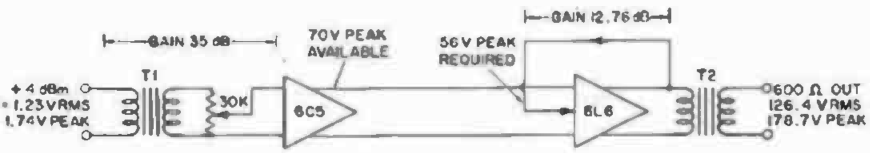


Fig. 12-217A. Block diagram for amplifier design problem.

For this problem, an impedance ratio of 600/30,000 ohms will be used, although a ratio of 50,000 to 100,000 ohms is common. The turns ratio for an impedance ratio of 600/30,000 ohms is:

$$\text{Turns ratio} = \sqrt{Z_r} = \sqrt{50} = 7.07$$

where,
Z_r is the impedance ratio.

This means that the signal voltage at the primary of the input transformer will be increased by a factor of 7.07.

As this amplifier is to present a 600-ohm impedance to the source, the secondary is terminated in a 30,000-ohm gain control. This reduces the signal voltage at the secondary to one-half the voltage obtained without the termination. Thus, the increase in signal voltage is one-half of 7.07 or 3.5. (See Question 12.170.) From the previous calculation, it was determined that 32 dB of gain was required in the first stage. To permit a small margin of safety, the gain will be taken as 35 dB.

Gain is achieved by the amplification of the first tube and the step-up ratio of the input transformer. The tube for the first stage is now selected for a voltage gain of ten or more, that will produce a peak signal voltage of 56 or greater at its plate. A medium-mu triode, such as the 6C5, resistance-coupled will produce about 70 volts rms at its plate. Therefore this tube will be used for the first stage. Referring to a resistance-coupled data chart for the 6C5, 70 volts rms signal voltage may be obtained under the following conditions:

Plate supply voltage	300 volts
Plate load resistance	50,000 ohms
Following grid resistance	100,000 ohms
Cathode resistor	2600 ohms
Coupling capacitor	0.04 μF
Voltage gain	11
Peak output volts	70 volts
Cathode bypass capacitor	2.3 μF

Both the cathode and coupling capacitors will be multiplied by a factor of 4 to extend the low-frequency response as explained in Questions 12.19 to 12.21. The next portion of the problem will be devoted to the design of the negative-feedback loop. Negative feedback will be taken from the plate of the 6L6 to its control grid through resistor R_{fb} and capacitor C_{fb} (Fig. 12-217B).

To calculate the feedback circuit, the gain of the power stage must be determined. As the stage develops 6.5 watts into a load impedance of 2500 ohms, the signal voltage at the plate is:

$$E = \sqrt{WZ} = \sqrt{6.5 \times 2500} = \sqrt{16{,}250} = 126.4 \text{ rms}$$

where,
E is the signal voltage developed at the plate,
W is the power developed by the tube,
Z is the load impedance.

To convert the rms plate voltage to peak voltage it is multiplied by 1.414, resulting in a peak voltage of 178.7 volts. The voltage gain is the peak output voltage divided by the peak input voltage (14) which is 12.76. Beta may now by calculated:

$$(1 - A\beta) = 4.00$$
$$-A\beta = 3.00$$
$$-\beta = \frac{3.00}{12.76} = 0.235$$

The equivalent resistance from the control grid of the 6L6 to ground is the resistance across which the feedback voltage is delivered and consists of the grid resistor R_g, the plate-load resistor

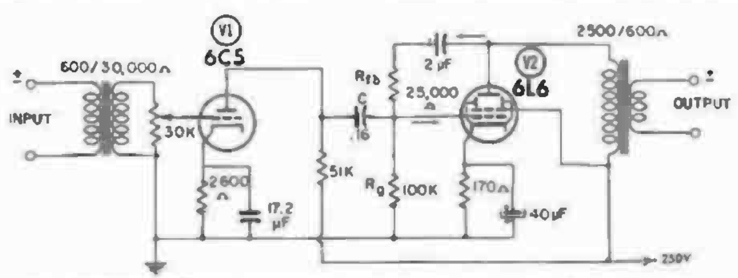


Fig. 12-217B. Power amplifier and driver stage for design problem.

R_g, and the plate resistance r_p of the preceding tube, all in parallel. The values of these components are:

R_g 100,000 ohms
R_g 50,000 ohms
r_p 10,000 ohms

The equivalent resistance R_{eq} of the above is 7700 ohms.

The feedback resistor R_{fb} may be calculated:

$$\beta = \frac{R_{eq}}{R_{fb} + R_{eq}}$$

$$R_{fb} = \frac{R_{eq}}{\beta} - R_{eq}$$

$$= \frac{7700}{0.235} - 7700$$

$$= 32{,}750 - 7700$$

$$= 25{,}050$$

where,
R_{eq} is the equivalent resistance,
R_{fb} is the feedback resistance.

The feedback loop extends from the plate to the control grid of the output tube. The resistor R_{fb} establishes the correct feedback voltage at the control grid. The capacitor C_{fb} is included to isolate the plate voltage from the grid circuit. This capacitor must have a reactance at least one-tenth the value of the total resistance in the feedback loop at the lowest frequency to be amplified.

The total resistance R_{eq} is 7700 ohms and R_{fb} 25,050 ohms making a total of 32,750 ohms. Therefore, the reactance of C_{fb} should be approximately 3300 ohms at the lowest operating frequency which, for this problem, is assumed to be 40 Hz. This will result in a capacitance of 2.0 μF.

The cathode resistor is 170 ohms and bypassed with a 40- to 50-μF electrolytic capacitor. The total plate current drawn by the amplifier will be the maximum plate and screen currents of the output tube, plus the current drawn by the first stage.

12.218 If the size of the capacitor in the feedback loop in the previous question is altered, what effect will it have on the frequency response?—Normally, the reactive element in a feedback loop is of such value it has little or no effect on the frequency response of the amplifier. This is accomplished by using a value of capacitance which has negligible reactance at the lowest frequency to be passed by the amplifier. However, if desired, the value of the capacitor in the feedback loop may be changed to obtain a rising characteristic at the lower frequencies.

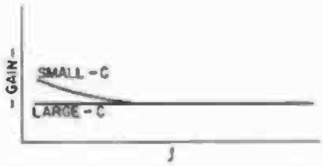


Fig. 12-218. The effect of decreasing the value of the feedback capacitor in a negative-feedback amplifier.

Assuming the value of the capacitance in the original design was selected for a flat frequency response, reducing the value of the capacitor results in less feedback at the lower frequencies; therefore, the gain of the amplifier at the lower frequencies is greater, resulting in a rising low-frequency characteristic. The effect of varying the value of capacitance in a feedback loop for a given set of conditions is shown in Fig. 12-218.

12.219 What is the purpose of connecting small capacitors from the grid or plate circuit to ground in a push-pull amplifier stage?—To prevent the generation of parasitic oscillations. If the amplifier stage employs a push-pull driver transformer, small capacitors may be connected across each half of the secondary, or from the control grid to ground. Both of these methods will affect the high-frequency response. However, if the amplifier is used only for paging or other services not requiring a wide frequency response, either method will be satisfactory. In either case, only enough capacity is used to prevent oscillation.

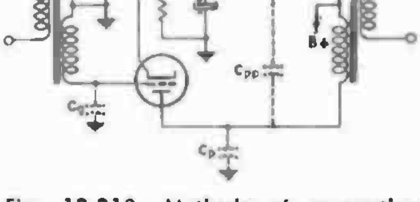


Fig. 12-219. Methods of connecting small capacitors in a push-pull amplifier to prevent parasitic oscillations.

Capacitors connected in the plate circuits may be somewhat larger in value than those used in the control-grid circuit because of the lower impedance of the circuit. At times, only a single capacitor connected from plate-to-plate will be necessary to prevent oscillation.

Parasitic oscillation is often caused by unbalances in the output transformer windings due to a greater amount of capacity existing between windings in one-half of the primary to ground. The connection of a single capacitor from one of the plates to ground to compensate for transformer unbalance will prevent oscillation. The proper plate must be found by trial.

A well-designed amplifier should not require compensation to prevent parasitic oscillation; however, because of manufacturing tolerances, this may be necessary at times. Typical methods of installing capacitors for the prevention of parasitic oscillations are shown in Fig. 12-219.

The values for the grid circuit vary from .00001 to .0005 μF. The value used, for the plate circuit is generally about .001 μF. Usually a .001 μF-capacitor is used if a single capacitor is connected from plate-to-plate. For a single capacitor from one plate to ground, use only as large a value as needed to prevent oscillation. The voltage ratings for the plate capacitors must be such they will stand the full dc plate voltage, and the peak signal voltage. Generally, 1000-volt mica capacitors are used for this purpose.

12.220 How should beam-power tubes be operated for the lowest harmonic distortion without negative feedback?—Beam-power tubes without negative feedback generally have from 7 to 13 percent total rms harmonic distortion, the greater part of the distortion being third harmonic. If the output load varies, as would be the condition when terminated by a loudspeaker, a capacitor and series resistor may be connected from plate to ground as described in Question 12.169, to help maintain a more constant load to the tubes.

A further reduction of harmonic distortion is obtained by operating the tube into a plate impedance one-fifth, or less, the plate resistance of the tube.

12.221 Why is the harmonic distortion increased when using high-frequency equalization?—Because the harmonic frequencies are accentuated and even small percentages of distortion may become objectionable. This type distortion is independent of any distortion induced by the equalized stage.

If the high frequencies are attenuated, the distortion is reduced. A good rule to remember is if the frequency response is to be extended, distortion must be reduced to a negligible amount.

12.222 How does the efficiency of an output transformer affect the power output of an amplifier?—At low-output powers the efficiency of the output transformer can be ignored; however, at powers of 10 watts and greater, the insertion loss of the transformer may become the limiting factor in the design of an amplifier. As an example, assume an output transformer of conventional design induces a power loss of 0.25 dB. For 20 watts of output power, the power loss between the primary and secondary amounts to approximately 1 watt.

If this same insertion loss occurred in an amplifier developing 100 watts of power, the loss between the primary and secondary would amount to about 5 watts of power. Therefore, it is extremely important that an output transformer used in an amplifier developing any amount of power have a low insertion loss or an efficiency as high as possible. This means that the transformer must have low leakage reactance.

Interstage transformers, as a rule, have about 1-dB insertion loss but, as these transformers are used in low-level circuits, it is of no particular consequence and the loss may be overcome by increasing the gain in the voltage amplifier stages 1 dB. The efficiency of an output transformer may be calculated:

$$E_{ff} = \frac{P_{sec}}{P_{pri}} \times 100$$

where,

P_{pri} is the power at the primary,
P_{sec} is the power at the secondary, operating with its normal load impedance.

Several new designs in output transformers have appeared in the last few years, in which the power loss is quite low. One of these designs is the McIntosh described in Question 12.231. The measurement of transformer loss is discussed in Question 23.71.

12.223 How is the insertion loss of a transformer calculated?—It may be calculated in two different ways—power or decibels. For the loss of power in watts, the equation is:

$$\text{Watts} = \frac{P_{pri}}{P_{sec}}.$$

For the loss in decibels:

$$\text{dB} = 10 \text{ Log}_{10} \frac{P_1}{P_2}$$

where,

P_1 is the power at the primary,
P_2 is the power at the secondary, when operating into its normal load impedance.

12.224 Can the frequency response of an amplifier be judged by its response to a square wave?—Yes. It may be said that if a square wave is applied to the input of an amplifier and not distorted by passage through the amplifier, the frequency response is uniform from f/10 to 10 f, where f is the fundamental frequency.

The square-wave generator is set to different fundamental frequencies and the shape of the square wave observed on an oscilloscope connected across the output of the amplifier. The frequency response of the oscilloscope must be such that it does not distort the square wave.

This can be checked by comparing the image at the output of the amplifier with that at the output of the square-wave generator. Square-wave testing is discussed in Question 23.154.

12.225 How are wideband vacuum-tube amplifiers compensated for frequency response?—As a rule, audio equipment requires a frequency response between 20 and 20,000 Hz. However, many devices used for testing audio-frequency equipment such as an oscilloscope, distortion-measuring sets, and vacuum-tube voltmeters require an extremely wide frequency response, ranging from 10 hertz to 1 megahertz in width. The conventional resistance-coupled amplifier will not amplify all frequencies equally over this wide band without special frequency compensating circuits.

The loss of low frequencies is due principally to the impedance of the bypass capacitor C_k in the cathode circuit rising as the frequency is decreased. This may be overcome to some extent by making the capacitor quite large, say 150 to 1000 μF.

Low-frequency response is also affected by the value of the coupling capacitor and the grid resistor used to couple the following stage or located at the input of the wideband stage. The time constant of this combination is also of importance. If the time constant is made long to pass very low frequencies, serious phase shift is caused which results in distortion of the lower frequencies. Phase shift at the lower frequencies may be avoided by using a direct-coupled circuit as described in Question 12.39.

The bypassing of the screen dropping resistor is also quite important, as it also affects the low-frequency response. To assure that the power supply impedance is low, a large capacitor should be connected across the dc output. As some of these capacitors would become rather impractical because of the large value required, it is more satisfactory to compensate the amplifier at the low frequencies.

High-frequency loss is due to the Miller effect, internal capacities of the tube, and the distributed capacities of the circuit which are difficult to overcome; therefore, compensation is also used for extending the high-frequency response. (See Question 12.191.)

Whenever compensation circuits are used, a loss of gain takes place, which may be compensated for by the use of tubes of high transconductance, such as the 6AC7.

An RC coupled stage with both a low-frequency and a high-frequency compensating circuit is shown in Fig. 12-225. Compensation is obtained by designing the time constants of the low-frequency compensating circuit to equal the time constant of the cathode circuit elements R_k and C_k. The compensating circuit consists of capacitor C_c and resistor R_c which increases the plate loading at the low frequencies and compensates for the phase shift which is caused by coupling capacitor C and grid resistor R_g.

To properly compensate the circuit, resistor R_c must have a value greater than the cathode resistor R_k by an amount equal to the voltage gain of the amplifier stage. Capacitor C_k in the cathode circuit must also be greater than C_c by the same amount, to obtain equal time constants in two parts of the circuit. Therefore:

$$R_c \times C_c = R_k \times C_k$$

where,

R_c is greater than R_k by the voltage gain of the stage.

To illustrate the design procedure, assume the amplifier stage has a gain of 18 at the midfrequencies, the cathode resistor is 150 ohms, bypassed by a 250-μF capacitor. The value of R_c is then:

$$18 \times 150 = 2700 \text{ ohms}$$

$$C_c = \frac{250}{18}$$

$$= 13.9\ \mu\text{F}.$$

At the low frequencies, the plate load consists of the plate-load resistor R_p and R_c in series, and at the higher frequencies principally R_p. At frequencies of 50 to 10 Hz, the compensating capacitor C_c will have a reactance of approximately 230 and 1200 ohms, respectively. In practice, the compensating resistor R_c is adjusted by applying a square wave signal to the input while observing the output waveform on an oscilloscope.

The high-frequency compensating circuit consists of a 70 to 250 millihenry slug-tuned peaking coil connected in the plate circuit. The frequency response at the high frequencies is adjusted in a similar manner to that used for the low-frequency circuit. When both circuits are properly adjusted, the waveform at the output of the amplifier should be the same as the waveform at the input. The compensating circuits, because of their design, are independent of each other.

12.226 Describe the dynamic characteristics of a class-A push-pull amplifier.—The dynamic characteristics of two tubes connected to operate in class-A push-pull are shown in Fig. 12-226. The characteristic curve has been constructed from the individual dynamic characteristics of each tube. The characteristics of the individual tubes may be identified as V1 and V2.

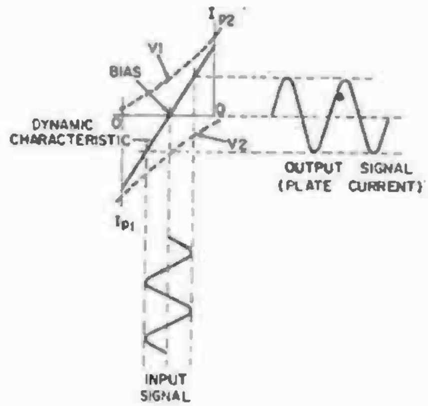


Fig. 12-226. Dynamic characteristics of two tubes operating class-A, push-pull.

The plot of the two tubes is the same because they are caused to operate 180 degrees out of phase. When operating as a single tube, class-A, little distortion exists because the grid signal operates in the most linear portion of the dynamic characteristic. In push-pull operation, the distortion is even less because of the increased linearity.

By projecting the signal at the control grid to the push-pull characteristic, the output waveform is obtained. As a rule, a greater grid voltage swing is possible in push-pull because the dynamic characteristics are linear over a greater voltage range. Thus, more power output is obtained.

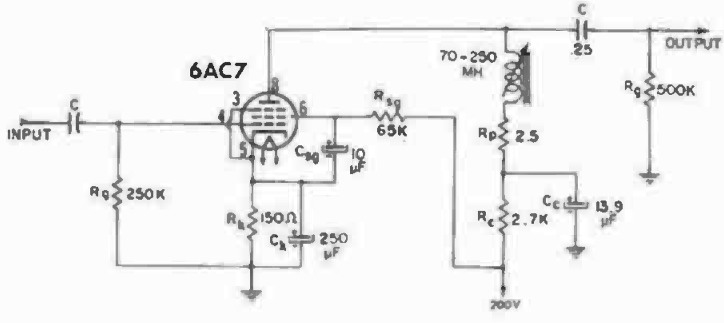


Fig. 12-225. A wideband RC amplifier compensated to cover a frequency band from 10 Hz to 1 MHz.

Efficiencies up to 30 percent are common with push-pull amplifiers where, with one tube only, 20 percent is average. Operating two tubes in class-AB can result in efficiencies up to 55 percent.

12.227 What are the dynamic characteristics of a class-AB$_1$ and -AB$_2$ push-pull amplifiers?—The dynamic characteristics are similar in appearance to that of the class-B$_1$ (Fig. 12-228A). The actual characteristic is a function of the bias voltage. As a rule, the output stage of a transistor amplifier is operated class-AB$_2$, class-B, or class-B$_1$.

12.228 Describe the dynamic characteristics of a class-B push-pull amplifier.—Although vacuum-tube amplifiers operating class-B and its subscripts are not as a rule used for driving recording equipment, extensive use is made of transistors operating in this mode. However, precautions must be taken to reduce crossover or notch distortion to a negligible amount. Amplifiers operating class-B are always designed for push-pull operation, as discussed in Question 12.67.

Two types of class-B amplifiers are in general usage, class-B$_1$ and class-B$_2$. In the class-B$_1$, grid current is not permitted to flow, while in the class-B$_2$ grid current is permitted to flow for a fraction of the input signal cycle; therefore, the harmonic distortion is higher than for the class-B$_1$ amplifier.

The dynamic characteristics of a typical class-B$_1$ push-pull amplifier are shown in Fig. 12-228A. A higher value of bias voltage is used than for the class-AB$_1$. It will be noted the dynamic curve for the class-B$_2$ amplifier (Fig. 12-228B) has a bend at the crossover point which induces harmonic distortion. Class-B amplifiers require very stable power supplies with good voltage regulation because the tubes in their quiescent state draw very little current; however, when a signal voltage is applied to the control grids, the plate current rises quite rapidly to a rather large value.

It is again emphasized that in the use of class-B amplifiers, crossover or notch distortion must be reduced to a negligible value for high quality recording and reprodnction.

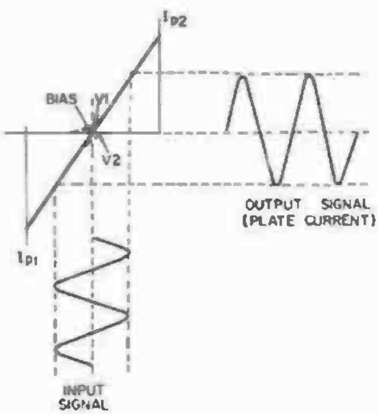


Fig. 12-228A. Dynamic characteristics of a push-pull amplifier operating class-B$_1$

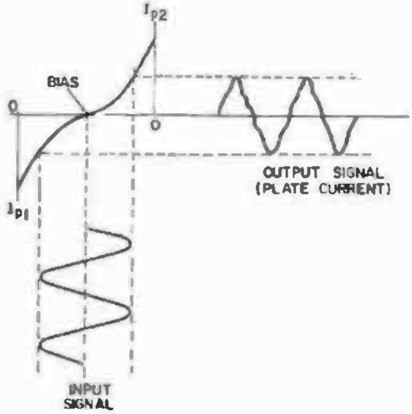


Fig. 12-228B. Dynamic characteristics of a push-pull amplifier operating class-B$_2$. Tubes are biased to plate-current cutoff.

12.229 How is fixed bias obtained?—Fig. 12-229A shows a fixed-bias circuit using a resistor R1 in the common negative return of a full-wave rectifier circuit normally used for supplying plate current. Because the total plate current drawn by the amplifier flows through resistor R1, a voltage drop is created which is used for a bias voltage. Only a small amount of filtering is required which may be obtained with a 100,000-ohm resistor (R2) and a 10-μF electrolytic capacitor. The fixed-bias may also be obtained by means of a rectifier connected to one-half of the high-voltage winding of the power transformer in Fig. 12-229B. Filtering is obtained by means of resistor (R) and a 20-μF capacitor C. The circuit in Fig. 12-229C is similar, except the ac voltage is obtained from a separate winding on the power transformer.

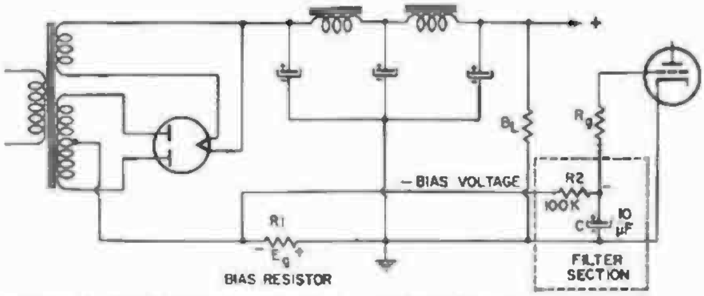


Fig. 12-229A. A fixed-bias circuit using a resistor in the negative return of a full-wave rectifier circuit.

12.230 Define the terms crossover and notch distortion.—When operating vacuum-tube amplifiers class-B or any of its subscripts, the bias voltage is adjusted for a condition of near plate-current cutoff. This type of operation is the cause of considerable distortion termed *crossover or notch distortion*, and occurs each time the signal passes through the zero plate-current point. This distortion is aggravated also by the imperfect coupling existing between the primary and secondary windings of the output transformer.

Crossover distortion can be corrected to a great extent by the application of a small amount of bias voltage applied in the cathode circuit of the individual tubes. For a 6L6 tube, this amounts to about 0.6 to 1.5 volts, and can be obtained by the use of a resistor of 5 to 10 ohms in each cathode circuit. After reducing the distortion in this manner, the remaining distortion can be further reduced by the use of a generous amount of negative feedback. As a rule, special class-B interstage transformers are employed to drive the output stage, with a low-impedance secondary because of grid-curreut flow.

Values of 1-percent or less THD are not difficult to obtain in a transistor amplifier for power outputs of 50 to 100 watts. What is important is the distortion at power output levels of 1 watt or less. Distortion at these levels can be up to five times greater than at full power output because of crossover or notch distortion. Distortion at the lower levels is usually generated in the output stage because of class-B operation. Therefore, it is highly important that several measurements be made at low power output to check for this type of distortion. The making of such measurements is discussed in Question 23.208.

Since an output transformer is seldom used with transistors, its contribution to the overall distortion is eliminated, and only the notch distortion has to be considered. The efficiency of a class-B push-pull stage ranges from 60 to 70 percent. (See Questions 11.143 and 23.210.)

12.231 Describe the McIntosh amplifier circuit.—The McIntosh circuit is a highly efficient beam-power tube push-pull amplifier circuit, using a special output transformer, developed and patented by Frank H. McIntosh. The

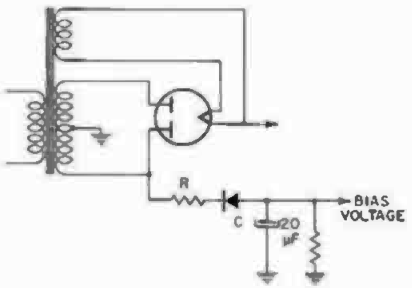


Fig. 12-229B. A fixed-bias circuit using a half-wave rectifier.

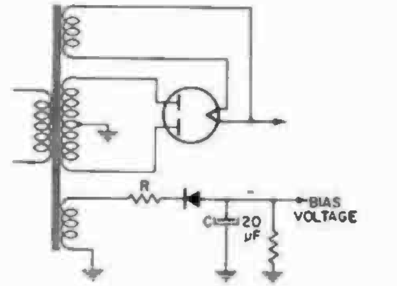


Fig. 12-229C. A fixed-bias circuit using a half-wave rectifier with a separate winding on the power transformer.

output stage attains its high efficiency through a combination of the output transformer and working approximately class-AB_2.

The transformer has four sets of windings, one connected in the cathode, one in the plate circuit, and two output windings, one of which is used for supplying a negative feedback voltage. A schematic diagram of the McIntosh MC-40 power amplifier is shown in Fig. 12-231A.

The output transformer of Fig. 12-231B provides two load impedances, one for the cathode and one for the plate. The primary and cathode windings are *bifilar* wound; that is to say, the two windings are made by winding the wires side-by-side, providing for all practical purposes unity coupling; thus, the leakage is greater than 80,000:1, which reduces the leakage inductance (lack of coupling) to a negligible amount. After the coils are wound,

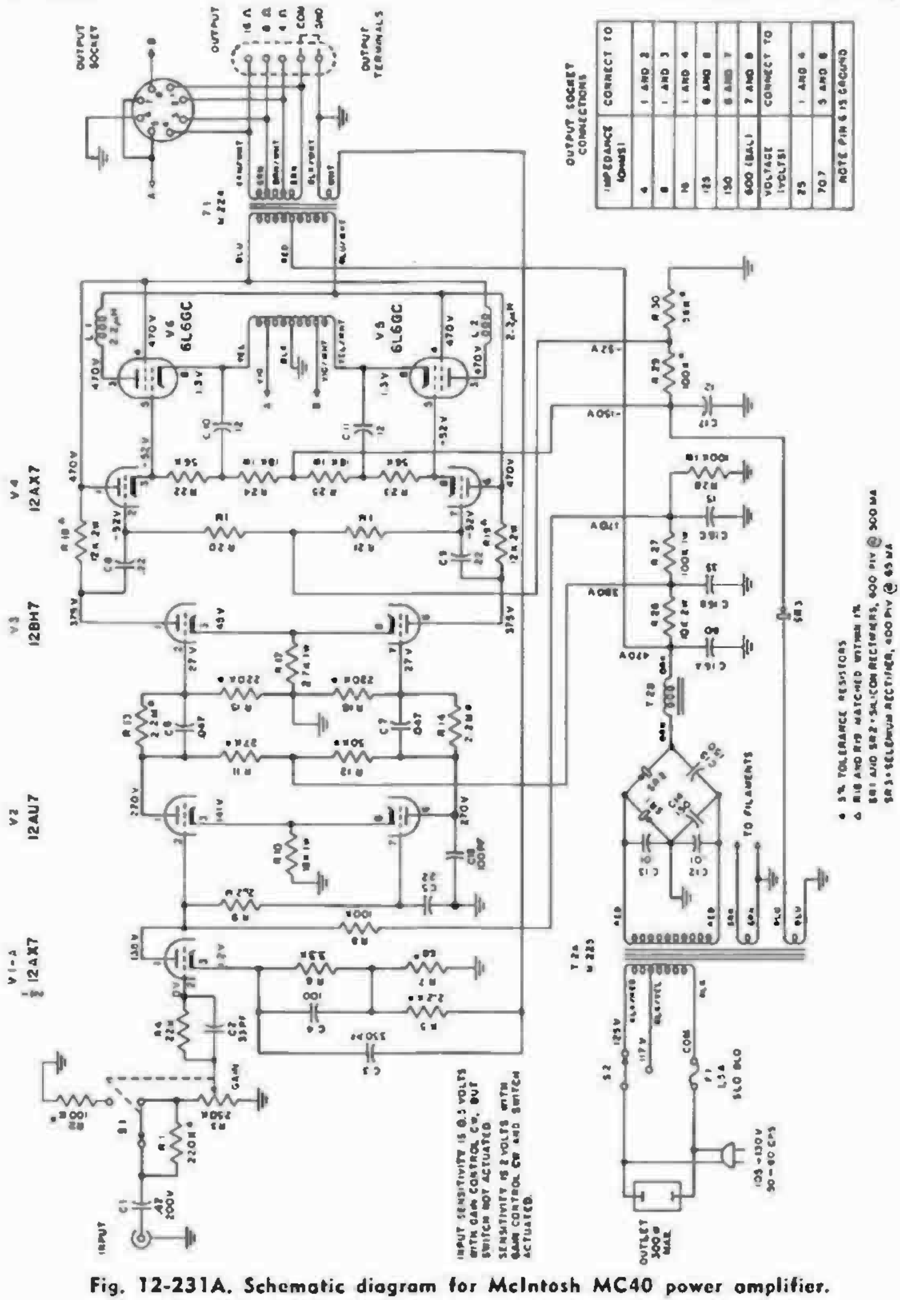


Fig. 12-231A. Schematic diagram for McIntosh MC40 power amplifier.

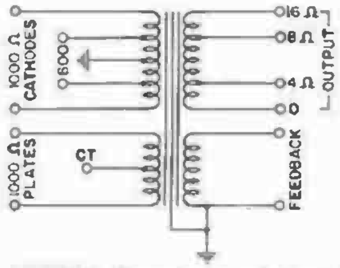


Fig. 12-231B. Impedance relations for McIntosh unity-coupled bifilar-wound output transformer.

grain-oriented steel "C" type cores are inserted in the coils, resulting in a shell-type construction as seen in Fig. 12-231C. (See Fig. 8-105.) The two cores weigh approximately five and one-half pounds each. The power bandwidth is thus increased to over 100,000 Hz.

In operating tubes push-pull class-A, class-B_2, and class-B distortion is generated, termed notch or crossover distortion. (See Question 12.230.) This distortion is particularly bad in the class-B modes. Notch distortion is generated at the instant the current is cut off during the idle period, and again when current starts to flow. This is caused by the imperfect coupling in the conventionally wound transformer. The use of a bifilar winding provides an extremely tight coupling between windings, removing a condition which aggravates notch distortion. Furthermore, the tubes are connected as partial cathode followers, half of the circuit being in the cathode and half in the plate circuit of each tube. This provides a local feedback loop which reduces distortion and generator impedance, and eliminates the need for balancing the circuits. The primary and secondary windings of the coils are interleaved no less than five times, to improve the coupling. Interleaving is accomplished by winding groups of primary layers, then secondary layers, until the whole transformer is completed.

Referring again to Fig. 12-231B, the small 5-ohm winding supplies a negative-feedback voltage, while the fourth winding is employed for coupling to the external-load circuit. Taps are provided at 4, 8, and 16 ohms. A 600-ohm output is also provided by the winding in the cathode circuit. Because of a small amount of dc current flow between the cathode and the center-tap gronnd connection, this winding cannot be used with grounded circuits. Only a symmetrical or balanced-to-ground circuit may be used. (See Question 8.25.) The small voltage drop across this winding has the effect of reducing the notch or crossover distortion mentioned earlier.

It will be observed in Fig. 2-231A that the screen grids are connected to the top end of each half of the primary winding, but the plate of a given tube is fed from the opposite end of the primary winding. Thus, the signal voltage at the cathode and the screen grid for a given tube is the same. Connected as shown, the screen grids are bypassed to the cathode providing ideal conditions for beam-power tube operation. The impedance relationship existing between the several windings is shown in Fig. 12-231B. It will be noted the impedance of the primary from plate-to-plate is 1000 ohms, which is quite low compared to the conventional push-pull amplifier where the plate-to-plate impedance may range from 4000 to 10,000 ohms. The impedance from cathode-to-cathode is also 1000 ohms. Because the signal voltage at the cathodes is quite high, the circuit requires tubes in the output stage that will withstand a high

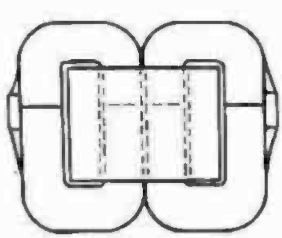


Fig. 12-231C. Interior construction of McIntosh output transformer. The coils are wound on grain-oriented steel "C" cores.

Fig. 12-231D. McIntosh MC40 power amplifier.

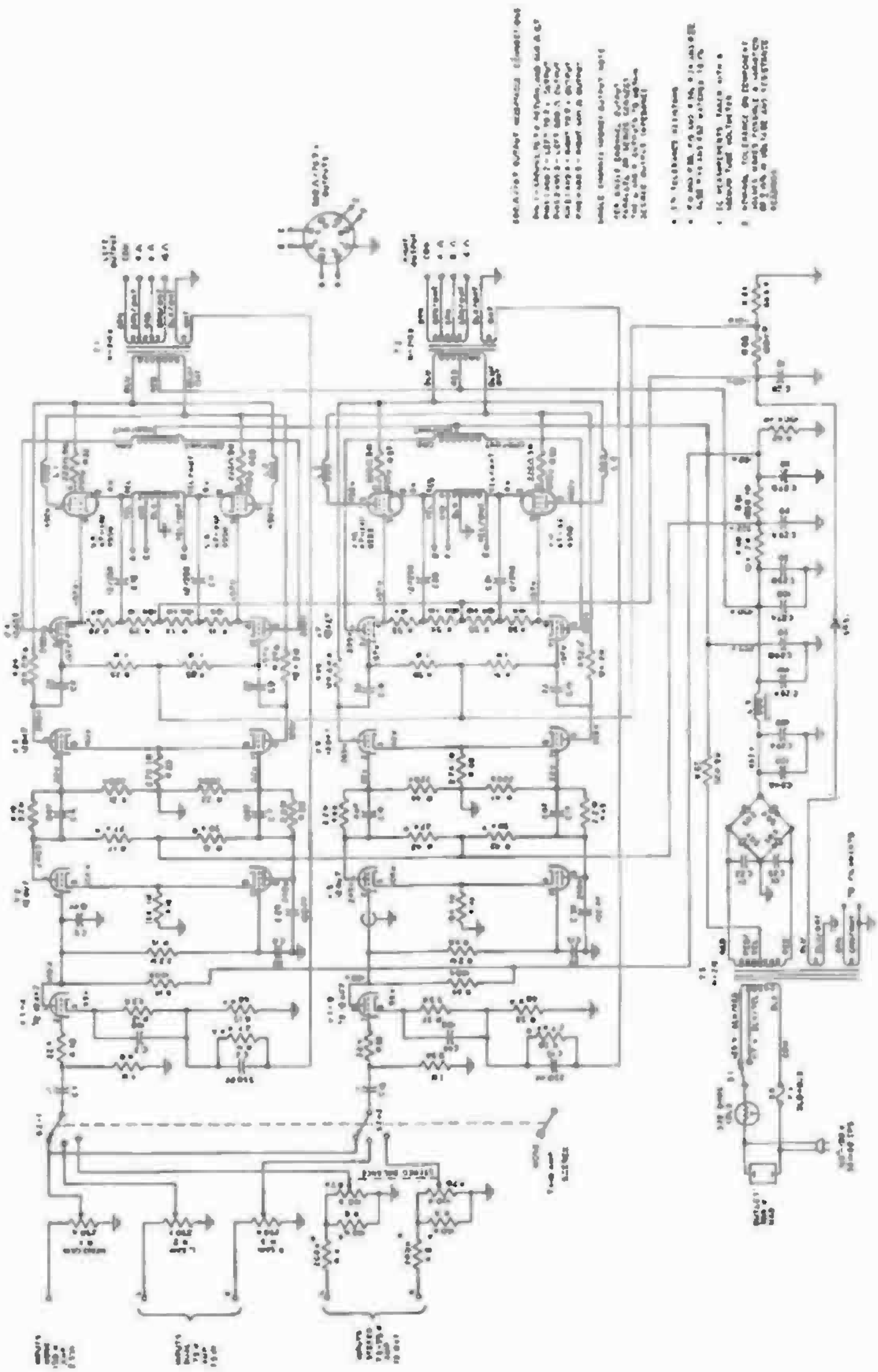


Fig. 12-231E. Schematic diagram for McIntosh MC275 stereophonic amplifiers.

heater-to-cathode potential. The 6L6G beam-power tube meets these requirements.

Referring again to Fig. 12-231A and tube V1A the input signal is applied to a single-ended amplifier stage, consisting of one-half of a 12AX7 tube, followed by a "long-tailed" phase inverter comprised of a 12AU7 tube. This stage is followed by a second push-pull stage, 12BH7 tube which drives a third push-pull stage, a 12AX7 tube connected as a cathode follower. The cathodes of this stage are direct-coupled to the control grids of the 6L6G beam-power tubes in the output stage.

Negative-feedback voltage is taken from a 5-ohm winding and carried back to the cathode circuit of the first stage. The 6L6G tubes in the output stage are operated with fixed bias supplied by a series half-bridge employing two silicon diode rectifiers, and two capacitors. (See Question 21.98.) Small radio-frequency chokes of approximately 2 microheuries are connected in the plate of each output tube to prevent parasitic oscillation. An input control and attenuator complete the circuit. The rated continuous power output is 40 watts, with less than 0.5-percent harmonic distortion at 20 to 20,000 Hz, and in the midfrequencies a 0.10-percent distortion. Intermodulation distortion is less than 0.5 percent for any combination of frequencies between 20 and 20,000 Hz, if the instantaneous peak power is less than 80 watts. Hum and noise is 90 dB below the output of 40 watts (plus 46 dBm). The internal output impedance is less than 10 percent of the load impedance. If the amplifier is to be used with a 70.7-volt distribution system, the load is taken from taps A and B on the 600-ohm winding in the cathode circuit of the output tubes. Phase shift is less than 8 degrees at 20 to 20,000 Hz. The completed amplifier appears in Fig. 12-231D.

To provide large amounts of power with good frequency response requires that an amplifier have a wide frequency range. To provide low distortion with a large amount of power requires a uniform frequency response with small amounts of phase shift. This design assures that the feedback voltage at fundamental high frequencies will maintain the proper phase relationship with the higher harmonics, thus permitting the maximum use of the feedback voltage loop to cancel distortion and improve the linearity. Small changes in the frequency response such as 0.10 dB will cause a phase shift of 10 degrees, and a change of 3 dB will cause a phase shift of 45 degrees. Therefore, the effect of the feedback is reduced in proportion to the cosine of the phase shift angles. As a rule, phase shift becomes apparent and measurable at one-seventh to one-teuth the frequency where a 2-dB change in the frequency response takes place. This indicates quite clearly that a bandwidth 7 to 10 times the highest frequency to be reproduced is necessary.

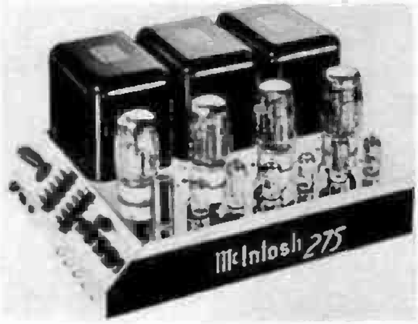


Fig. 12-231F. McIntosh MC275 stereophonic power amplifiers. Each channel develops 75 watts of power.

The schematic diagram for a dual 75-watt version, the McIntosh MC275, is shown in Fig. 12-231E. Except for the values of certain components, and the output tubes being changed to KT-88 or 6550 tubes, the theory and operation of the circuit is the same as for the MC40. An additional winding appears on the output transformer, which is fed from the plates of the driver stage V4, and is *trifilar* wound, improving the power

Fig. 12-231G. McIntosh Model C24 solid-state stereophonic preamplifier.

bandwidth. The output power rating for this amplifier is 75 watts continuously for each channel, or 150 watts monophonic. The frequency response, harmonic and intermodulation distortion, and phase-shift are the same as for the MC40. The complete amplifier is pictured in Fig. 12-231F. Either of the described amplifiers are designed to operate with any type preamplifier or stereo center.

In Fig. 12-231G is shown a solid-state preamplifier (stereo center) Model C24, especially designed to operate with the previously described amplifiers. In the schematic diagram of Fig. 12-231H, the preamplifier employs a total of 18 silicon planar transistors, eight in each side, with one in a third L + R channel (for driving an additional amplifier and loudspeaker) and one in the voltage regulator circuit in the power supply. The preamplifier section of the phono pickup and magnetic tape head employs three of the eight transistors. Standard RIAA or LP equalization may be selected by a slide switch. This is followed by a mode-selector switch, loudness control, channel balance, the first section of the volume control, and an emitter-follower transistor driving a Baxandall-type tone-control circuit. (See Question 6.86.) This is followed by two more amplifier stages, rumble and noise filter, and an output stage using two transistors. The second section of the volume control is next, followed by a voltage divider network feeding the L + R third channel.

The power supply consists of two silicon diodes in a full-wave rectifier circuit. The output voltage is regulated to 75 volts and further regulated by a zener diode to 10 volts. A second zener diode is connected in the emitter of the voltage-regulating transistor. An unusual feature of this unit is that the output circuits of the power amplifiers are returned to this unit before going to the loudspeaker voice coils. This permits the user to reverse the phasing of one speaker if required, and feed a pair of stereo headphones from a network connected between the left and right channels. The frequency response of the complete unit in the flat position is within 0.5 dB, from 20 to 20,000 Hz. The output is rated 2.5 volts for less than 0.10-percent distortion, with a slight increase to 0.30 percent at 10-volts output. The hum and noise is 78 dB below 2.5-volts output.

12.232 Describe a Williamson circuit using ultralinear operation.—In Fig. 12-232A is shown a schematic diagram for the well-known Williamson power amplifier using ultralinear operation. Starting at the left, the signal from a preamplifier is applied to the input control grid of tube V1, through a network consisting of a 15,000-ohm resistor, 0.10-μF capacitor, grid resistor, and a 15,000-ohm grid-limiting resistor. The 15,000-ohm series resistor in the input increases the input impedance of the amplifier. The 0.10-μF capacitor prevents any dc component in the signal from the preamplifier from affecting the control-grid circuit of the power amplifier, as would be the case if the preamplifier did not include a dc isolating capacitor. The 15,000-ohm grid-limiting resistor in the control grid prevents overloading of input stage if too great an input signal voltage is applied from the preamplifier (as would be the case if the signal were taken from the plate circuit of the output tube in a preamplifier).

The output of V1 is direct-coupled to the control grid of V2, a phase splitter employing equal values of load resistance in both the plate and cathode circuits. Because of the large value of resistance in the cathode circuit of V2, an 88-volt positive potential is developed at the cathode. The control grid of this tube, being connected to the plate of V1, is also 88 volts positive. This means the control grid is operating at a zero potential. By direct-coupling to the input of the phase-inverter tube V2, phase shift at the lower frequencies due to the presence of a coupling capacitor is avoided.

The signal at the cathode of V2 is in phase with the signal at the control grid, while the plate signal swings in the opposite direction. These two signals are of equal amplitude but 180 degrees out of phase. This balanced signal is applied to the control grids of the driver stage V3 and V4, through coupling capacitors of 0.10 μF each. Two signals of equal amplitude and 180 degrees out of phase are again available at the output of V3 and V4 and are applied to the control grids of the 5881 or 6L6G push-pull output tubes through

1.0-μF coupling capacitors. The use of large coupling capacitors increases the low-frequency response, besides providing a 10:1 ratio of time constants between the two amplifier stages and thus reducing the tendency to oscillate at the low frequencies. The signal is finally amplified by the output tubes

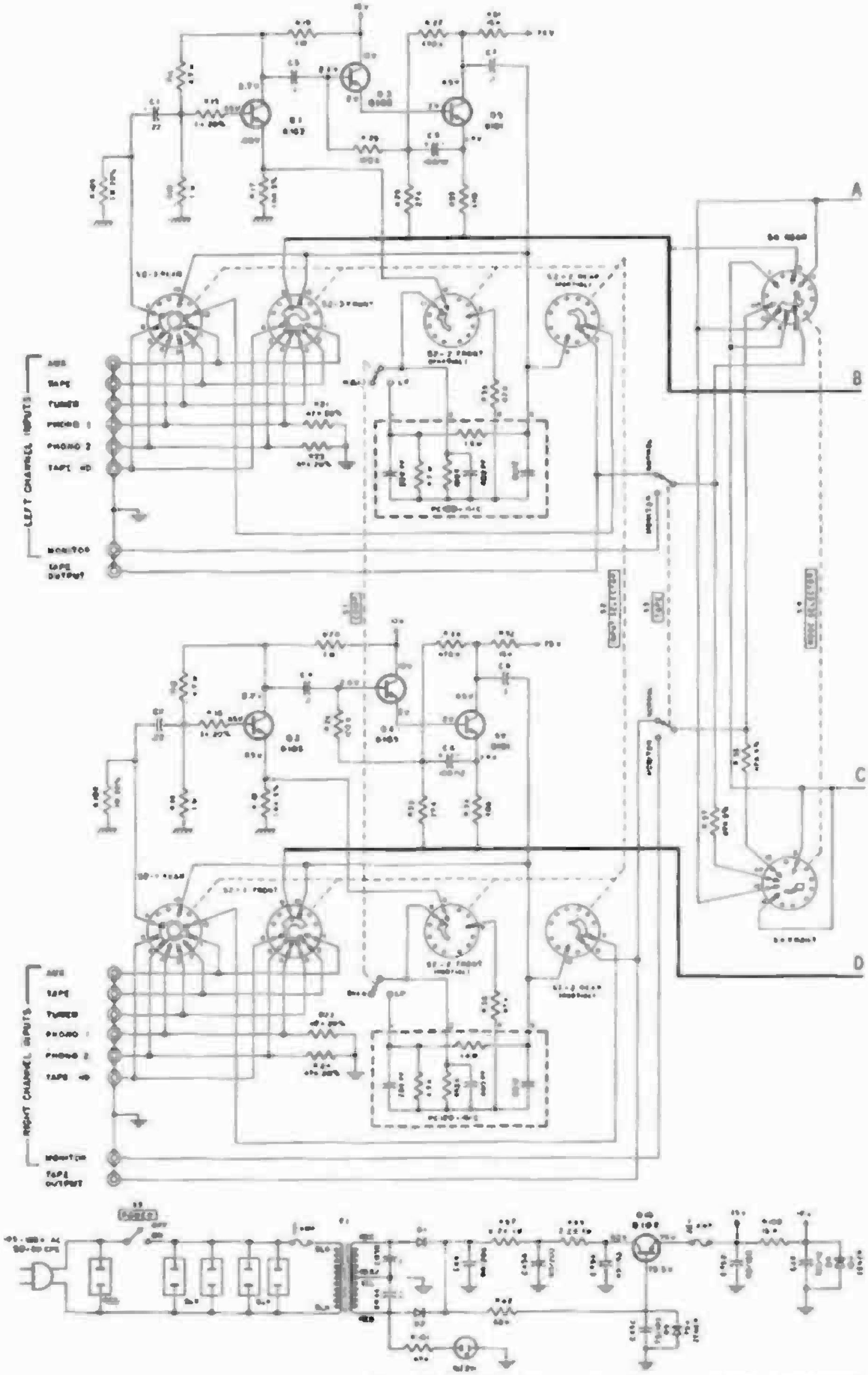
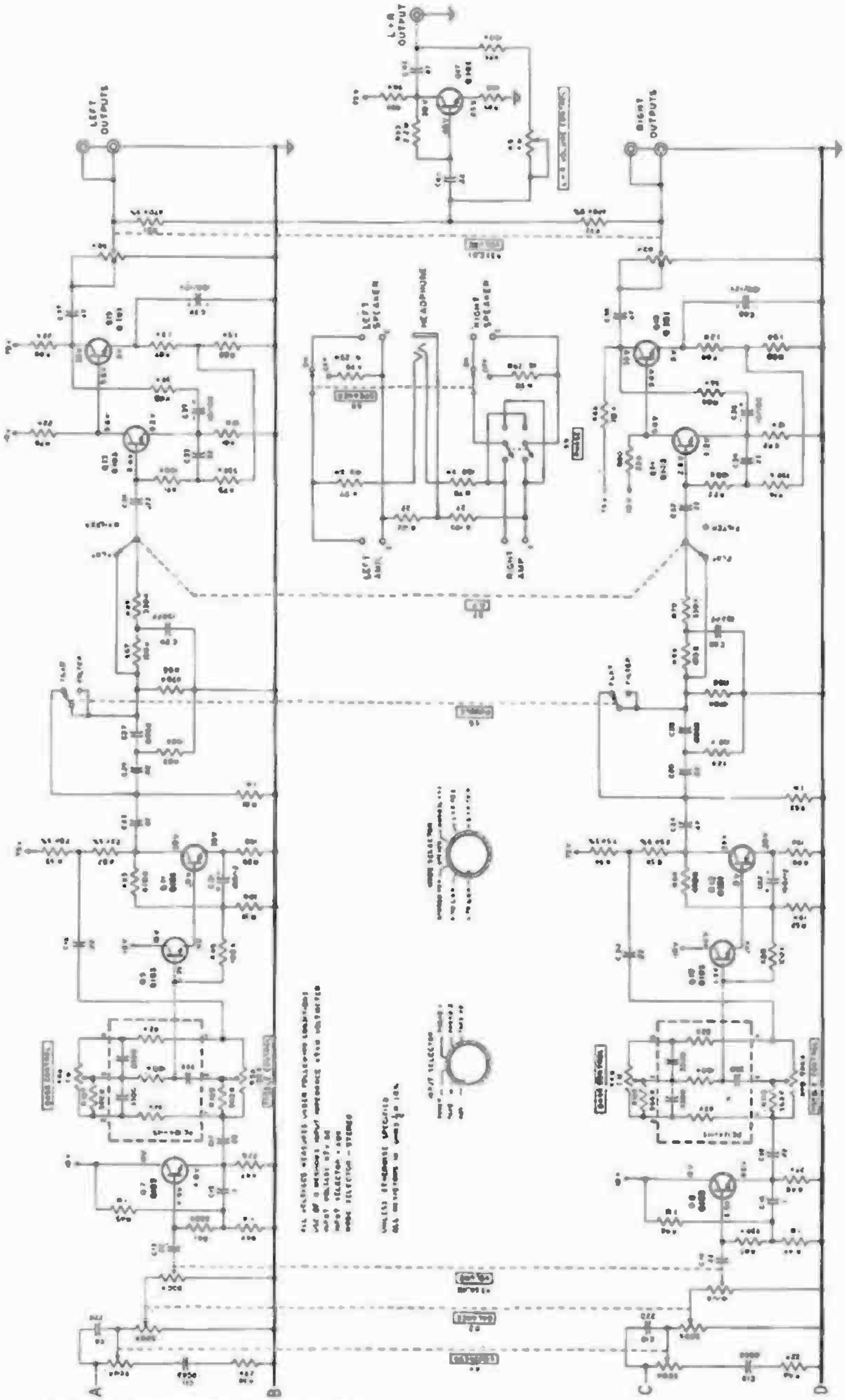


Fig. 12-231H. Schematic diagram of the McIntosh



Model C24 stereophonic preamplifier.

and applied to the output load by means of the output transformer T1.

Negative-feedback voltage is taken from the output winding of T1 and carried back to the cathode of the first stage V1A, through a 4700-ohm series resistor shunted by a 240-pF capacitor, which increases the high-frequency response above 50,000 Hz and assures a 180-degree phase shift at the higher frequencies around 100,000 Hz.

To obtain the lowest distortion with this circuit, the plate currents of the output tubes must be balanced as closely as possible. To achieve this balance, a network consisting of potentiometer P1 and several fixed resistors is connected in the control-grid return circuit of the push-pull input circuit. A pin jack is connected to the upper end of each cathode element. To balance the plate currents, a dc voltmeter is connected across the pin jacks and the potentiometer P1 adjusted for a minimum or zero indication on the voltmeter. The accuracy of the volt-

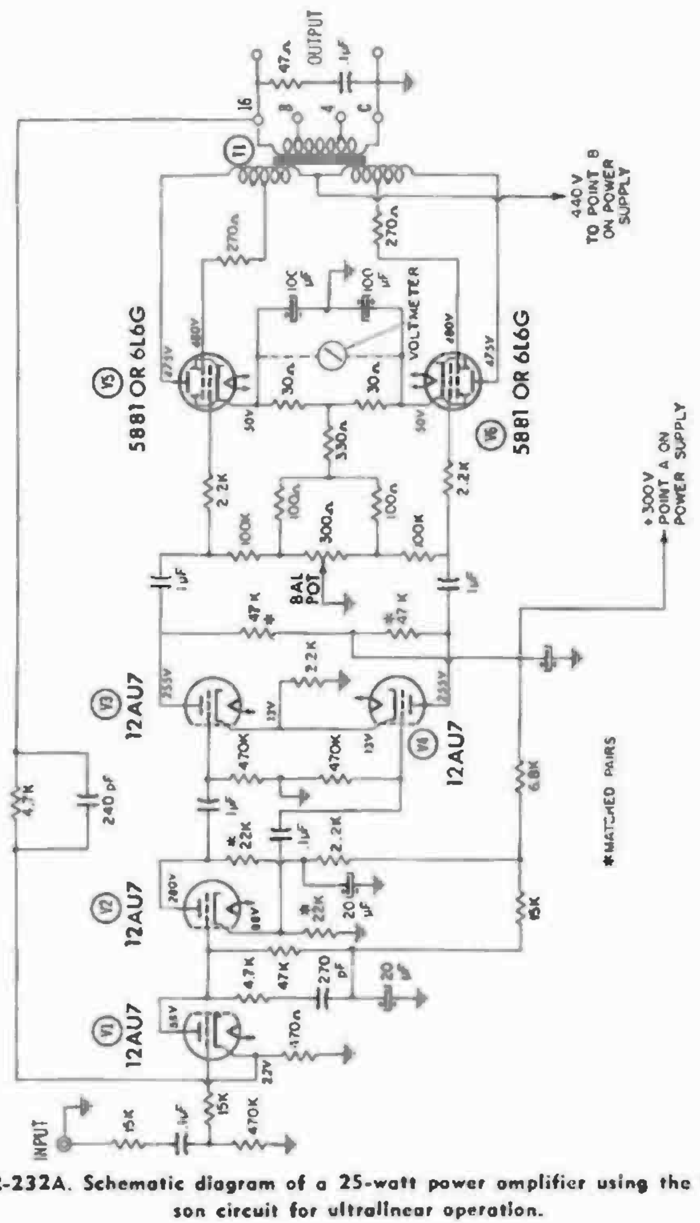


Fig. 12-232A. Schematic diagram of a 25-watt power amplifier using the Williamson circuit for ultralinear operation.

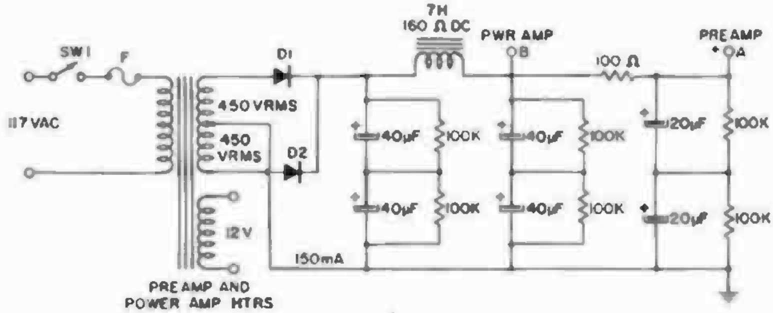


Fig. 12-232B. Schematic diagram for Williamson 25-watt ultralinear amplifier power supply.

meter does not enter into the adjustment, as it is used only as a null indicator. When a balance is achieved, the voltage drop across each cathode resistor is the same; therefore, the plate current in each tube is the same. The potentiometer P1 minimizes the unbalanced dc in the primary of the output transformer, reducing the core saturation. Thus, the low-frequency response is increased and the distortion below 50 Hz reduced. (See Question 23.45.)

A limiting circuit, consisting of a resistor and capacitor connected in series and in parallel with the secondary of the output transformer T1, prevents damage to high-frequency speaker units caused by transient disturbances. The values of resistance and capacitance will not affect the frequency response in the normal frequency spectrum.

Although the power supply for the above combination is shown in a separate diagram, Fig. 12-232B is actually a part of the power amplifier assembly, and consists of a power transformer, two diode rectifiers, and an LC and RC filter section. The high voltage for the output tubes is taken at point B and for the preamplifier at point A. To obtain a high value of filter capacitance in the filter circuits and still have a large margin of safety relative to the operating voltage, electrolytic capacitors of 40 μF are connected in series at the input and output of the filter choke, with resistors of 100,000 ohms connected across each capacitor, to assure an even voltage distribution across each capacitor. The 8-pin plug at the right of the diagram is for providing power to the preamplifier.

Fig. 12-232C is a plot of power output versus intermodulation distortion. The frequency response for 5 and 25 watts is given in Fig. 12-232D, with the phase-shift characteristics given in Fig. 12-232E. Hum and noise is 99 dB below the output level of 25 watts (plus 44 dBm), or 45 dB with reference to 1-milliwatt.

12.233 How is the bandwidth of an amplifier designated?—It is stated as the difference between the half-power point at the upper and lower frequencies, or the frequency of 3 dB loss with respect to the flat portion of the passband as shown in Fig. 12-233.

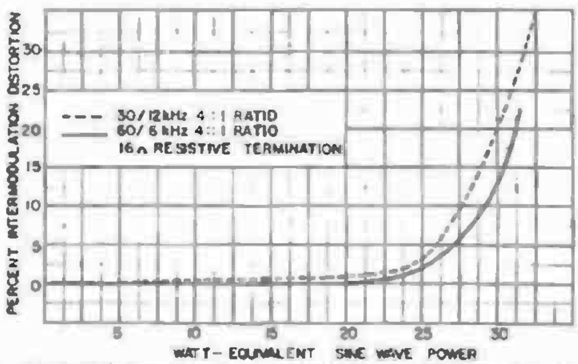


Fig. 12-232C. Power output versus intermodulation distortion.

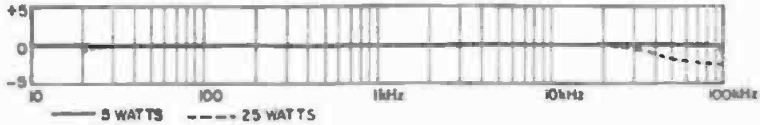


Fig. 12-232D. Frequency response of the power amplifier in Fig. 12-232A.

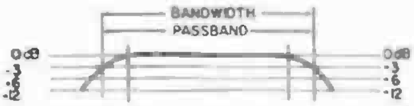


Fig. 12-233. Bandwidth of a transformer.

12.234 What is the purpose of connecting a small capacitor in parallel with the resistor in a negative-feedback loop. —It is employed as a phase-shifting device to assure the signal being 180 degrees out of phase at the midrange frequencies. Phase shift at the high frequencies is maintained by the stray capacitance due to components and internal wiring.

12.235 What precautions should be observed when mounting coupling capacitors in a wide-range amplifier?—They should be mounted up and away from the metal chassis to prevent the distributed capacity from affecting the high frequencies. Metal bathtub type capacitors should be avoided for interstage coupling.

12.236 What is the purpose of paralleling an electrolytic capacitor with a paper or mica capacitor?—To nullify the inductive effect of electrolytic capacitors, thus providing better bypassing at the higher frequencies. Generally, about 0.10 to 0.025 μF will suffice for frequencies up to 50,000 Hz.

12.237 How does a hum potentiometer across a heater winding reduce the hum frequencies in the output?—When the pot is in balance, equal voltages but opposite in sine are fed to the control grid by induction from the heater circuit cancelling out the induced hum frequencies.

12.238 What is the principal cause of distortion in an output transformer? —When the maximum operating flux is reached, the core material cannot produce any more lines of force because of saturation. Under these conditions strong harmonics are produced, the order being dependent on the amplifier design.

12.239 How is the input level of an amplifier calculated to drive it to full output?—Assume an amplifier is stated to have 64 dB gain and to be capable of developing a plus 43 dBm at the output, what level at the input is required to drive it to plus 43 dBm?

$$\begin{aligned} dB_{in} &= (dB_{gain} - dB_{out}) \\ &= (64 - 43) \\ &= 21 \text{ or } -21\ dBm_{in}. \end{aligned}$$

Thus a signal level of minus 21 dBm at the input will drive the amplifier to a plus 43 dBm.

12.240 What is the equation for calculating the voltage gain in decibels when an ideal transformer of a given impedance ratio is inserted between a generator and the load?—The gain in decibels may be computed:

$$dB = 20 \operatorname{Log}_{10} \sqrt{Z_r},$$

or $$= 20 \operatorname{Log}_{10} T_r,$$

or $$= 10 \operatorname{Log}_{10} Z_r,$$

where,

Z_r is the impedance ratio of the transformer,

T_r is the turns ratio of the transformer.

12.241 If the output stage of a voltage amplifier employs a screen-grid tube and must be transformer-coupled at the output, what type transformer is recommended?—If the tube is a 6J7 or similar type, an output transformer with a 30,000-ohm impedance primary

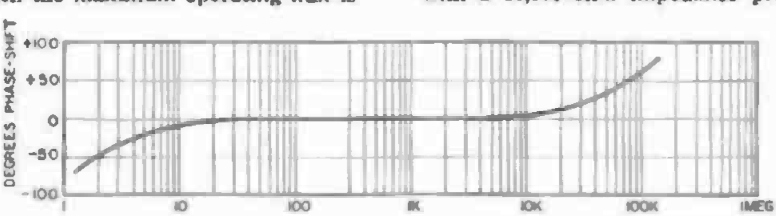
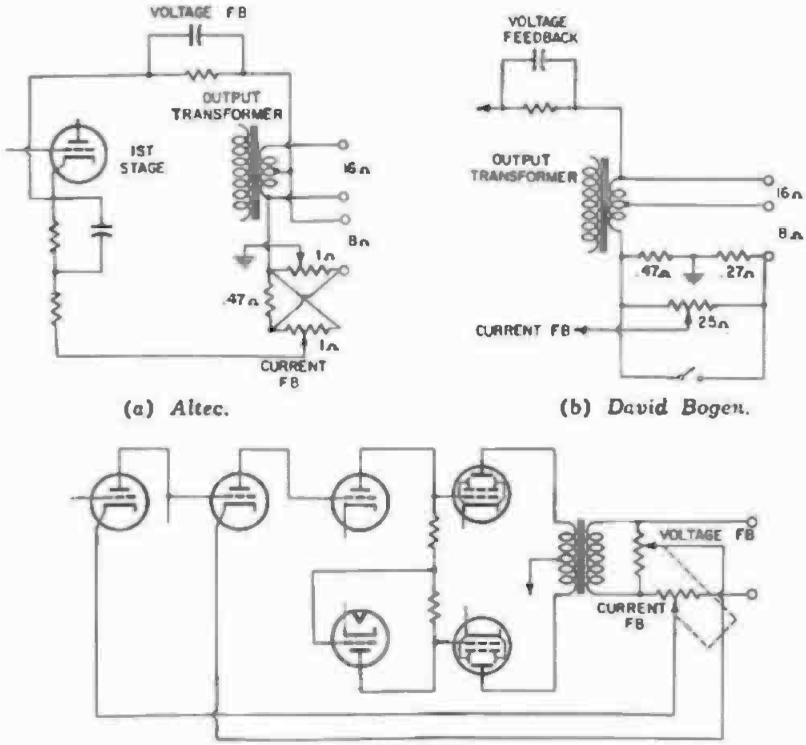


Fig. 12-232E. Phase shift of the power amplifier in Fig. 12-232A.



(a) *Altec.* (b) *David Bogen.*

(c) *Scott.*

Fig. 12-242. Variable damping circuits.

may be used, with a secondary impedance to suit the requirements. The load supplies the terminating impedance.

12.242 What is variable damping and how is it used?—Variable damping is used in the output circuit of an amplifier for the purpose of matching the loudspeaker impedance to that of the amplifier. The damping factor of the amplifier is varied by means of a control which permits the use of either negative voltage or current feedback. Three typical variable damping control circuits are shown in Fig. 12-242. It is claimed by the designers of these circuits that the various effects noted in loudspeakers because of impedance variations are eliminated by the use of variable damping. Also, that when properly adjusted, the amplifiers will present the correct impedance match to the loudspeaker voice coil at all times. Variable damping circuits have been the subject of much discussion but have not yet been accepted as the remedy for the mismatch between the amplifier output and the loudspeaker. Measurements of amplifiers using variable damping are discussed in Question 23.139.

12.243 Show the circuit for a class-AB push-pull amplifier using fixed bias.—Such a circuit is shown in Fig. 12-243 for an amplifier capable of producing 30 watts of audio frequency power. The output stage is driven from a cathode-coupled phase inverter. It will be noted the negative bias is obtained from a small half-wave rectifier and is applied to the cathode return of the inverter stage. The amplifier, although it has considerable harmonic distortion, is quite suitable for public address work.

12.244 How are harmonics generated in a nonlinear amplifier?—Harmonic distortion in an audio amplifier may be caused by several factors, such as the tubes not operating on the straight portion of their characteristic curve, operating in an overloaded condition caused by applying too great a value of signal voltage to the input resulting in a flow of grid current, improper operating voltages, or poor regulation in the power supply.

Amplitude distortion is caused by the nonlinear action of the tubes. If a pure

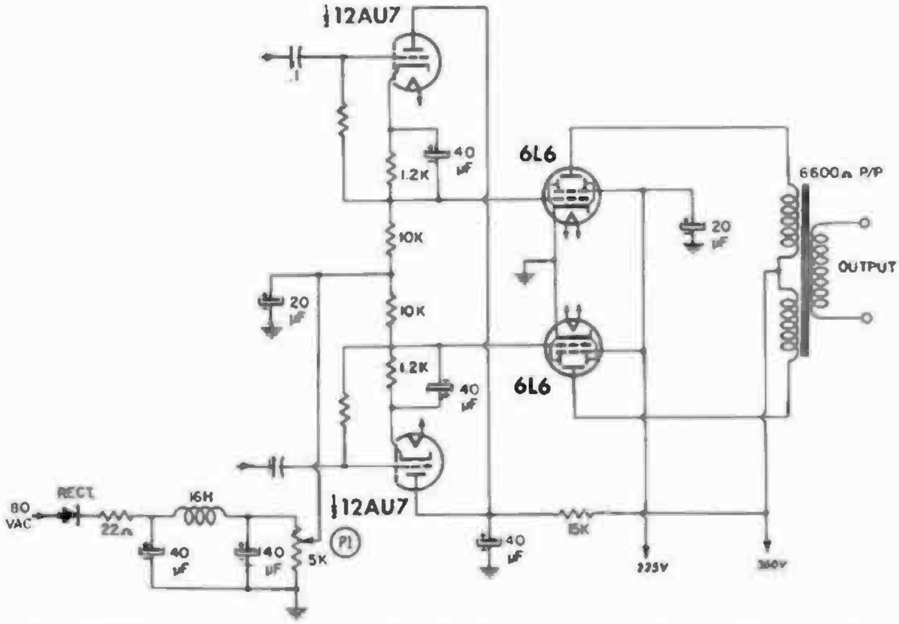


Fig. 12-243. A cathode-follower phase-inverter and push-pull amplifier operating class-AB_2 using fixed bias.

sine wave (a single frequency which contains no harmonics) is applied to the input of a vacuum tube which is supplied with the correct operating voltages and the amplitude of the sine wave is such that it does not overload the input, a pure sine wave may be expected at the output or plate circuit. The only departure from the original waveform at the input will be the increase of amplitude and a phase reversal of 180 degrees.

To illustrate the cause of nonlinear distortion, the plot of the grid-voltage plate-current characteristics of a general purpose triode biased for class-A operation is shown in Fig. 12-244A. The tube selected for this illustration requires a normal bias voltage of minus 6 volts. This point is indicated on the plate curve as the normal operating point.

Operating under the foregoing conditions, if a sine wave is applied to the control grid, a sine wave is reproduced in the plate circuit. However, this will

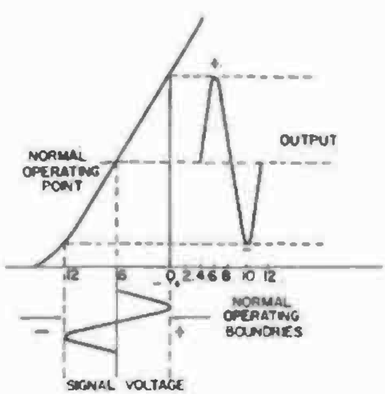


Fig. 12-244A. Plate current for a medium triode that is biased for class-A operation.

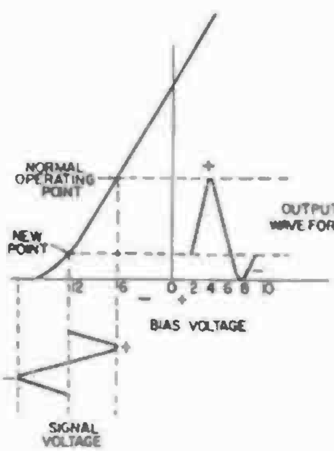


Fig. 12-224B. Output waveform distortion caused by overbiasing.

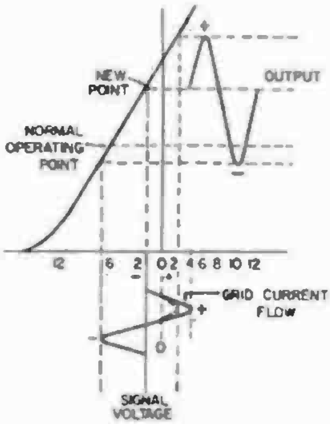


Fig. 12-244C. Output waveform distortion caused by underbiasing.

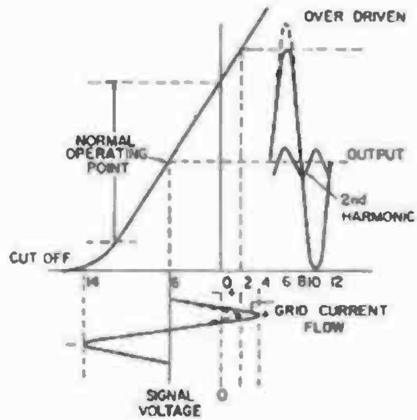


Fig. 12-244D. Distortion in a correctly biased tube caused by the application of too large a signal voltage at the control grid.

only hold true if the applied signal voltage does not exceed the normal operating boundaries. These boundaries are the points on the straight portion of the curve starting at minus 12 volts bias and continuing up to the zero bias point. If the signal voltage does not exceed these limits, a sine wave will be reproduced in the plate circuit; hence, no distortion of the input signal will occur.

Now consider an amplifier in which the bias voltage is incorrect, or too large, as shown in Fig. 12-244B. Here the negative bias voltage has been increased until the operating point (value of plate current) has been shifted downward towards the toe or bend in the plate current curve, or to the minus 12-volt point. Under these conditions, if a sine wave is applied to the control grid there will be a greater flow of plate current for the positive half of the grid swing than for the negative half, resulting in an unsymmetrical output waveform in the plate circuit.

If the bias voltage is reduced to a value of minus 2 volts, as shown in Fig. 12-244C, the operating point will be shifted upward. Now when the control grid swings positive, the amplitude of the positive half of the plate-current waveform will be less than the negative half because of being driven into the saturation region of the plate-current characteristic. When the control grid is driven above the zero point, it will draw grid current because the grid is now positive with respect to the cathode. This flow of grid current causes a voltage drop across the load in the grid circuit, and as this voltage is in opposition to the signal voltage, the amplitude of the positive peak of the signal voltage is reduced. This region is indicated by the dotted line in the positive half of the signal voltage at the control grid. Conditions as described in the foregoing cause strong second harmonics to be generated within the tube, distorting the output waveform in the plate circuit. Thus it is apparent the control-grid bias voltage and the amplitude of the applied signal play an important part in controlling the distortion in a vacuum tube.

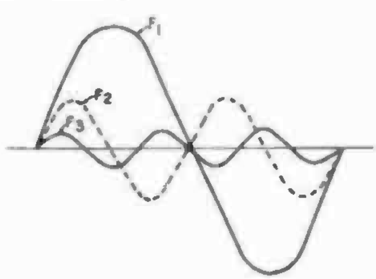


Fig. 12-244E. A fundamental frequency F_1, and its second and third harmonics F_2 and F_3.

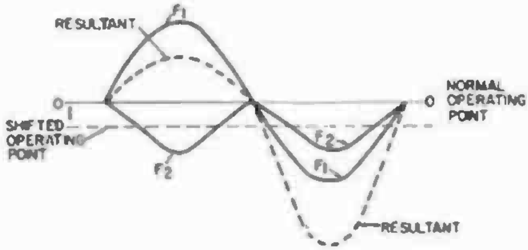


Fig. 12-244F. Adding the fundamental frequency F_1 and the second harmonic F_2 of Fig. 12-244E produces a new frequency-designated resultant.

The effect of overdriving the control grid by applying too great a signal voltage is shown in Fig. 12-244D. In this illustration the tube is again biased to operate at its normal operating point using a minus 6 volts at the control grid.

For the sake of illustration, the signal voltage has been increased to a point where it drives the control grid into the positive region or saturation and cutoff, the negative toe of the plate current curve. Driving the grid so far negative practically cuts off the plate current on the negative swings of the control grid. This action flattens off the negative peak, while the positive peak is driven into saturation and flattened off. (Saturation is a point where the plate current is increased to such proportions that no further output is obtained.) The slightest flattening or distortion of either the negative or positive peak will cause distortion of the output waveform producing both even and odd order harmonics.

Harmonics generated in a vacuum tube because of overloading is the result of inherent characteristics of the tube. First the plate resistance is not always uniform and secondly, the so-called straight portion of the plate current curve is not really straight, but has some curvature. The most objectionable harmonics are the second, third, fifth, and seventh because they are generally of higher amplitude and fall in the audible band.

To further illustrate the generation of harmonics within an amplifier, in Fig. 12-244E are shown three superimposed sine waves. F_1 is a fundamental frequency, F_2 the second harmonic, and F_3 the third harmonic. Adding the second harmonic and fundamental frequency algebraically (Fig. 12-244F) results in a new frequency indicated as the resultant. This new frequency is the same as the fundamental except the amplitude of the negative peak has been increased, while the amplitude of the positive peak has been reduced. Under these conditions the second harmonic will predominate, and is similar to a condition where the bias is too low.

The fundamental frequency and the second harmonic in a different phase relationship are shown in Fig. 12-244G. For this condition, the positive or resultant peak has a greater amplitude than that of Fig. 12-244F. It is of interest to note that for the condition in Fig. 12-244G the second harmonic is generated as if too great a grid bias voltage has been used, and is similar to the conditions in Fig. 12-244B.

In Figs. 12-244F and G a horizontal line has been added to indicate the shifted operating point. This line represents a dc component caused by self-rectification within the tube. Self-recti-

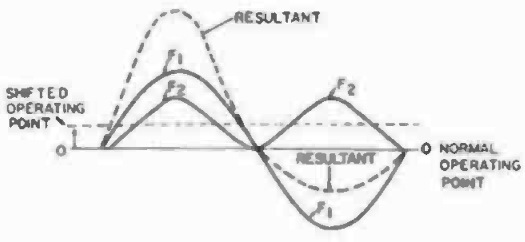


Fig. 12-244G. Resultant waveform when the operating point has been shifted due to self-rectification.

fication shifts the operating point on the plate curve of the tube, moving either upward or downward depending on the conditions to which the tube is being subjected at the moment.

The operating point has shifted upward due to the increase of the positive half-cycles and the decrease of the negative half-cycles in Fig. 12-244G, causing the introduction of second harmonics. When large values of second harmonics are present, there is an increase in plate current for an increase of signal voltage. In Fig. 12-244F the reverse has taken place.

When the distortion consists of a flattening of only one peak in the output waveform, the principal distortion is second harmonic. This is accompanied with a change in the average plate current and is called self-rectification.

Whenever both peaks are flattened, third harmonics predominate and if the distortion is considerable, a fourth harmonic will be generated, particularly if the control grid is driven positive.

When the control grid is driven positive, grid current will flow and a voltage drop is caused by the flow of current through the input load impedance, which may be a transformer or resistor in the grid circuit. This voltage drop flattens off the peaks of the input signal which is in turn reflected in the output waveform, along with the distortion caused by driving the tube into saturation.

When the voltage drop caused by the flow of grid current is 20 percent of the peak signal voltage, a second harmonic of 5 percent will be generated. As a rule when second harmonics are generated, third harmonic distortion is also generated at an amplitude approximately the same as second harmonic.

It is important in the design of class-B amplifiers (in which grid current flows an appreciable part of the input cycle), that the input load resistance be small and that the voltage drop caused by the load impedance in the grid circuit be small, as the voltage drop caused by the flow of grid current increases the bias voltage at the control grid, making it more negative than normal. This is why an input transformer designed to drive a class-B stage always has a low resistance secondary.

When a triode is overloaded, second-order harmonics are generated, while pentodes generate third-order harmonics. Even-order harmonics may be eliminated or reduced to a negligible value by the use of push-pull circuits. Odd-order harmonics generated by pentodes may be reduced by the use of push-pull circuits and the selection of the proper load impedance.

The addition of negative feedback will reduce the odd harmonics to a negligible value. As a rule a pentode is operated into a load impedance ⅕th to ⅛th the plate resistance of the tube. As the value of the load impedance is rather critical for a given harmonic distortion, the manufacturer's recommendations relative to load impedance and operating voltages should be closely adhered to.

12.245 What are the basic reasons for the difference in sound reproduction between transistor and vacuum-tube amplifiers?—The difference in sound reproduction of a transistor amplifier and a vacuum-tube amplifier of comparable power and frequency characteristics has long been a subject of contention between advocates of the two types of amplifiers.

As most transistor power amplifiers employ class-B operation in a push-pull power-output stage, the greater part of the distortion is generated within this stage. The disadvantage of using class-A output stages with transistors (single-ended or push-pull) is that collector current flows continuously. As a result, transistor dissipation is *highest* when no signal is present. Dissipation can be greatly reduced by the use of class-B push-pull operation. When two transistors are connected class-B push-pull, one transistor amplifies half of the signal and the other transistor amplifies the other half. These half-signals are combined in the output circuit to restore the original waveform.

Ideally, transistors used in class-B operation should be biased to collector cutoff so that no power is dissipated under zero-signal conditions. At low signal inputs, however, the resulting signal would be distorted, because of the low forward-current transfer ratio of the transistor at very low currents. This distortion is termed crossover or notch distortion, as discussed in Question 12.230.

In early designs of transistor amplifiers this type distortion was overlooked because the amplifiers were tested at their maximum power output, similar to vacuum-tube amplifiers; thus, crossover distortion was missed. Crossover distortion has a faculty of generating high-order odd harmonics, which are more annoying than second or third harmonics. Such reproduction leads to listening fatigue.

Another very important point is that transistors change their characteristics with temperature (contrary to the general belief); therefore, temperature-compensating circuits may have to be included and means taken to keep the amplifier at a reasonable operating temperature. Amplifiers are also affected by line voltage changes as discussed in Question 23.210. Several other factors also enter, such as the time-constant of the coupling circuits, and the bandwidth of the transistors which has a very definite effect on the feedback loop and can also induce distortion. Many other factors enter into the design, which are more or less in common with vacuum-tube amplifiers.

One of the features of a transistor amplifier is that it will deliver more power into a *reactive* load than appears, using a *resistive* termination, because it is a *current amplifier*. Therefore, the harder it is driven, the more current it delivers (assuming the load impedance is of low impedance). This is one of the reasons wby transistors in the output stage can be easily damaged.

Transistors adapt themselves easily to direct coupling, thus eliminating components and phase shift. By the use of opposite type transistors (pnp and npn), phase inverters are eliminated. Also, because of the low impedance of transistors, stray capacitance is reduced and therefore phase shift at the higher frequencies is reduced. These helpful factors permit the use of more negative feedback, increasing the stability and lowering the distortion and noise. To summarize, the difference in transistor sound has in the past been principally because of certain types of distortion being overlooked, and other design problems. Transistor amplifiers, although basically similar to vacuum-tube amplifiers, require special treatment in design concept and in testing procedures.

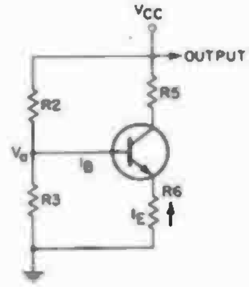


Fig. 12-246. Bias-voltage network for design problem.

12.246 Describe the procedure for designing a transistor bias network.— Assume a bias voltage network for a transistor is to be designed, using the network configuration in Fig. 12-246, for a low-level application operating at near-room temperature. The collector-to-emitter voltage V_{ce}, collector supply voltage V_{cc}, and emitter current I_e are selected by the designer, using values taken from the manufacturer's data sheet for small-signal parameters. The base current I_b equals $I_e/h_{fe} + I_{co}$; the voltage at point V_b then equals $(I_e \times R6) + V_{be}$ (dc). The base to emitter voltage V_{be} is approximately 0.2 volt for germanium and 0.7 volt for silicon. The voltage drop across R6 should be at least five times greater than V_{be}. The current through R3 is chosen to be at least five times I_b. Resistor R3 is then V_b/I_b, and R3 may be computed:

$$R3 = \frac{V_{cc} - V_b}{I_{R3} + I_b}$$

The load resistance R_L may be computed:

$$\frac{V_{cc} - I_E R6 - V_{CE}}{I_E} \text{ or } \frac{V_{cc} - V_{CE}}{I_E} - R6$$

Several other methods for developing bias voltage for transistors are discussed in Question 11.133. Voltage stabilization using thermistors is discussed in Question 11.157.

12.247 What effect does negative feedback have on transistor circuitry?— Negative feedback is discussed in some length in this section, as applied to vacuum-tube circuitry. Negative feedback may also be used with transistors for equalization and to reduce distortion and noise. The feedback voltage can be made proportional to the output current or voltage, and may be applied to either the input voltage or current. If the feedback voltage is proportional to the *output current*, the output impedauce will be *increased*. If the feedback voltage is proportional to the *output voltage*, the output impedance will be *decreased*. A negative-feedback signal applied to the input current decreases the input impedance, while a negative voltage applied to the input voltage increases the input impedance. If the feedback voltage is positive, opposite effects take place. As a rule, most of the information relative to negative feedback for vacuum tubes is applicable to transistor circuitry, when care is taken to observe the proper signal polarities.

12.248 Describe a solid-state plug-in microphone preamplifier.—The Model AT240 microphone solid-state preamplifier, manufactured by McCurdy Radio Industries, Ltd. (Canada) is typical of high-quality amplifiers designed for recording and broadcast use. The amplifier employs silicon transistors, and because of its low noise characteristics and wide input dynamic range, it will permit input levels of minus 60 dBm to minus 20 dBm to be used with very low distortion. The input impedance is quite high to prevent loading the microphone circuit, and its output impedance is low to provide maximum isolation against cross talk, when it is used in a mixer console or similar complex system. The mechanical assembly consists of glass-epoxy card design, using gold-plated printed circuitry. Because of its small physical size, 12 amplifiers may be plugged into a mounting tray of standard rack dimensions. Its outward appearance is shown in Fig. 12-248A.

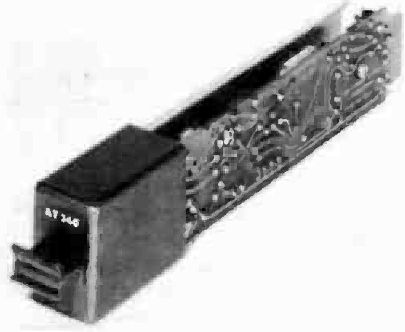


Fig. 12-248A. Model AT240 solid-state plug-in preamplifier manufactured by McCurdy Radio Industries (Canada).

Referring to the schematic diagram in Fig. 12-248B, the input is coupled through matching transformer T1 to the base of the common-emitter amplifier Q1. This transistor provides a large portion of the amplifier gain, while presenting a sufficiently high input impedance to avoid loading the source impedance. The output of Q1 is direct-coupled to Q2, which operates in

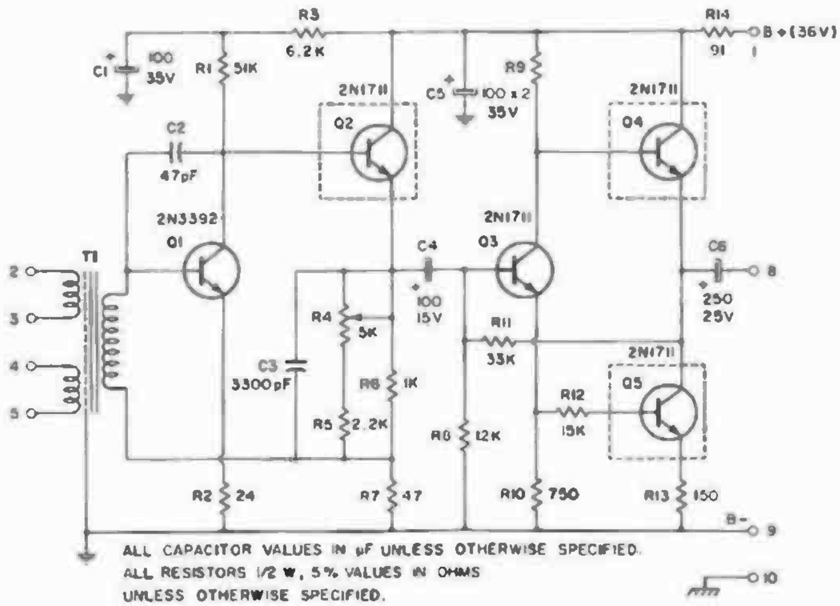


Fig. 12-248B. Schematic diagram for McCurdy Radio Industries Ltd. (Canada) Model AT240 plug-in solid-state microphone preamplifier.

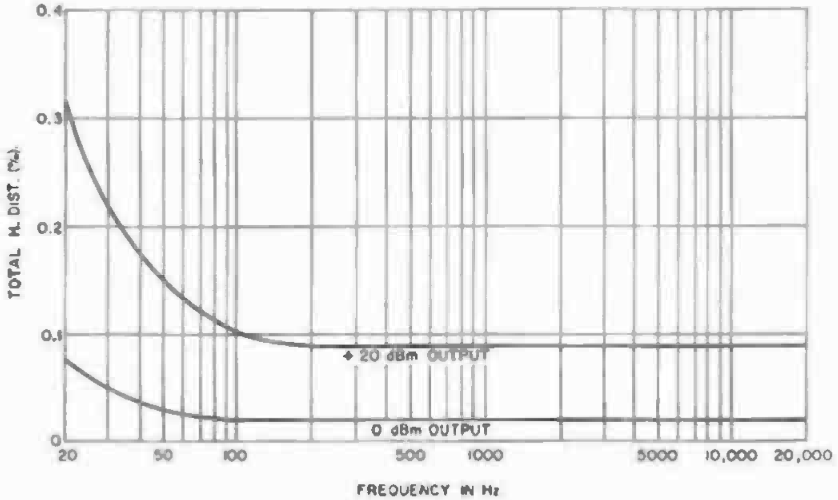


Fig. 12-248C. Distortion characteristics for McCurdy Radio Industries Ltd (Canada) AT240 solid-state microphone preamplifier.

an emitter-follower configuration and feeds the output amplifier. The gain-control circuit comprising potentiometer R4 provides a limited gain adjustment of approximately plus or minus 1 dB, by varying the negative feedback to the base of Q1 via T1. The signal developed at Q2 emitter is coupled through C4 to the base of Q3. Transistor Q3 functions as a phase splitter, since signals of opposite polarities appear at its collector and emitter. The outputs of Q3 are direct-coupled to the base of Q4 and Q5 respectively. Transistors Q4 and Q5 form a single-ended push-pull output stage, with the signal at the emitter of Q4 in phase with that appearing at the collector of Q5. These signals are added in the common output circuit and the resultant signal is coupled to the load resistance through C6. Transistors Q3, Q4, and Q5 operate class-A, with the dc operating point set by resistors R11 and R8. Resistor R11 also provides negative feedback from the output to the base of Q3 to reduce distortion and stabilize the gain. Additional feedback is provided by the unbypassed emitter resistors R2, R7, R10, and R13. Transistors requiring heat sinks are shown in dotted lines.

Specifications are: Gain is 40 dB plus-minus 1 dB; with frequency response from 20 to 20,000 Hz, plus-minus 0.5 dB. Input noise unweighted 10 to 100,000 Hz, minus 124 dBm, or minus 127 dBm, 10 to 20,000 Hz (bandwidth specified at the minus 3-dB point). Maximum output is plus 20 dBm. Source impedance is 37.5 and 150 ohms. Input impedance is greater than ten times the source impedance. Internal output impedance is 25 ohms. The distortion characteristics are given in Fig. 12-248C.

12.249 Describe a solid-state mixer (booster) plug-in amplifier.—In Fig. 12-249A is shown the schematic diagram for the Model AT241 McCurdy

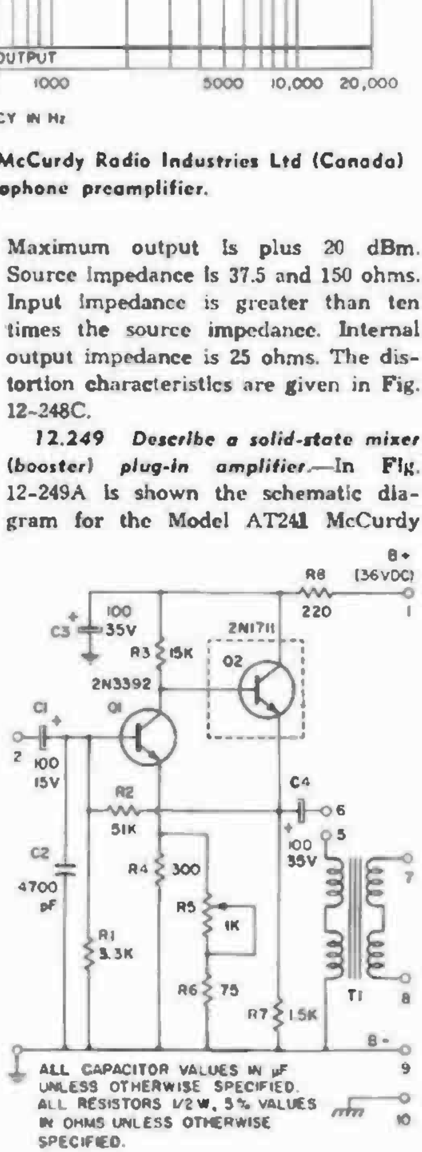


Fig. 12-249A. Schematic diagram for McCurdy Radio Industries Ltd. (Canada) Model AT241 plug-in solid-state mixer amplifier.

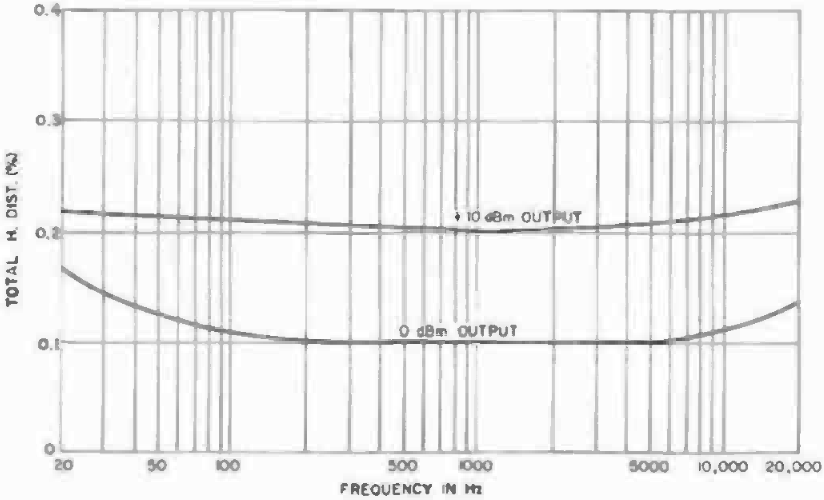


Fig. 12-249B. Distortion characteristics for McCurdy Radio Industries Ltd (Canada) AT241 solid-state mixer or booster amplifier.

Industries Ltd. (Canada) silicon transistor mixer or booster amplifier. Normally a mixer or booster amplifier is used after the mixer network to compensate for network-insertion loss. The amplifier is similar in mechanical construction to the AT240 amplifier, described in Question 12-248.

Referring to the schematic diagram, transistors Q1 and Q2 form a direct-coupled pair, with Q1 operating as a common-emitter amplifier, and Q2 operating as an emitter-follower. The input signal is coupled through C1 to the base of Q1, where it is amplified and direct-coupled to the base of Q2. Transistor Q2 provides the necessary power gain and its low output impedance provides optimum isolation when working into a 600-ohm load.

Resistors R1 and R2 supply voltage feedback to the base of Q1 to set the impedance of this stage at 600 ohms. Unbypassed emitter resistors R4 and R6 at Q1 supply negative feedback to reduce distortion and stabilize the gain, with R5 providing for a small gain adjustment. Transformer T1 is normally connected in the output; however, its position may be reversed and used as an input transformer by re-strapping.

Specifications are: Gain 35 dB, plus-minus 1 dB. Frequency response 20 to 20,000 Hz, plus-minus 0.25 dB. Input noise unweighted, minus 120 dBm to 100,000 Hz (bandwidth specified at minus 3-dB points). Harmonic distortion is less than 0.2 percent at zero dBm, output 20 to 20,000 Hz. Maximum output is plus 10 dBm. Source impedance is 600 ohms. Internal input impedance is plus-minus 10 percent of source impedance. Internal output impedance is 50 ohms; with transformer connected for 600-ohm load, it is 150 ohms. Load impedance is 600 ohms. Distortion characteristics appear in Fig. 12-249B.

12.250 Describe a transistor monitor amplifier operating class-A.—Many times a small monitor amplifier is required, having an output between 1 and 10 watts. A monitor amplifier operating within this range, using transistors and operating class-A, is shown in Fig. 12-250A. This amplifier manufactured by Langevin, may be used as either a program or monitor amplifier. Provision is made for reducing the power output to 1 watt, by removing a single strap connection. The Model AM17, to be described, is unusual in its design as most transistor amplifiers are operated as class-AB or class-B.

Starting at the input, the secondary of the input transformer is split and a balancing control R10 is inserted for adjusting the balance of the output stage, which is measured by connecting a voltmeter between the emitters of transistors Q9 and Q10. Transistors Q1 to Q6 are voltage amplifiers, with the outputs of Q5 and Q6 coupled to Q7 and Q8, through diodes D1 and D2. The purpose of these diodes is to prevent the output stage from drawing excessive current. With the output stage balanced, there is no appreciable current change in Q9 and Q10 with a change of signal level, which is normal for a proper class-A amplifier stage. (See Question 12.226.) Choke L1 and capacitor C7 comprise an interference filter which attenuates any rf interference that may be picked up on the incoming dc power-supply connections. It will also remove video vertical-sync pulses when used around video equipment.

With the amplifier strapped for 1-watt operation, plus 30 dBm, the current demand is about 1 ampere, and for 8 watts (plus 39 dBm), it is 2 amperes. The lower power connection is used when it is operated as a program amplifier. All transistors are silicon. The fre-

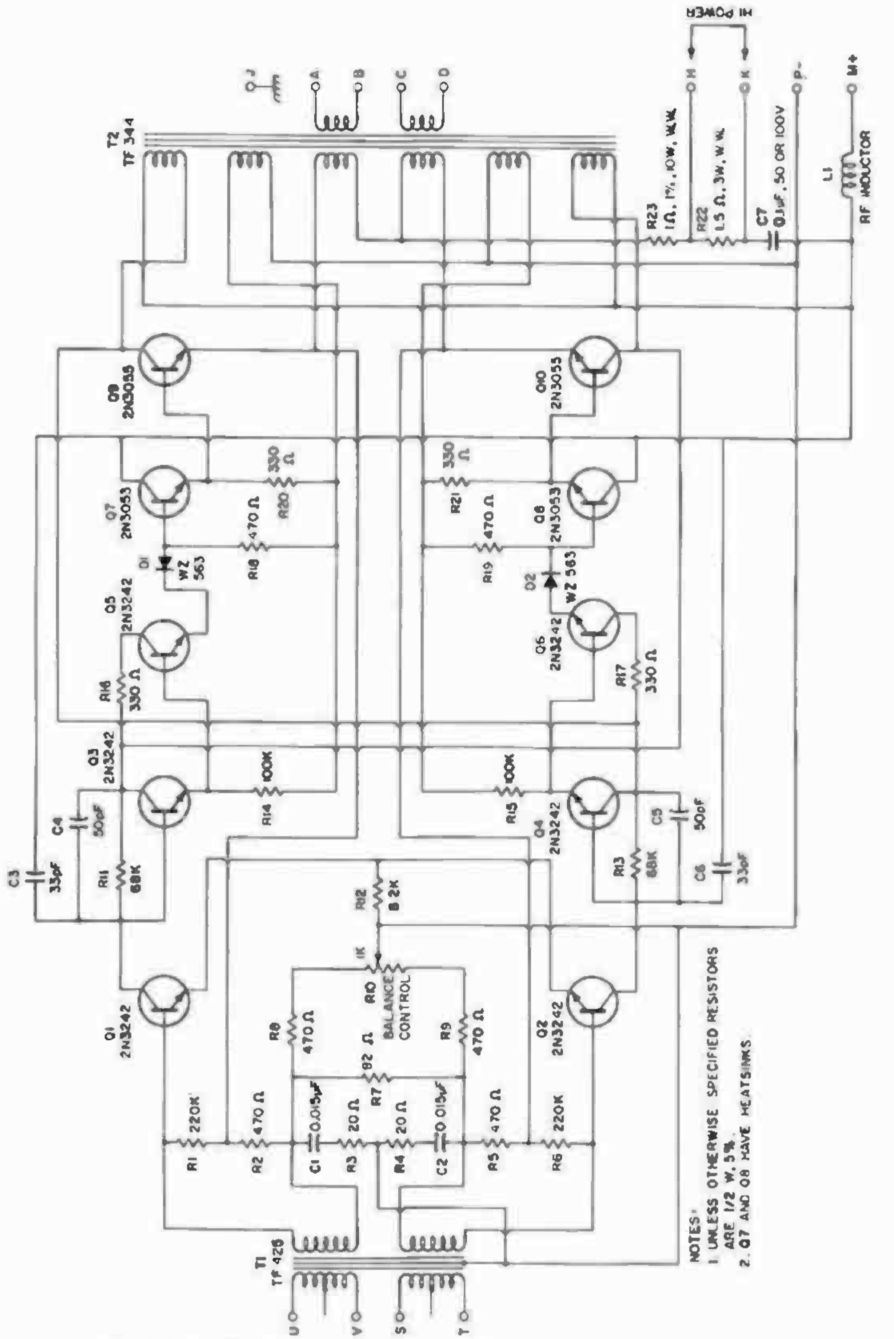


Fig. 12-250A. Langevin Model AM17 transistor power amplifier.

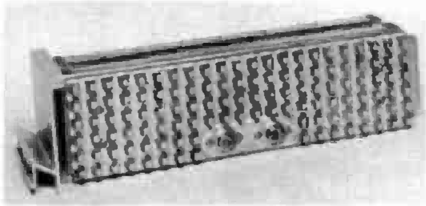


Fig. 12-250B. Langevin Model AM17 transistor class-A power amplifier.

quency response is plus-minus 0.5 dB, 20 to 20,000 Hz, with a gain of 57 dB. Harmonic distortion is 1 percent at plus 24-dBm output. Equivalent noise is minus 115 dBm. The appearance of the complete device is pictured in Fig. 12-250B. A separate power supply is required. Four of the units shown may be plugged into a supporting tray.

12.251 ***Describe the basic principles of a transistor complementary-symmetry amplifier operating class-B.***—Due to the nature of the pnp and npn transistors, they may be connected to have all the benefits of the conventional push-pull amplifier circuit, without the need of a phase-inverter stage, driving transformer, or output transformer. The basic circuit for this type amplifier is shown in Fig. 12-251A. Input-coupling capacitor C charges through one transistor during the positive half-cycle of the input signal, and discharges through the other transistor during the negative half-cycle of the input signal. This action eliminates the need for discharge diodes with capacitance coupling, as required in conventional transistor class-B push-pull amplifiers.

The circuit of Fig. 12-251A shows two transistors connected in a complementary-symmetry circuit. Transistor Q1 is an npn type and Q2 is a pnp type. If a negative-going input signal is ap-

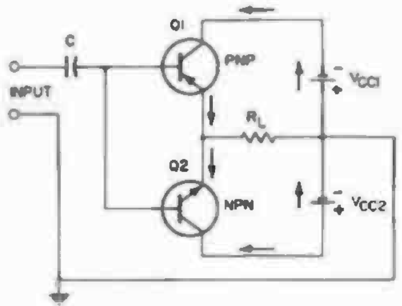


Fig. 12-251A. Basic circuit for a complementary-symmetry transistor amplifier operating with zero bias.

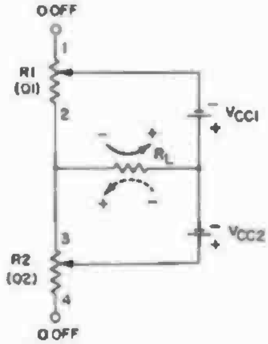


Fig. 12-251B. Simplified circuit of a transistor complementary-symmetry amplifier.

plied to capacitor C, transistor Q1 is forward biased and it conducts. With a positive-going signal, Q2 is caused to conduct. As one transistor is conducting, the other is nonconducting. The resulting action in the output circuit may be better understood by referring to Fig. 12-251B. The external emitter-collector circuit of transistor Q1 is represented by variable resistor R1 and that of Q2 by resistor R2.

With no input signal and the amplifier biased for class-B operation (zero emitter-base bias), the variable arms of the resistors can be considered to be in the off position. As the arm of R2 moves toward point 3, current passes through the series circuit consisting of battery V_{CC2}, resistors R_L and R2. The amount of current flow depends on the magnitude of the incoming signal, the variable arm moving towards point 3 for increasing forward bias, and toward point 4 for decreasing the bias. Current flowing in the direction of the arrows produces voltages with the indicated polarities.

When the input signal goes negative Q1 couducts and Q2 becomes nonconductive. The same action is repeated as for resistor R1. Current flows through V_{CC1}, resistor R1, and load resistor R_L in the direction shown by the solid line arrow and produces a voltage across R_L with the indicated polarities.

12.252 ***Describe a complementary-symmetry transistor amplifier operating class-A.***—To operate a complementary-symmetry amplifier class-A, a forward bias voltage is applied to both transistors, so the collector current is not cut off at any time. Referring to Fig. 12-251B, under class-A operation the variable resistors will not be in the off position at any time. The dc bias current in the output flows out of the negative battery terminal V_{cc2}, into the positive terminal of battery V_{cc1} through resistors R1 and R2, and into the positive terminal of battery V_{cc2}. No current flows through the load resistor R_L. Under these conditions the output circuit may be considered to be that of a balanced bridge, consisting of resistors R1, R2 and batteries V_{cc1} and V_{cc2}. With a positive-going input signal, transistor Q2 conducts more, and Q1 conducts less.

In the simplified circuit of Fig. 12-251B the arm of resistor R1 moves toward point 1, and that of resistor R2 moves toward point 3. This results in an unbalanced bridge condition and electrons flow through resistor R_L, in the direction of the dashed-line arrows, producing the indicated polarities. With a negative-going input signal, transistor Q1 conducts more and Q2 conducts less. The variable arm of R1 moves to point 2, and that of R2 moves toward point 4. Again the bridge is unbalanced, and electrons flow through R_L in the direction of the solid-line arrows, producing voltage with the indicated polarities. For either class-A or -B operation there is no direct current flow through load resistor R_L; therefore, a loudspeaker voice-coil winding may be directly connected in place of load resistor R_L.

12.253 Describe a direct-coupled complementary-symmetry amplifier.—A common-emitter complementary symmetry circuit is shown in Fig. 12-253A. In this circuit, transistors Q3 and Q4 are driven directly by another common-emitter complementary-symmetry stage, Q1 and Q2. The signal for Q1 and Q2 is taken from a single-ended stage. When the input goes positive, Q1 conducts and Q2 is nonconductive because of the 180-degree phase reversal in the common-emitter configuration, the collector of Q1 goes negative which causes Q3 to conduct, and Q3 collector goes positive. With a negative input signal, Q2 conducts and its collector goes positive, which causes Q4 to conduct and its collector to go negative. Transistors Q1 and Q3 are nonconductive during this period. Battery V_{EE1} supplies the required biasing for Q1 and Q3, and battery V_{EE2} biases Q2 and Q4. The emitter-base junction of Q3 is in series with the collector-emitter circuit of Q1 and battery V_{EE1}. As a result the emitter of Q3 is positive with respect to its base (forward bias) and the collector of Q1 is positive with respect to its emitter as is required for electron flow. A similar arrangement exists for Q2, Q4, and battery V_{EE2}.

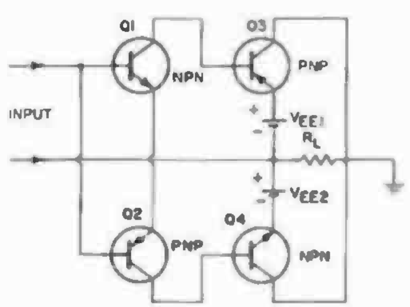


Fig. 12-253A. Common-emitter direct-coupled complementary-symmetry amplifier.

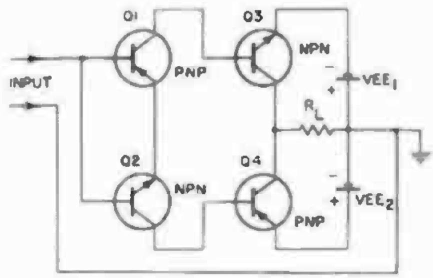


Fig. 12-253B. Common-collector direct-coupled complementary-symmetry amplifier.

A common-collector configuration is shown in Fig. 12-253B. As for a single transistor, maximum voltage, current, and power gain are obtained with the common-emitter connection; however, its input resistance is low. The common-collector configuration will provide higher input resistance with less current, voltage, and power gain.

12.254 What is a compound-connected transistor amplifier?—By the use of a compound configuration, the drop-off of collector current or the reduction in audio feedback at higher emitter currents can be made negligible. Fig. 12-254A shows a variation of collector currents versus emitter current for a single transistor, and two transistors in a compound configuration. Compound-connected transistors can be single-ended, push-pull, or in a complementary-symmetry configuration; the latter circuit appears in Fig. 12-254B. This

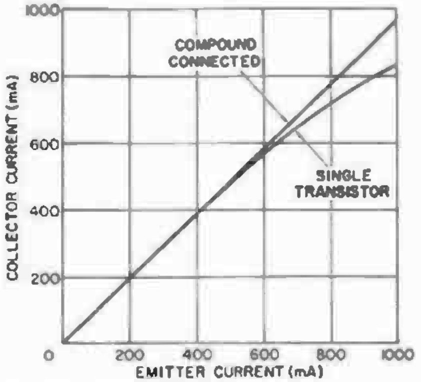


Fig. 12-254A. Variation of collector current with emitter current for a single transistor and for compound-connected transistors.

circuit is somewhat similar to that of Fig. 12-253A. Transistor Q1 is replaced by the compound connection of Q1A and Q1B. The gain for such a configuration is greater than that of the complementary-symmetry configuration. Such circuits are used in power-amplifier output stages.

12.255 ***Describe an 80-watt transistor power amplifier, using a half-bridge output stage.***—The basic design for a half-bridge push-pull power amplifier given in (a) of Fig. 12-255 is capable of developing 80 watts of audio power, with less than 2.0-percent total harmonic distortion (THD) up to full output. The general form is that of a two-stage transformer-coupled half-bridge push-pull output circuit, direct-coupled to an 8-ohm load. This arrangement allows considerable design latitude in achieving high sensitivity and high values of negative feedback, with the least number of active components, as compared to the symmetry approach. In addition, the circuit is thermally stable relative to the bias for the output transistors. The use of a driver transformer presents no particular problems, as the design is quite simple.

The output circuit is of standard half-bridge configuration, which drives the load directly from a balanced plus-minus power supply. Because of the balanced power supply, there is only a negligible dc component across the load, caused by bias-voltage tolerances. This dc component and the ac ripple accompanying it are easily nulled out by small bias pots, if desired. The bias network itself is a stable voltage divider which presents a fixed voltage to the base of each power transistor to allow an idling current determined by the inherent transconductance characteristics of the device. The 0.47-ohm emitter resistor provides a small amount of degeneration, enhances the stiff bias voltage to present a thermally stable circuit, and offers a small amount of local ac feedback. The output circuit employs the driver-transformer secondary dc resistance as a part of the bias network, since it can be held to tighter tolerance than a separate resistor. This arrangement minimizes the total drive voltage required from the transformer secondary so that the emitter diode of the power transistor is avalanched only a minute amount.

Shunt-bias networks are only practicable when the output transistors are driven from the low-impedance source (as reflected through the transformer) to minimize the effects of the shunt negative-voltage feedback loop, which degrades the input impedance and increases the required driving power. Extensive tests have indicated this type voltage drive optimizes the basic linearity throughout the whole bandwidth and extends the frequency range of the output circuit. The low source impedance provides a constant voltage to the output transistors, as their input impedance changes with load and frequency. This permits a low distortion gain through the exceptionally linear transconductance characteristic.

The driver section consists of two common-emitter stages, direct-coupled and feedback-biased for maximum sta-

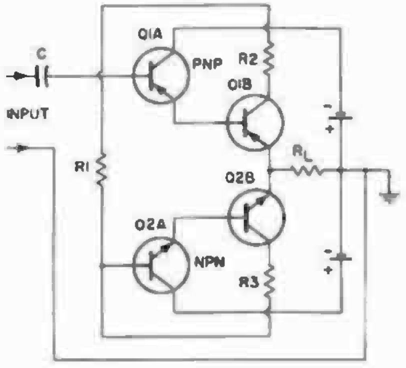
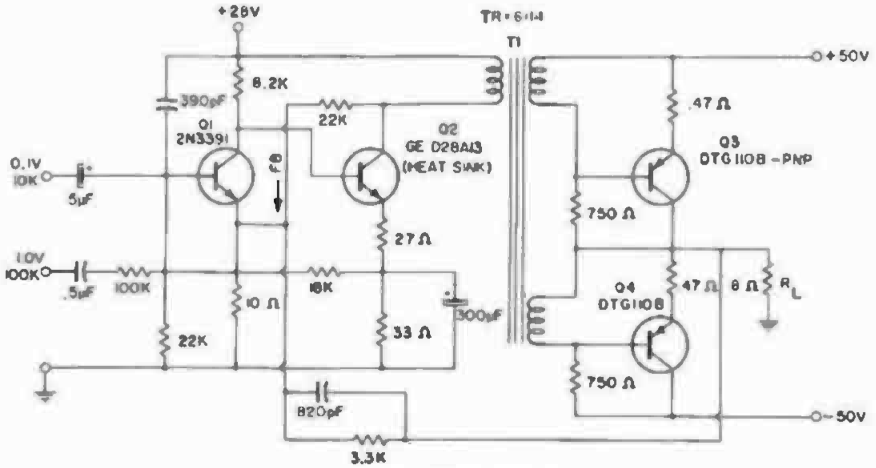
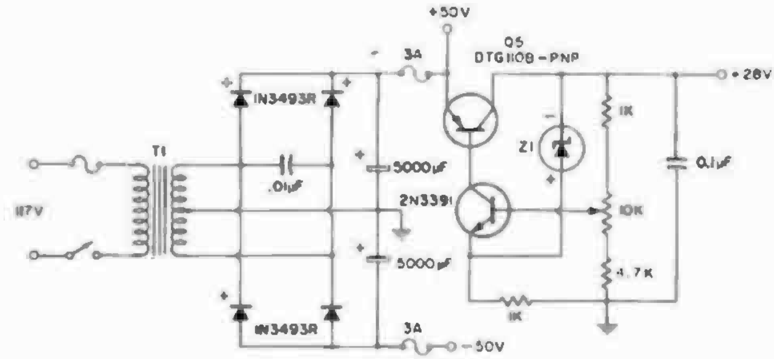


Fig. 12-254B. Compound-connected transistors in a complementary-symmetry configuration.



(a) *Amplifier circuit.*



(b) *Power-supply circuit.*

Fig. 12-255. Schematic circuits for 100-watt half-bridge transistor power amplifier. (Courtesy, Delco Radio Division.)

bility. The dc feedback bias is derived from the emitter of Q2 and applied back to the input of Q1. This provides good thermal stability and compensates for dc parameter variations between the two transistors. About 16 dB of ac negative feedback is derived from the collector of Q2 to reduce the output impedance of the driver stage, thereby providing a low-impedance (voltage) source to drive the output stage.

The driver transformer is of split-primary design, using grain-oriented steel core material. The split-primary is bifilar wound and permits a high degree of coefficiency between the primary and secondary, resulting in a flat frequency response throughout the audio spectrum, and rolling-off in a frequency range most beneficial to the total gain-phase behavior of the composite amplifier. This favors maximum driver efficiency. (See Question 8.109.)

A 6:1 turns ratio was chosen, for the driver transformer (Goslin Corp. TAD 3257) so that a large drive voltage swing from a low current signal device would drive the DTG 110A's (Delco) output transistors to full output. The GED28A13 transistor is quite satisfactory for this purpose when mounted on a 1" × 1" × 1/16" aluminum plate insulated from the chassis. The circuit shown provides exceptional sensitivity for the rated output of 80 watts, with 34 dB of total negative feedback. This is achieved by staggering the frequency-response characteristics of the silicon planar transistors Q1 and Q2, the low impedance of the driver transformer, and the DTG 110A output transistors.

The power supply is capacitor filtered, which provides sufficient regulation for full continuous output of two 80-watt amplifiers. A simple feedback regulator Q5 and Q6 provides the driver bias supply with very low ripple content for optimum noise and power-supply feedback rejection.

The amplifier will deliver up to 80 watts to an 8-ohm load from an input signal of 0.10 volt from an input impedance of 10,000 ohms, or driven from a 1-volt source with a 100,000-ohm input impedance. Bandwidth is plus or minus 1 dB, 20 to 20,000 Hz at 1 to 20 watts output. Total harmonic distortion is less than 2 percent at 1 kHz up to full output. Intermodulation distortion is less than 2 percent up to full output, using 60 and 6000 Hz in a 4:1 ratio. Signal-to-noise ratio (thermal) is 80 dB below rated output. Internal output impedance is 0.5 ohm and rise time is 10 microseconds to full output.

The output transistors must be mounted on heat sinks with at least 165 square inches of surface, with mica washers using silicon oil, or mounted directly on the heat sinks, then insulating the heat sink from the chassis for greater dissipation of heat. When making square-wave tests, the waveform at the output can be improved (if necessary) by trimming the 390- and 820-pF capacitors. The output may be reduced to 50 watts by reducing the operating voltage to 35 volts, rather than the 50 volts called for. This will require changing the values of the 750-ohm resistors in the output stage to 550 ohms each.

12.256 Describe a 200-watt stereophonic transistor amplifier, using the Sharma circuit.—The attainment of high power output in transistor amplifiers is rather difficult and has been the subject of much research. An interesting circuit has been developed by Madan M. Sharma, of Mattes Electronics Inc., using a double class-B or totem-pole output-stage configuration (Fig. 12-256A). It has been found that the presence of bias voltage in the output stage does not entirely eliminate notch distortion common to class-B amplifiers, and imposes limitations on the driver stage. Therefore, the power-output stage of the Sharma circuit does not use bias voltage. Referring to Fig. 12-256B, the basic circuit, the output stage employs simulated negative-impedance characteristics driven by a low-power, low-distortion driver amplifier, which is basic in design.

The low output of the driver stage stabilizes the power stage and provides the drive signal. Near the zero crossing, when the power transistors are below the threshold of conduction, the output load is driven entirely by the driver stage. When the signal rises, the parallel combination of the load impedance and the negative output impedance of the power stage appears across the output of the driver as a very high impedance. The driver stage drives the load when the signal is low, but drives a high impedance when the signal level is high. In either condition, the driver stage is required to supply only a small amount of power.

To achieve maximum power output, this circuit was modified (Fig. 12-256C) to use two latching diodes, D1 and D2, and resistors R1 and R2. The combination is called a latching circuit and disconnects the power stage from the output of the driver stage when the instantaneous voltage across the output terminals is greater in magnitude than the driver power-supply voltage. This action isolates the two amplifiers when isolation is required, yet permits the system to operate as a basic circuit during the balance of the time. At low levels, the drive is connected to the output terminals through latching diodes D1 and D2, permitting dc feedback to stabilize the dc output at the output terminals. High-level signals are transformer-coupled, offering the best impedance match between the driver and the output transistors. Using separate power supplies, the power stage may be operated at any voltage.

The driver stage also supplies power. The proportion of power delivered by the two amplifier stages depends on the signal level. The two amplifiers continue to drive the load collectively until a signal level is reached when the power amplifier supplies the power. At this point, the driver stage supplies no power but acts as a control signal. At the higher levels, when the power transistors are supplying all the output power, they tend to drive themselves through the driver transformer. Since the driver stage presents a low output impedance (lower than the impedance of the power stage), the excess current which would have caused oscillation returns to ground through the driver stage. When the output is larger than

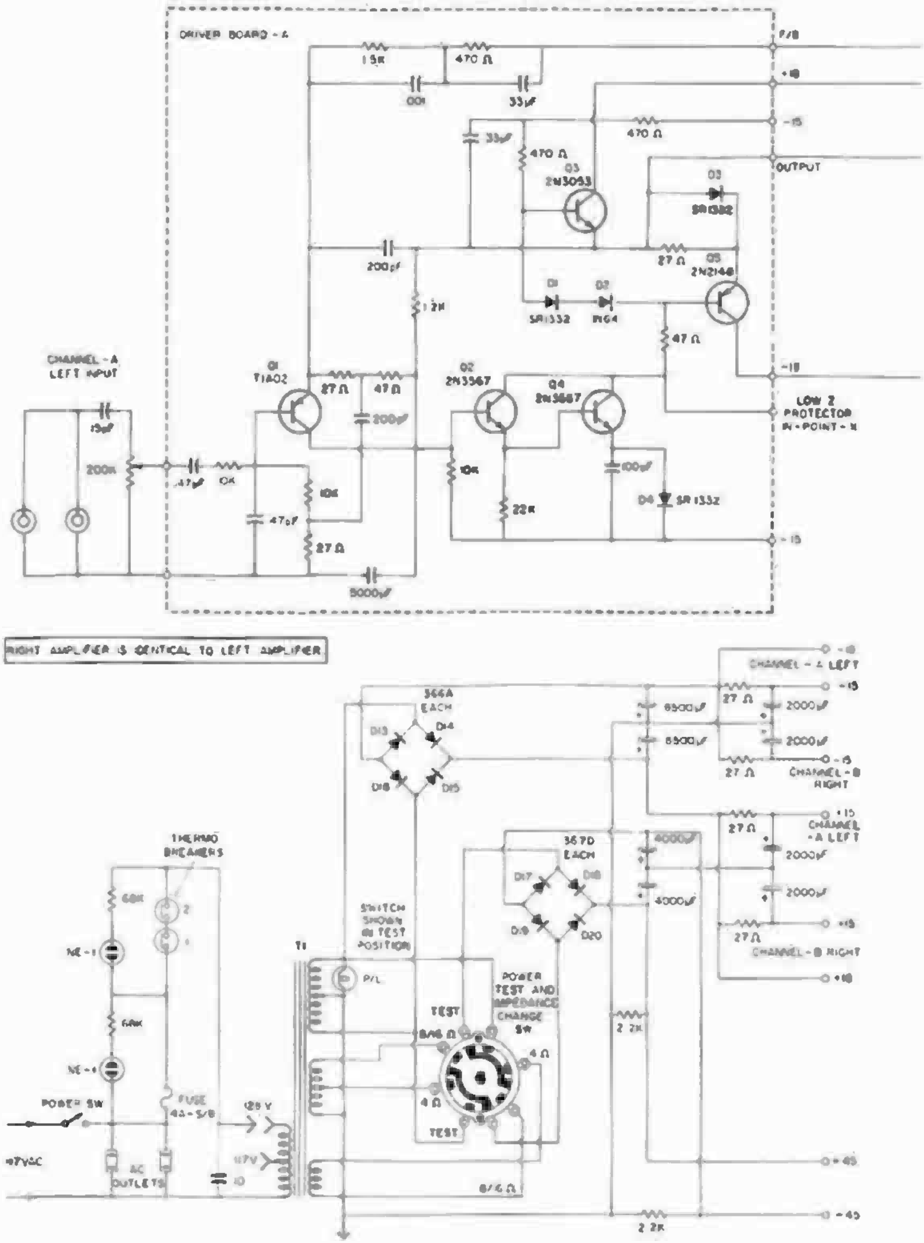
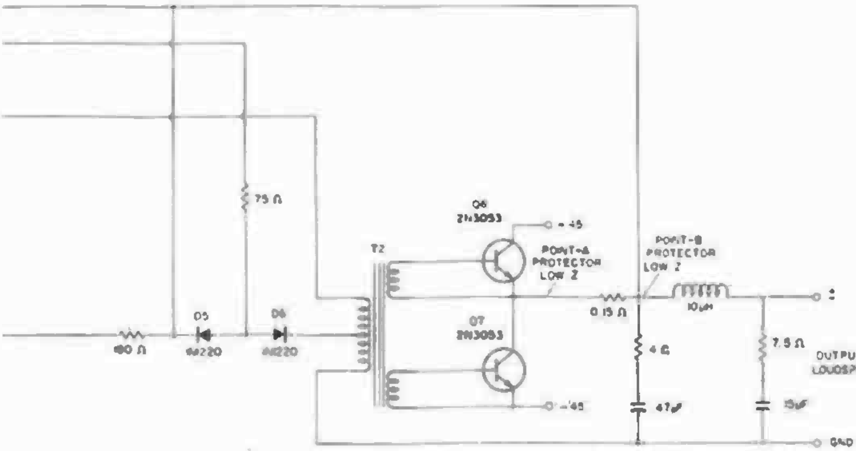
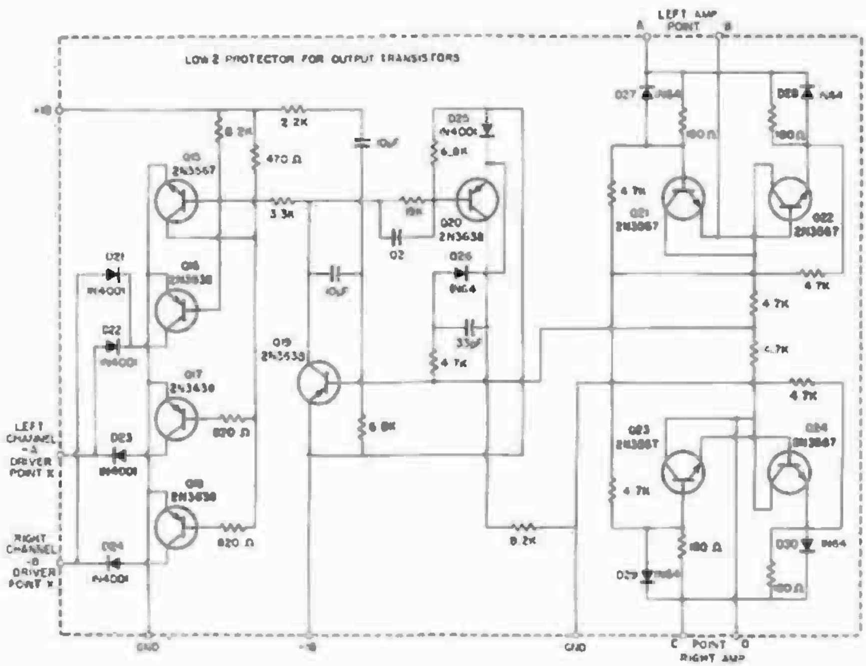


Fig. 12-256A. Schematic diagram for Mattes Electronics, Inc.

the supply voltage of the driver, the latching diode stops conduction. This isolates the two amplifiers. Now the signal flow is maintained through the driver transformer only, and the amplifier operates like any transformer-coupled amplifier. Because of the negative feedback, transients between the different modes are smoothed out and linearity established throughout.

The complete circuit with its voltage amplifier stages is shown in the overall schematic diagram in Fig. 12-256A. It will be observed a driver amplifier limiter has also been included. The two amplifiers are each capable of developing 100 watts of continuous power into a load of either 4 or 8 ohms. For a load of 16 ohms, the maximum power output drops to 60 watts per channel. The intermodulation distortion, working into 4 or 8 ohms is 0.07 percent, using the





200-watt stereophonic transistor amplifier using the Sharma circuit.

SMPTE test method as set forth in Section 23. The harmonic distortion is 0.5 percent maximum at 100 watts. Frequency response is plus-minus 0.5 dB, 15 to 30,000 Hz. Damping factor is 250, with an effective internal output impedance of 0.04 ohm. Hum and noise are 90 dB below the rated output, or 1 millivolt. One of the important features of the Sharma circuit is that the output terminals may be operated with a direct short circuit without damaging the output transistors. Also, because of the design capacitive loads of 0.5 microfarad may be driven; however, maximum power output cannot be achieved.

To summarize, at very low signal levels, the power transistors are below the threshold of conduction, the driving power being supplied by the driver stage. As the signal level is increased, the power transistors begin to conduct and supply power to the external load circuit.

The output stage has an efficiency of 78 percent, the theoretical maximum for a class-B amplifier. The driver transformer has a turns ratio of 3:1, using *pentafilar* windings of low dc resistance. The output transistors are rated 15 amperes collector current. If a lower rating is used, it is possible for them to be damaged, if the output terminals should become short-circuited. Balanced component values are not necessary, standard 10-percent resistors are quite satisfactory.

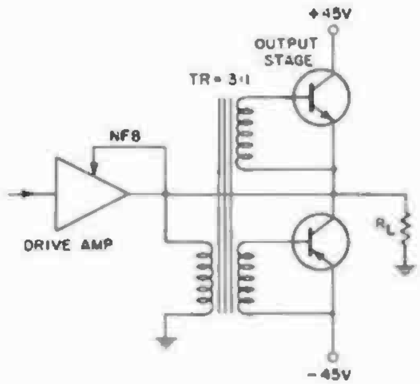


Fig. 12-256B. Basic circuit of Sharma circuit used in the Mattes Electronics, Inc. stereophonic solid-state amplifier.

Because of the 100-watt output capabilities of the amplifier, a test switch has been incorporated to decrease the power output to about 12 watts when setting up for testing. The 10-microhenry chokes and the capacitors in the output circuit constitute a low-pass filter which removes very high frequency components from the signal, that are not entirely eliminated by the feedback circuit due to transient-time effects present in all power transistors. The cutoff frequency is high enough to not affect the response of the amplifier as a whole. Suggested specifications for the driver transformer are given in Question 8.109.

Connected across the output of each power transistor stage is a protective circuit consistiug of 5 transistors and 5 diodes. It is a fact, but not generalized, that the internal input impedance of a loudspeaker system at the very low frequencies may present a *dead short* circuit to the output power transistors. If they are not adequately protected instaut failure is the result.

With the protective circuitry shown at the lower right of Fig. 12-256A, the output transistors of each channel are protected against changes in phase angle (caused by the speaker characteristics), and against a short circuit across the output. If the built-in limits set by the manufacturer are exceeded even by a partial short circuit, the amplifier will cease operating instantly, then sample the load impedance every second for a few milliseconds duration to determine if the amplifier load impedance has returned to normal.

It will be noted that the actuating signal is taken from across a 0.15-ohm resistor connected in series with the output line, at points A and B in the left channel and C and D in the right channel. The outputs of the protective circuits are returned to the driver stage at points X and Y. Thus for a malfunction in the output circuit, the driver circuits instantly become inoperative,

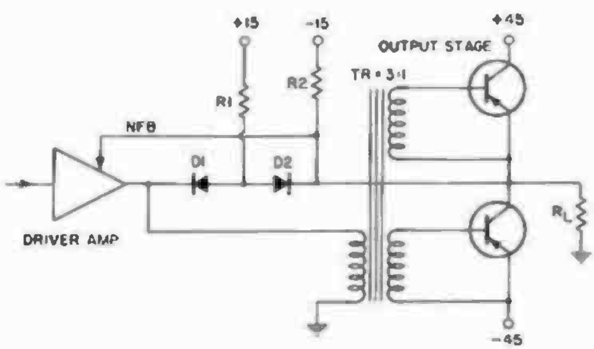


Fig. 12-256C. Modification to basic circuit of Fig. 12-256B used in the Mattes Electronics, Inc. stereophonic solid-state amplifier.

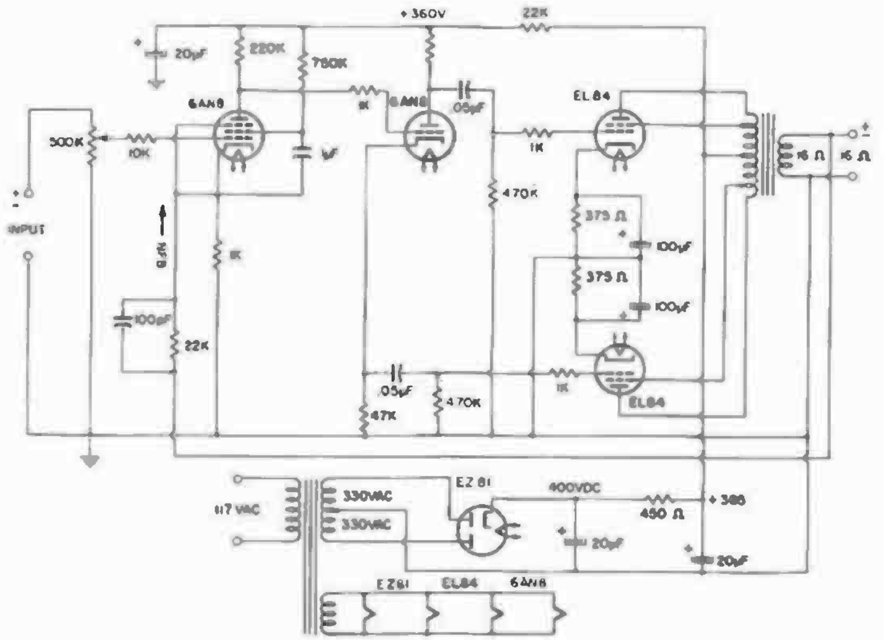


Fig. 12-257A. 15-watt Williamson ultralinear amplifier used in FET conversion.

and damage to the output transistors is prevented. The operation of the sensing control circuit produces a characteristic sound in the speaker system, warning the listener of difficulties in the output circuit. When trouble has been cleared, the amplifier returns to normal operation. (See Question 20.138.)

12.257 Can a field-effect transistor (FET) be directly substituted for vacuum tubes in an amplifier?—Yes, under certain conditions and if they are used only for *voltage-amplifier stages.* Considerable work in this direction has been done by Rheinfelder. All tubes, except the power output stage were replaced, and only a slight modification to the power supply circuitry was made to reduce the voltage. The output stage of a typical 15-watt Williamson ultralinear amplifier is shown in Fig. 12-257A, with its conversion in the voltage stages to FET's shown in Fig. 12-257B. After conversion, measure-

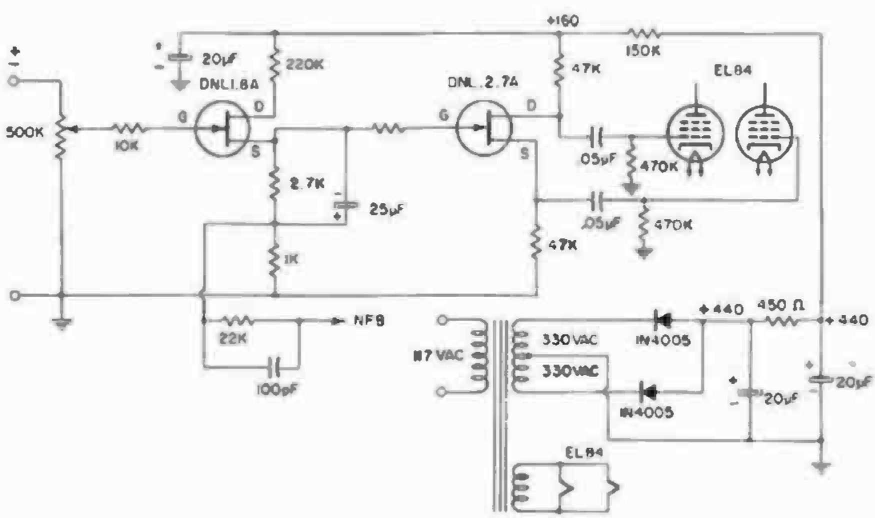


Fig. 12-257B. 15-watt Williamson ultralinear amplifier voltage stage and phase-splitter after conversion to FET's.

ments indicated the same overall gain, improved frequency response, lower hum and noise for over a 10-percent increase in power output for 1-percent distortion. The rectifier tube in the power supply was replaced with two diode rectifiers, and the voltage was dropped by the use of a 150,000-ohm resistor in the high-voltage lead to the voltage amplifier stages, rather than the 22,000-ohm resistor in the original circuit. The FET's were soldered directly to the tube sockets. The original values of the components in the negative-feedback loop were retained. (See Questions 12.128 and 12.232.)

12.258 Describe a source follower, using a field-effect transistor (FET).— A source follower is quite similar in nature to the cathode-follower circuit used with vacuum tubes. In the circuit of Fig. 12-258A, the output voltage is taken from the source element and follows the input (gate) due to the action of the negative series-voltage feedback, because the output voltage is applied in series with the input signal voltage. Therefore, the gain is:

$$A' = \frac{A}{(1 + A)}$$

where,

A is the gain before feedback,
A' is the gain with feedback.

The input impedance (R'_{in}) with feedback is derived:

$$R'_{in} = \frac{R_{in}}{1 - A'}$$

where,

R_{in} is the input impedance without feedback,
A' is the voltage gain of the source follower.

The output impedance may be approximated:

$$R_o = \frac{1}{G_m}$$

where,

G_m equals the effective circuit transconductance.

A typical example: A gain of (A) equal to 70 results in a net gain of 0.985, with feedback, and an impedance multiplying factor of 666 (666 megohms with a gate impedance of 10 megohms). Under these conditions, the output impedance is 1000 ohms, depending on the type FET employed. The output capabilities for a source follower are identi-

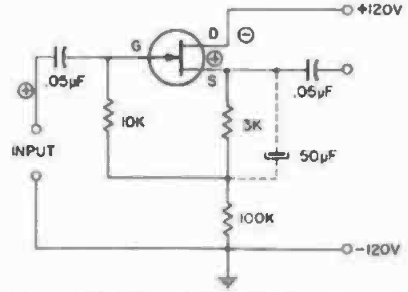


Fig. 12-258A. Field-effect transistor (FET) source follower. Signal polarities are given to indicate the phase reversals between input and output signals.

cal to a conventional source amplifier stage; however, the distortion below overload is considerably better because of the negative feedback. The source follower finds its greatest use where an extremely high input impedance is required, with a low output impedance. It should be operated into a load impedance of 100,000 ohms minimum, and from a source impedance higher than the input resistance, before the feedback was applied (10 megohms).

Optimum performance is obtained with the source element set to 65 percent of the voltage applied to the drain, which may be obtained for the circuit shown, using the specified FET, with a 3000-ohm resistor in the source circuit. This resistor may be left bypassed as little effect is noted with it unbypassed. The frequency response of the follower circuit shown is down 3 dB at 290 kHz, using a source resistance of 100,000 ohms, and with a total capacitance of 5.3 pF at the input, including the socket.

The optimum gate voltage for a source follower is close to one-half the supply voltage, the same as for the conventional source amplifier. A direct-

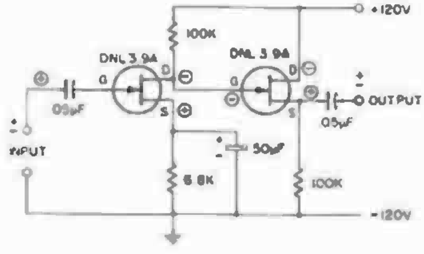


Fig. 12-258B. Direct-coupled field-effect transistors (FET) signal polarities are given to show the phase reversals between the input and output signals.

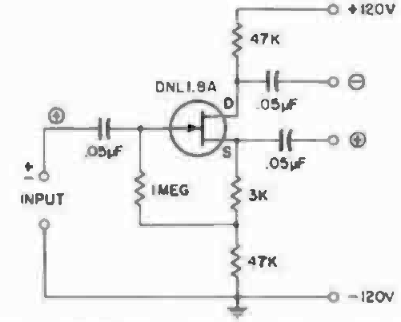


Fig. 12-259A. Phase-splitter circuit using a single FET (after Rheinfelder).

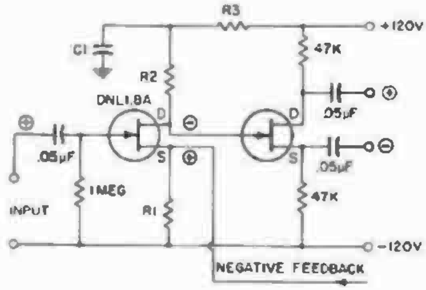


Fig. 12-259B. Direct-coupled phase splitter using two FET's (after Rheinfelder).

coupled source follower is shown in Fig. 12-258B. The characteristics for several different values and source resistance are given below.

Drain resistance	*Source resistance*	*Gain*	*3-dB bandwidth*
100K	6.8K	35.6 dB	30 kHz
220K	15K	39.0 dB	23 kHz
470K	39K	41.2 dB	15 kHz

The above values are for a Dickson DNL 3.9A FET. (See Question 11.145.)

12.259 ***Describe a phase-splitter circuit using a field-effect transistor (FET).*** —The FET phase splitter performs the same function as its vacuum-tube counterpart. The circuit shown in Fig. 12-259A is an adaptation of the source follower shown in Fig. 12-258A. Resistors of 47K in both the drain and the source were found to be quite satisfactory, regardless of the supply voltage. As the signal is split at the output, only half the input voltage is available at each output; however, the feedback is nearly as effective in reducing distortion as in the source follower. Greater output may be obtained by increasing the supply voltage until the breakdown is reached. In this manner, sufficient signal-output voltage may be obtained to drive a pair of EL 34 tubes in push-pull to 100-watts output, with only 23.4 volts for each grid. An FET operating with 160 volts can easily supply this voltage with less than 1-percent distortion.

A direct-coupled phase splitter is shown in Fig. 12-259B, and a source-coupled phase inverter is shown in Fig. 12-259C. Signal polarities have been indicated to show how phase-splitting and inversion is obtained. Referring to Fig. 12-259B, in this circuit the biasing is adjusted in favor of the phase splitter. As a rule, optimum performance is achieved with the drain-to-source voltage at 45 percent of the supply voltage; the gate voltage should run about plus 33 volts dc. The optimum drain voltage

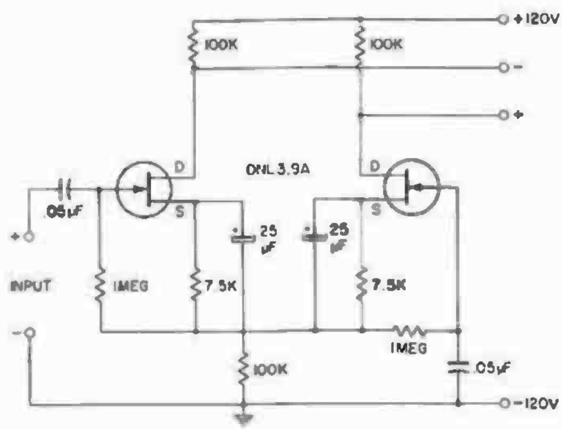


Fig. 12-259C. Source phase inverter using two FET's (after Rheinfelder).

for the first stage is plus 55 volts dc. Therefore, improved performance may be obtained by lowering the supply voltage at the first stage by the insertion of resistor R3 and capacitor C1. These components also aid in decoupling the two stages. Lowest distortion is obtained when resistor R1 equals 2100 ohms unbypassed, and R2 is 100,000 ohms. The distortion for these values is 0.97 percent, with an output signal voltage of 15 volts each side, and a gain of 20 dB. It should be realized that the distortion figures for this and other circuits are those *without* the benefit of negative feedback, which would be normally used in a high-quality amplifier. Changing the value of R1 to 7500 ohms, R2 to 220,000 ohms, and with R3 at zero ohms, and the supply voltage at 200 volts, the signal voltage on each side for 1-percent distortion is 23 volts. Increasing the input sufficiently to obtain 30 volts at the output results in a distortion of 3.4 percent, with a gain of 43 dB. If negative-feedback is used, it is applied to the top of R1, which is not bypassed. (See Question 11.145.)

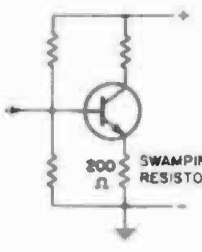


Fig. 12-260. Configuration of a swamping resistor connected in the emitter circuit of a transistor.

12.260 Describe a swamping resistor.—This term is used with resistors employed for several different purposes. One of the most common uses of such resistors is its connection in the emitter circuit of a transistor (Fig. 12-260). The base-emitter junction generally has a negative temperature coefficient; that is, the resistance decreases with the increase of temperature, permitting leakage currents to increase. Typically, the V_{e-b} will rise about 2.5 millivolts per degree centigrade, forward biasing the transistor, with the possibility of causing thermal runaway. If the swamping resistor is large in comparison to the emitter-base junction resistance, it will effectively swamp out the negative-resistance effect. This resistor is generally referred to as an emitter resistance, and for the average small transistor equals about 200 ohms.

12.261 Describe the construction of an integrated circuit element.—Integrated circuitry is a technology developed over the past decade by several companies whereby a *single* tiny module can contain several transistors, diodes, and passive elements on a single substrate. The most striking property is their microscopic size. A single module may contain up to 10 to 20 transistors and 40 to 60 resistors on a single piece of silicon 1/10 to 1/20-inch square. Arrays of resistors and capacitors are created in the form of *thin-films*.

The fundamental requirement of an integrated circuit is that its components be processed simultaneously. By the use of sophisticated thin-film techniques, *active* as well as *passive* components are formed. However, most integrated circuits are not based on thin-film techniques. Silicon planar techniques similar to the manufacture of transistors are used.

A typical integrated circuit is shown in Fig. 12-261A. The circuit for a typical element is given in Fig. 12-261B. Employing conventional construction methods, this circuitry would require very many external connections. As shown here, the external connections are reduced to 10. In addition to the tremendous reduction in size, the operating characteristics are greatly improved.

The basic procedure for the manufacture of an integrated circuit is given in Figs. 12-261C to I. The process is started by diffusing two n-type crystals in a single uniform crystal of p-type characteristics (Fig. 12-261C) from a uniform p-type crystal (Fig. 12-261D), using the masking properties of silicon and photochemical techniques with the control of time and temperature. In this manner *isolated nodes* are achieved. Diodes are thus formed by the p-type substrate and n-type nodes. Transistors are formed by the addition of p-type and n-type regions (Fig. 12-261E). The silicon wafer is then coated with an insulating oxide layer, and the oxide opened to permit the metalization and interconnections (Fig. 12-261F).

Resistors are formed by omitting the n-type emitter diffusion and two con-

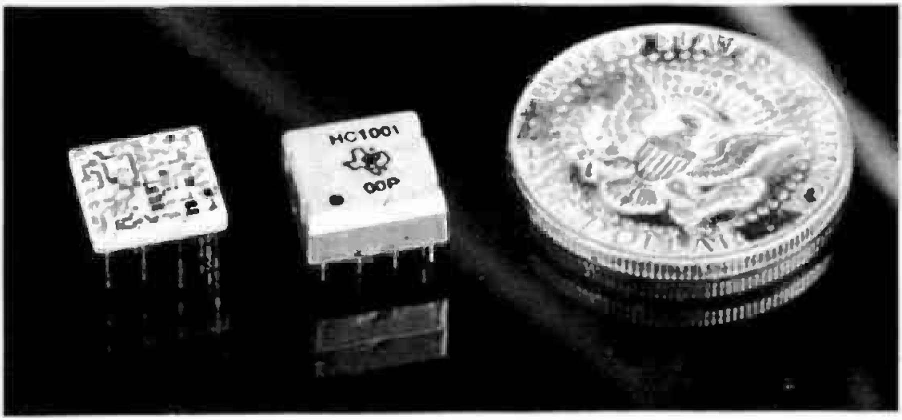


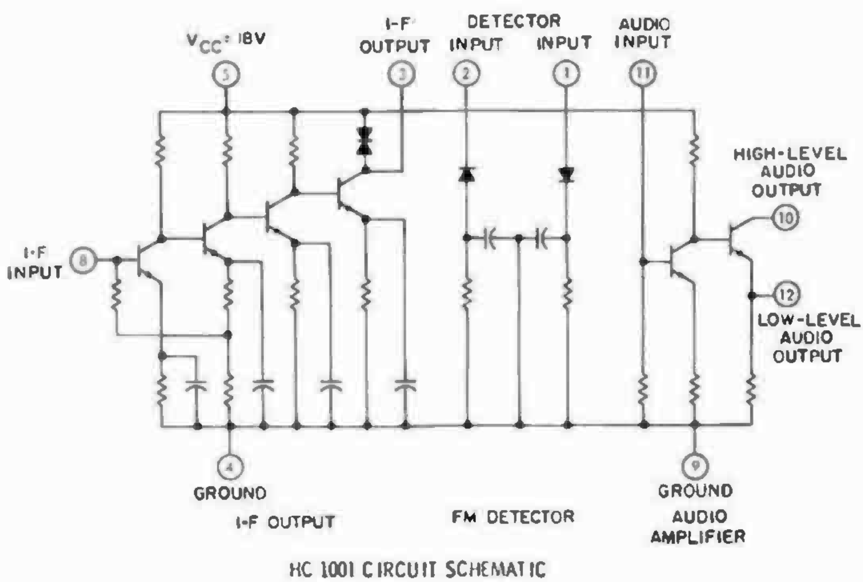
Fig. 12-261A. A typical hybrid IC circuit, used as an fm sound system module. (Courtesy, Texas Instruments, Inc.)

tacts are made to a p-type region (Fig. 12-261G). For capacitors, the oxide is used for the dielectric (Fig. 12-261H). A typical combination of three different elements is shown on a single wafer (Fig. 12-261I).

By employing the above mentioned techniques for the formation of an integrated circuit, the circuit elements thus formed are similar to discrete elements. However, integrated resistors are quite different from those of discrete design, as these are normally manufactured to a standard form factor, with the value of resistance obtained by a variation in the resistive material. In the design of integrated circuits the value of the semiconductor resistor depends on its geometry. Large values of resistance are long and narrow whereas small values are short and squat.

The value of a capacitor is equal to the product of its area, the ratio of the dielectric constant of the diffused material, and the thickness of the oxide. In practice, the area is kept constant and the capacitance varied directly with the area.

An important factor in the design of semiconductor resistors is operating temperature, making it rather difficult to achieve a close tolerance. Therefore, the ratios of the photolithographic masks



HC 1001 CIRCUIT SCHEMATIC

Fig. 12-261B. Interior circuit of IC module shown in Fig. 12-261A.

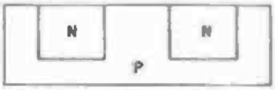


Fig. 12-261C. Diffusion of n-type areas in order to provide isolated circuit nodes.

Fig. 12-261D. Starting material for an integrated circuit, using a silicon wafer.

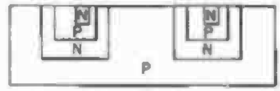


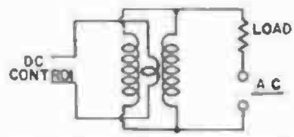
Fig. 12-261E. P-type and n-type regions to form transistors.

used in the manufacture are closely controlled.

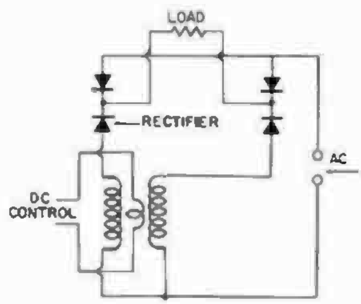
Where several transistors are included on the same chip, they have a number of operating advantages over that of the discrete unit design. Because of their proximity, they receive almost identical processing, are closely matched and because of the close spacing, suffer from minimum temperature differences. Thus a close match is possible over a wide range of operating temperatures. In addition, many more transistors can be included in a given area with their circuit elements than could be with the discrete element design.

12.262 ***Describe a magnetic amplifier and its purpose.***—Magnetic amplifiers are used in servomechanisms for the control of electromechanical devices. They are also known as direct-current transformers, current transformers, saturable reactors, and amplistats. Such amplifiers consist of one or more saturable reactors, the controlling voltage being from a dc source. Magnetic amplifiers can be designed to deliver almost any amount of power in the 60 to 400 Hz range.

Magnetic amplifiers can also be designed to operate at radio frequencies; however, their frequency range is limited by the transformers and rectifier elements.



(a) Ac output, dc control voltage.



(b) Rectified ac output controlled by dc.

Fig. 12-262. Magnetic amplifiers.

Time-constants play a large part in the design of magnetic amplifiers. To reduce the time-constant to a minimum, large amounts of both negative and positive feedback are employed. To increase the gain several stages may be cascaded, the total gain being that of the individual stages. In such systems a gain of one million is not uncommon; however, a low time-constant can only

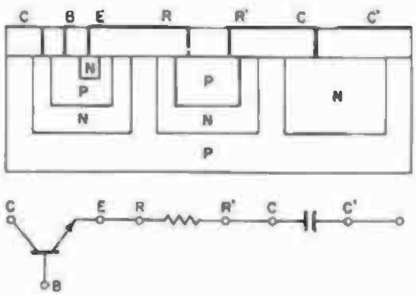


Fig. 12-261I. Completed integrated chip, containing a transistor, resistor, and capacitor.

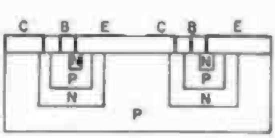


Fig. 12-261F. Metalized contacts for transistors.

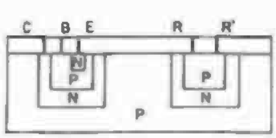


Fig. 12-261G. Contact connection to p-type region to form integrated resistor.

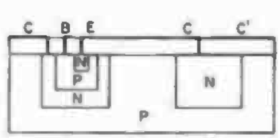


Fig. 12-261H. Diode used as a dielectric to form an integrated capacitor.

be achieved with low gain. Typical circuits are shown in Fig. 12-262.

12.263 Describe a three-terminal amplifier.—It is a European expression to denote a stereophonic amplifier where two amplifiers are combined on a common chassis. As there are two plus-minus terminals and the ground common to both, it is termed a three-terminal amplifier (Fig. 12-263).

12.264 What is a fluid amplifier?—Fluidic amplifier systems are composed of devices employing aerodynamic and mechanical principles, and are designed to operate similar to an electronic amplifier. Such devices may be constructed to produce the action of a gate, resistor, flip-flop, binary counter, and many other devices similar to their electronic counterparts. They require no moving parts, and have very low wear factor, and are not affected by nuclear radiation.

The fluid use for their operation is air maintained at a constant pressure. The amplifier devices consist of channels and passages. The advantages of such devices are they are free from vibration and have been operated up to 15,000 G's with few adverse effects.

12.265 Give a block diagram for a typical stereophonic control center.—Stereophonic control centers consist of a group of control circuits and preamplifiers for the reproduction of phonograph records, microphones, magnetic-tape reproducers, radio, and any other source of electrical pickup. Stereo centers do not contain power amplifiers for driving loudspeakers; these are separate units. Stereophonic-amplifier assemblies containing power amplifiers are called integrated stereo amplifier systems. In some instances the center

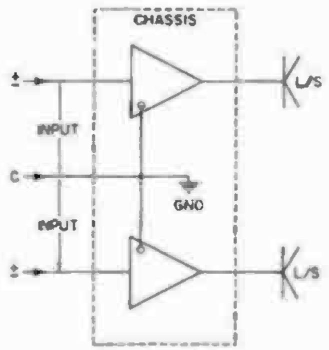


Fig. 12-263. Three-terminal amplifier.

may include an additional output circuit for driving a center or remote speaker derived from the left and right channels.

Referring to the block diagram in Fig. 12-265A, the signal will be traced for the left channel. The right channel is the same in all respects, except for one slight modification. Assume the signal is from pickup unit A, which is fed to selector switch B. This switch selects the desired input signal and equalization (magnetic or ceramic pickup, response flat for microphone and radio) and applies it to preamplifier C, then to switch D. This switch permits the user to select the proper circuitry for balancing the left and right channels, reproduction for either stereo or monophonic records, and the reversing of the right and left sides.

The signal is next applied to a rumble filter E for reducing turntable rumble (or rumble recorded in the record), then to high- and low-frequency equalizers F and G (tone controls). This signal then passes to a scratch filter H for reducing surface noise. This is followed by an overall gain control I (volume control), in combination with a switch for converting the volume control to a *loudness* control. In this latter position, the frequency response is equalized to compensate for the human ear characteristic, when reproducing at low levels (see Question 5.65). The signal is now sent through amplifier J and on to the output terminals, an external power amplifier, and loudspeaker.

The right channel is the same except for the additional amplifier stage L. This stage operates in combination with switch K, a phase-reversal switch, for reversing the phase of the signal fed to the right power amplifier. This may be necessary when reproducing stereophonic records that were recorded before the phasing standard for recording stereo records was initiated.

Connected across the left and right output circuits is potentiometer O and amplifier stage N for deriving a center or remote-amplifier signal. Potentiometer G is for achieving a balance between the two sides for the best stereophonic reproduction, and is adjusted using a monophonic or special balancing record.

It will be observed that switches B and D are two-section ganged switches with identical circuitry in both channels. The overall gain control I is also a two-gang component. The two sections of the control must track within 1 dB over their entire range to assure that the balance between the two sides is maintained as the volume control is adjusted. Also, precautions must be taken that initially the left and right power amplifiers and loudspeakers are in phase. This latter subject is discussed in Section 20.

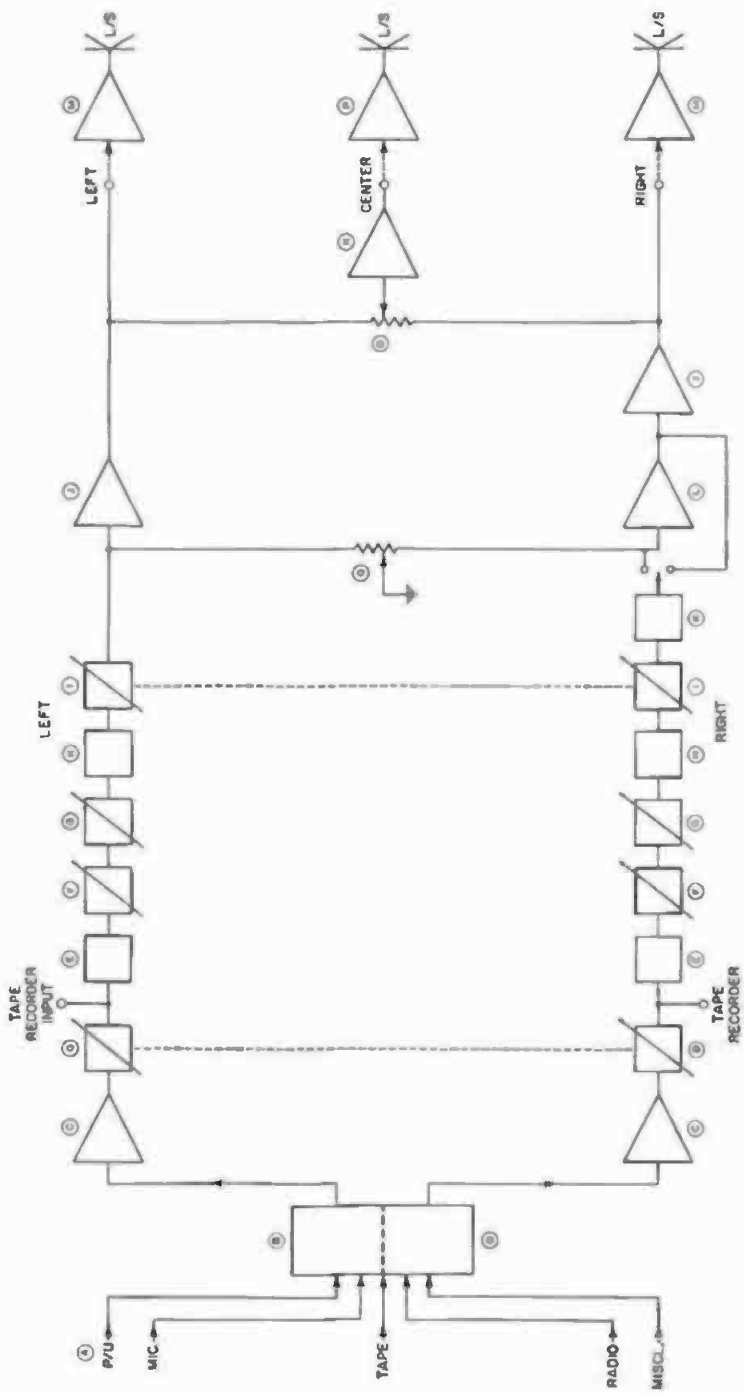


Fig. 12-265A. Block diagram for a typical stereophonic control center.

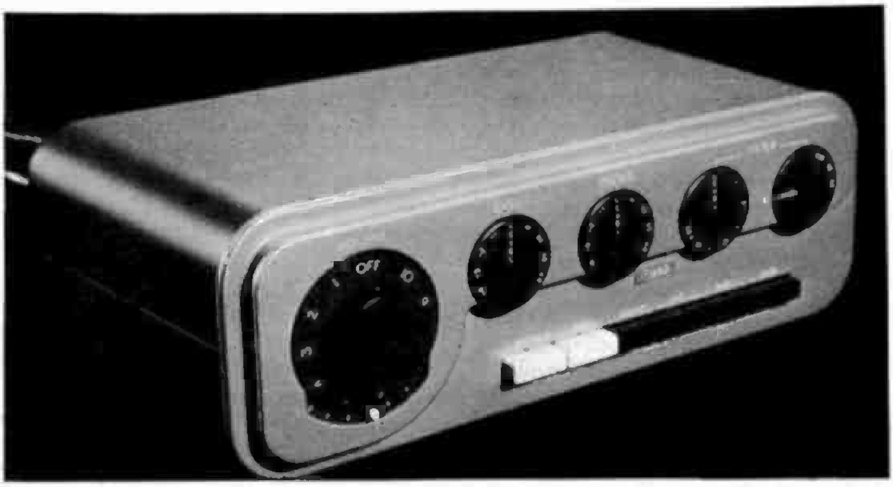


Fig. 12-265B. Stereophonic control center Quad 22, manufactured by Acoustical Manufacturing Co., Ltd. (England)

The block diagram does not represent any particular manufactured unit. The circuits shown are only to illustrate the controls and components found in an average stereo center and may vary considerably from that shown. In addition, a circuit is generally provided for driving a magnetic tape recorder, and for monitoring the program material being recorded.

Pictured in Fig. 12-265B is a stereophonic control center manufactured by Acoustical Mfg. Co. (England). At the left is the master volume control, with a stereo balance control in the slot below the knob. Along the bottom is a group of push-button switches for selecting the type of reproduction with the desired input and the correct equalization. Above the push buttons are four knobs. The first two are the conventional treble and bass controls. The remaining two are controls not found on all stereo centers. The right control of this pair is an m-derived low-pass filter with cutoff frequencies of 5, 7, and 10,000 Hz, which may be switched in half-octave steps. This permits the high end to be adjnsted for surface noise of the best reproduction within the scope of the program material. The remaining knob is for adjusting the effectiveness of the filter. There is a choice of four playback characteristics, obtained by pressing a combination of buttons in accordance with a chart supplied by the manufacturer.

At the rear are provided facilities for connecting the sources of signal inputs, tape recording, and outputs for power amplifiers and miscellaneous facilities. The overall frequency response of the unit is 20 to 20,000 Hz; distortion is 0.10 percent at 1.4 volts output; noise is minns 70 dB below 1.4 volts; crosstalk is 40 dB, 20 to 20,000 Hz.

Shown in Fig. 12-265C is an integrated solid-state stereophonic amplifier assembly, manufactured by Sherwood Electronic Laboratories, Inc. In

Fig. 12-265C. Integrated stereophonic amplifier assembly Model S-9000A manufactured by Sherwood Electronic Laboratories, Inc.

this assembly are included two solid-state power amplifiers capable of developing 120 watts of continuous sine wave power, and 160 watts of music power, both at 8-ohms output impedance. The front panel controls include an input selector switch, stereo-monophonic switch, bass and treble controls, and loudness and balance controls. In this unit, the base and treble controls are common to both channels, which is quite satisfactory if the speakers are matched. At the lower portion of the panel are a group of switches for monitoring, high- and low-pass filters, loudness control, and speaker on-off switch. Also included is a headphone jack which cuts off the speakers when it is used.

The specifications for this unit are: harmonic distortion at sine-wave rated output 0.25 percent, below 10 watts 0.05 percent; intermodulation distortion at rated output 0.25 percent, below 10 watts, 0.10 percent; damping factor is 40; hum and noise is 70 dB below rated output, using a weighted response curve. Power bandwidth for 0.5-percent THD is 12 to 25,000 Hz.

The solid-state components consist of 23 silicon transistors and 4 silicon diodes. (See Question 12.231 and Fig. 12-231H.)

12.266 Describe the basic design principles of a Darlington circuit pair.—The Darlington circuit (Fig. 12-266A) is a direct-coupled transistor pair, whereby an emitter-follower Q1 is used as the load for a previous emitter-follower. The gain of this circuit is approximately *unity*, with a current gain equal to the product of the beta's of the transistors involved. The two transistors may be viewed as a single unit. The resistor in the emitter of Q1 is also across the emitter-base circuit of Q2. The voltage drop across R1 is equal to its own voltage drop, plus that of R2 in the emitter of Q2.

A second version of a Darlington pair is given in Fig. 12-266B. Here a diode D1 having the same voltage characteristics as the base-emitter junction of Q2 is connected. Assuming that the voltage across the diode and resistor R2 are the same, the currents through both resistors are the same. Such circuits are often used in integrated circuitry (IC), where a *high* input resistance is required. The input resistance is the input resistance of Q2 multiplied by the beta of Q1. In some instances, resistor R1 (Fig. 12-266A) is omitted.

A Darlington pair (General Electric D16P4 I/C) connected for use with a

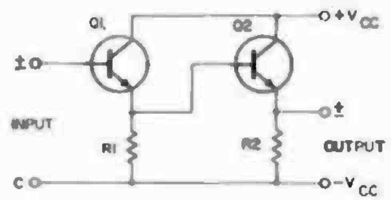


Fig. 12-266A. Basic circuitry for a Darlington pair transistor preamplifier.

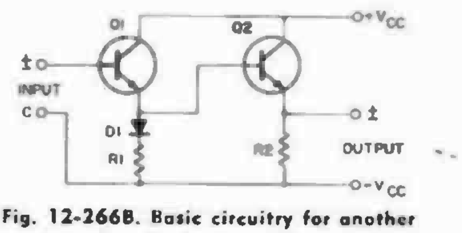


Fig. 12-266B. Basic circuitry for another version of a Darlington pair transistor preamplifier.

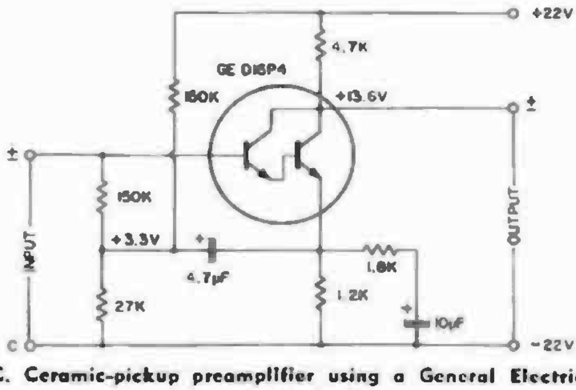


Fig. 12-266C. Ceramic-pickup preamplifier using a General Electric D16P4 integrated circuit.

ceramic phonograph pickup is given in Fig. 12-266C. In this circuit, the input resistance is on the order of 2.7 megohms at 50 Hz. The frequency response is 40 to 12,000 Hz within ±2 dB, with a THD of 0.2 percent, 1 to 15kHz. Intermodulation distortion is approximately 1.5 percent, using 60 and 6000 Hz in a 4:1 ratio for a 1.9-volt output. This level is 10 dB *below clipping*. The unweighted noise level is 80 dB, referred to 1.9 volts. An input signal level of 250 millivolts is required to produce an output of 1.9 volts.

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Section 13

Disc Recording

The invention of the cylindrical phonograph and record by Thomas A. Edison in 1877 resulted in a method of recording sound using a constant groove-velocity vertical sound track, without tangent error. However, because of the nature of the record, each had to be a master; therefore, large production became rather a problem.

Emil Berliner, with his invention of the laterally recorded flat disc record, sacrificed the feature of constant groove-velocity recording; thus, tangent error and a loss of high-frequency response resulted as the smaller diameters were approached. However, the disc record was more practical and economical and could be easily reproduced in great quantities by stamping from a metal master. Because of these features, the disc record became the standard for the phonograph industry. Eldridge Johnson, a machinist with Berliner, contributed many improvements to the original reproducing machine of Berliner.

Since the introduction of electrical recording in 1927, the search has been endless for a better recording media, mechanical drives, speed stability, less tangent error, recording heads, pickups, reproducing machines, and materials for pressing the final disc.

Although Blumlein, in 1931, first proposed the 45/45-degree stereophonic method of recording, it has only been in recent years confirmed as standard by the recording industry.

Stereophonic and microgroove recording demonstrate the mutability of the overall disc record industry, all the way from foil to vinyl. Coarse pitch to microgroove, pickup arms weighing from pounds to grams, and frequency ranges extending beyond audibility are all part of the progress of the industry.

Various types of disc recording and the techniques are discussed in this section, with descriptions of recording lathes and their associated equipment.

13.1 What is an electrical transcription?—A disc recording used in radio broadcasting for the transmission of program material, recorded on 10- or 12-inch discs, using microgroove techniques. Before the advent of microgroove recording, transcriptions were recorded in 16-inch disc records, using either lateral or vertical recording at 96 to 150 lines per inch. Many of the transcriptions heard today are actually magnetic tape.

13.2 Describe a disc recording lathe.—A commercial recording lathe, manufactured by Neumann of West Germany, is pictured in Fig. 13-2A, with its principal components indicated. Basically the recorder consists of a heavy steel base, mounted on shock mounts supported by a steel cabinet A. Turntable B weighs 65 pounds, with three stroboscopic rings on its outer rim for three speeds which are illuminated by a neon light. The turntable is isolated from the drive system below by means of an oil-filled coupling, thus preventing rumble and flutter being transmitted from the drive to the turntable. The turntable is driven by a film of oil between two concentric cylinders at C. The lathe bed D is of the slide type, with two ball-bearings riding on top the bed to relieve strain placed on the sled E by the weight of the cutter suspension and cutting head F.

Directly below the lathe bed is a calibrated scale on which are mounted the starting cams and end-groove stop.

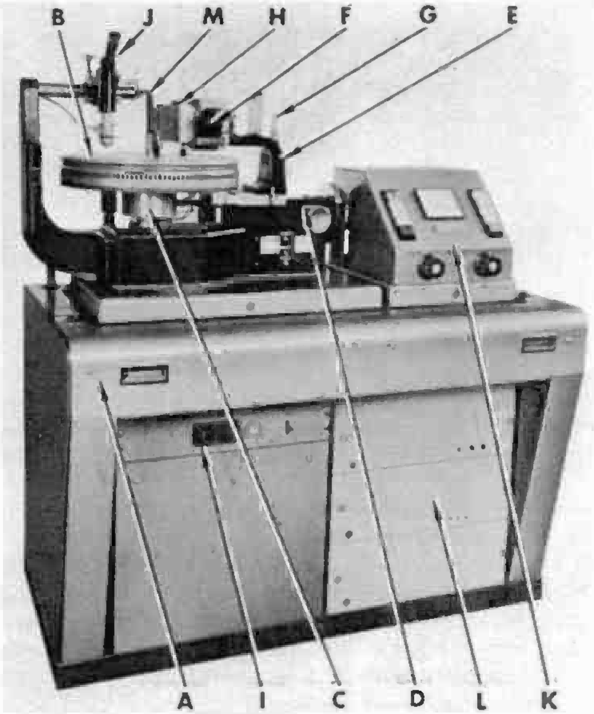


Fig. 13-2A. Neumann master disc recording lathe Model AM-32b with automatic pitch and depth control. (Courtesy, Gotham Audio Corp.)

Three cams for 7-inch, 10-inch, and 12-inch discs are provided. An adjustable end-groove stop for the three standard RIAA groove diameters cause the cutting head either to lift immediately (with eccentric grooves) or with an adjustable delay to provide for a locked groove. A lead screw engaging lever G is interlocked in such a way that the cutter will lift at any time it is not being driven by the lead screw, with a braking assembly to prevent the lead screw from coasting when the end is reached. A vacuum system provides a means of holding down blank discs from 10- to 17¾-inches in diameter on the turntable, with a disabling valve to shut off the vacuum holes when they are not in use.

Cutting-head connections are brought to the rectangular box H on the transport sled E by means of a plug to permit the exchange of cutting heads without disturbing the alignment. The connector plugs (for stereo) consist of 6 pins which carry the audio signal, feedback loop, and dc for the stylus heating coil. A release solenoid lifts the

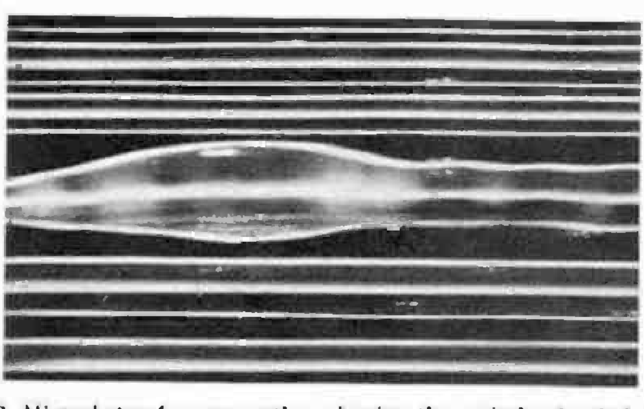


Fig. 13-2B. Microphoto of groove action, showing the variation in pitch and depth.

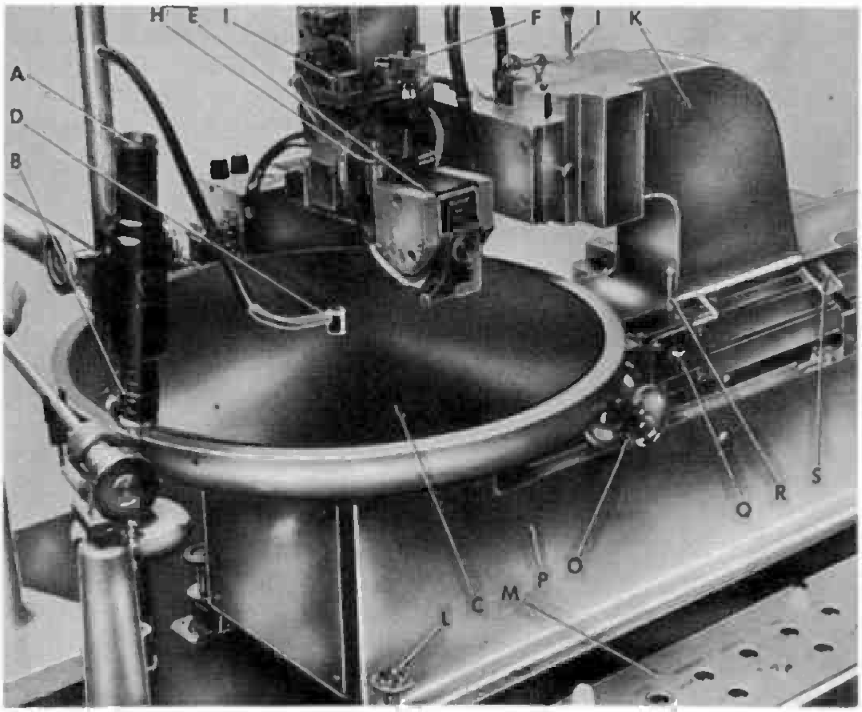


Fig. 13-2C. Scully Recording Instruments Corp. master disc recording lathe with automatic pitch and depth control.

cutting head whenever the stop button I is depressed, when the sled hits the end groove-stop, or when the lead screw is disengaged. A dash pot on the front of the cutting-head suspension mechanism is equipped with a perforated piston, with an adjustable shield over the perforations, to allow a wide latitude of adjustment. A tilting mechanism is connected to a moving-coil system which, together with the depth-of-cut control, provides electronic-depth variation, thus eliminating the advance ball generally used.

The depth-of-cut control supplies direct current to a moving-coil system in the cutting-head suspension mechanism, thus relieving the cutting-head pressure on the disc. This operation is controlled by a potentiometer, which is adjusted while observing the cut through a microscope J. A second control presets the increased depth used with the lead-in, lead-out, and spiraling grooves. The variable-pitch control K is a separate piece of equipment situated at the right end of the lathe. The pitch-control mechanism is self-driven and coupled to the lead screw by a four-way shock-isolated coupling. The variable-pitch motor is connected by means of a belt, and through an oil-filled flexible coupling to the lead screw. A copper disc on the shaft of this motor runs over the pole of an electromagnet, in which a direct current flows, producing a braking action, which stabilizes motor revolutions. A second motor, identical to the belt-connected motor, connected to an overdrive in the gear train, serves for the speed-up of pitch for lead-in and lead-out spiraling.

The turntable motor is of the synchronous type. The motor consists of a gearlike armature about 10 inches in diameter, rotating inside a similar inside gear. By means of a winding, a rotating magnetic field is set up, causing the armature to rotate. The wow and flutter of this particular machine is 0.035 total rms.

For automatic pitch control, three control amplifiers are necessary, and are mounted in the lower portion of the cabinet. A preview head is mounted on the magnetic-tape playback transfer machine to provide the control amplifiers with advance knowledge of the modulation to be fed to the cutting head. The output of the control amplifiers associated with the preview head is fed to the control amplifiers. For monophonic recording, the depth of cut is held *constant*, while the pitch is varied as a function of the preview information. In stereophonic recording, the pitch control is actuated by the *sum* of the left and right channel signal, while the depth of cut is varied according to the difference signal obtained from a stereophonic preview head. It is the function of the pitch-control amplifier to translate the preview signal through an equalizer into variations of braking current, which in turn are applied to the pitch-control motor to vary its speed and, with it, the lines per inch of recording.

The depth-control amplifier is identical to the pitch-control amplifier; however, its output to the solenoid in the cutting-head suspension produces a varying relief of cutting-head pressure, acting against a counterbalancing spring on the cutting-head mounting mechanism. A microphotograph of a group of recorded grooves, showing the action of the variable pitch and depth control appears in Fig. 13-2B. In the recording of both lateral and vertical modulation (stereo), increased depth requires increased pitch, so any deepening of the groove caused by the depth-control amplifier must be translated into increased pitch. This is accomplished by an integrating amplifier, which adds to the pitch-control current whenever increased depth is required. It is claimed that such a system, when properly adjusted, can add up to 6 minutes of recording time on a 12-inch record.

The microscope is 156 power with concentric illumination, brightly lighting the groove and leaving the land between the grooves dark, and is moved across the turntable by means of a rack and pinion gear. The microscope support arm also acts as the vacuum conductor for the vacuum-chuck turntable. The microscope graticule is calibrated to read in 0.001-inch graduations. The chip collector is contained in the lower portion of the cabinet. A pickup arm is generally monnted at the left for playback purposes, and may be used for simultaneous monitoring of the playback signal, while also monitoring the signal from the negative-feedback loop to the cutting head. A half-speed converter permits the turntable to be rotated at 16⅔ rpm, and half-speed for 45 or 78 rpm for experimental work or for cutting frequency discs.

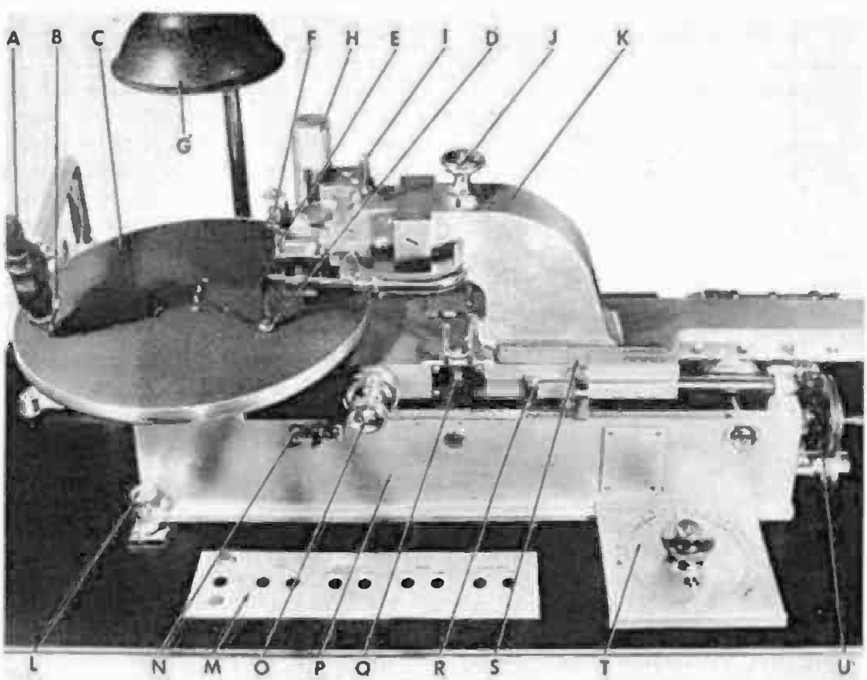


Fig. 13-2D. Scully Machine Co. recording lathe.

A second recording lathe, manufactured by the Scully Recording Instrument Corp., appears in Fig. 13-2C. Like the Neumann lathe, it also has variable pitch and depth control and many of the features previously described. At A and B is the microscope and its light, C the turntable, D a vacuum-suction pipe for removing the chip from the disc, E the cutting head, F depth-control adjustment, H cutting-head dash pot, I plug connection for the cutting head, K sled, L leveling screws, M push-button control panel, O carriage hand knob, P recorder base, Q feed-nut adjustment, R carriage-limit stop, and S line-per-inch indicator scale. The motor with its gear box and mechanical filter are mounted below the table.

In Fig. 13-2D is shown an overall view of a similar recorder with a monophonic cutting head. The letter symbols correspond to the previous view. By means of the control T, the lines per inch may be varied over a range from 70 to 400 lines per inch. At U is a lead-screw hand wheel. Automatic control of pitch and depth are accomplished in a manner somewhat similar to the Neumann lathe. Anticipation amplification is approximately 20-dB greater in the 50 to 60-hertz range, returning to normal frequency response around 750 Hz. A gain control adjusts the amplitude of the original signal from the control head on the magnetic-tape reproducer. An excursion control determines the amount of decrease in the number of lines per inch for a particular recording. A return control adjusts a time-delay network in the output of a signal-control amplifier. Delay in returning the feed from coarse to fine permits the continuance of sufficiently coarse lines to prevent post-groove over-cuts.

13.3 Define the term lateral recording.—Lateral disc recording was invented in 1888 by Berliner and Johnson for engraving a sound track on a disc record. A stylns mounted in a cutting head (recording head) is actuated by the sound modulatious and moves at right angles to the direction of travel of the recording blank. In lateral recording, the dept of cut and groove width are nearly constant. Both monophonic and stereophonic sound may be recorded by this method.

13.4 What is vertical recording?—Vertical recording is the principle used by Thomas A. Edison in his original cylinder recorder pictured in Fig. 13-4. The sound waves are picked up by a small metal horn and directed to a metal diaphragm to which is attached a stylus. Pressure changes in the horn, caused by the sound waves, moved the diaphragm causing the sylus to indent a sound track in a vertical direction on the surface of a revolving tin-foil cylinder. The sound track recorded on the tin-foil cylinder was reproduced by the same stylus riding in the indented sound track. The movement of the sty-

Fig. 13-4. Thomas A. Edison with his original tin-foil phonograph. Taken in Washington, D.C., April 18th, 1887, where he demonstrated the machine before the National Academy of Science. (Courtesy, Smithsonian Institute.)

lus caused the diaphragm to which it was attached to vibrate, disturbing the air column in the horn. Thus, sounds were generated which were similar to the waveforms inscribed on the tin-foil cylinder.

In a modern vertical-recording head, the actuating mechanism is designed to eliminate any lateral motion. Thus, distortion caused by the combination of a lateral and a vertical motion is eliminated. The sound track made by a vertical recording head varies in both depth and width. Because the variation in width of the groove is small, a greater number of lines per inch may be recorded in a given space.

At one time, vertical recording was considered to be superior to lateral recording; however, with the development of microgroove lateral recording, vertical recording is now considered to be obsolete.

In the original Edison cylinder recorder shown, the record revolved at a constant speed; therefore, the machine recorded with a constant-groove velocity. In the present day recorders, the angular velocity of the turntable is constant and the groove velocity decreases as the smaller diameters are approached. (See Question 13.22.)

13.5 What is hill and dale recording?—It is another name for vertical recording.

13.6 What is a long-playing record?—A disc record having a playing time longer than 5 minutes, 10 to 12 inches in diameter and turning at 33⅓ rpm, recorded with approximately 150 to 300 lines per inch.

13.7 Define a microgroove recording.—A microgroove recording has from 200 to 300 or more lines per inch, with a groove width in the pressing suitable for reproduction, using a stylus having an included angle of 40 to 55 degrees and a tip radius of 0.001 inch.

For stereophonic recording the lines are the same, but the groove width in the pressing must be such that it may be reproduced using a stylus of 45 to 55-degrees with a tip radius of 0.5 to 0.7 mil. Styli dimensions are discussed in Section 15.

13.8 What are the advantages of microgroove recording?—Longer playing time with low surface noise, wider frequency range, lower distortion and greater dynamic range. The weight of the pickup on the record has been reduced from 27 grams for older type recordings to 0.5 to 3 grams. This lower playback weight results in longer life of the record, lower distortion, and better tracking, and removes the pressure from the side walls of the groove. Because of the light pressure, it is important that the playback turntable be level; otherwise, the pickup will skate across the record if jarred.

The reproducing stylus for microgroove recordings has a tip radius of 1 mil (or less) and the included angle is 45 degrees. (See Question 15.5.)

13.9 What are standard turntable speeds?—Standard speeds are 33⅓ and 45 rpm. It is recognized that both 78.26 rpm and 16-inch coarse-pitch transcriptions running at 33⅓ rpm are still in limited use; however, these two latter speeds are no longer considered standard. A speed of 16⅔ rpm is also used for special recording purposes, but it also is not considered standard.

13.10 What medium is used for recording disc records?—Originally soft wax was employed, but in the early 1930's this was replaced with the modern acetate disc, which has also undergone several changes in its material composition since its introduction.

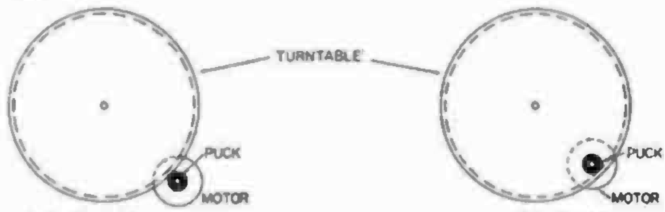
13.11 What materials other than acetate are used for recording on disc records?—For purposes where quality of recording is not of prime importance, materials other than acetate are aluminum, paper, and many different types of plastics.

13.12 What is a pregrooved recording blank?—A disc with a prepared groove in which the recording stylus rides. This type of blank was nsed in early-day home recording equipment to eliminate the costly lead screw and half-nut.

13.13 What is a recording lathe?—It is another name for a disc recording machine. (See Question 13.2.)

13.14 What is a direct-drive disc recorder?—One in which the power is transmitted to the turntable from the motor by means of gears. (See Question 3.71.)

13.15 What is a rim-driven turntable?—One in which the motive power is transmitted to the turntable by a puck on the motor shaft in contact with either the inside or outside edge of the turntable. When driven on the outside



(a) *Outside.* (b) *Inside.*

Fig. 13-15. Rim-driven turntables.

edge, a rubber tire is placed on the outer edge of the turntable which makes contact with a metal puck on the motor shaft. These two methods are used more than any other type of drive system. Typical rim drives are shown in Fig. 13-15.

13.16 What is an idler?—A wheel placed in the mechanical transmission system between the driving member and the turntable in a recorder or reproducing machine.

13.17 Does an idler between the motor and the driving puck have any effect on the speed?—No. It only serves as a means of transmitting the power from the driving source to the turntable. The turntable size and the diameter of the driving puck on the motor determine the speed. (See Question 3.71.)

13.18 What is a puck?—A circular piece of metal or fiber used to transmit power from the driving source to the driven member.

13.19 What is linear speed?—A constant velocity in a given direction.

13.20 What is a constant angular velocity device?—A device driven at a constant speed in rpm.

13.21 What is a weight-driven recorder?—Because of the lack of constant speed drives, early recorders were driven by means of a weight mechanism similar to that of a clock. The recorder was set on a tripod about six feet above the floor. Heavy weights, wound up by means of a windlass, supplied the motive power which was controlled by a centrifugal governer.

13.22 What is a constant-groove velocity recorder?—A disc recording machine so designed that as the recording progresses across the surface of the record toward the smaller diameters, the turntable speed is increased. In this manner, the groove velocity is held constant throughout the whole recording and is the same for any given diameter. The steady increase of speed is accomplished by a planetary drive system contacting the underside of the turntable. The design of this machine permits a greater number of lines per inch to be recorded, thereby increasing the playiug time. Twenty-seven minutes may be recorded at 300 lines per inch on an 8-inch disc.

The principal drawback to this type recording is that the playback machine must be of the same design as the recorder in order to obtain exactly the same increase of speed as when recorded. Generally, such machines are used with embossing-type recording heads using a constant-amplitude recording characteristic as described in Question 14.5. It has often been stated that the combination of constant-groove velocity recording and constant-amplitude recording is the classical system of recording.

Constant-groove velocity recorders use an embossing-type stylus with a 1-mil tip radius. The groove velocity is generally around 2 inches per second, at 300 lines per inch. This slow velocity is used only for dialogue recording and will give 93 minutes of recording time on an 8-inch disc. For music recording the groove velocity is increased to 10 ips. The frequency range is approximately 80 to 5500-Hz. (See Questions 13.204 to 13.206).

13.23 Describe a monogroove recording.—In an early attempt to record stereophonic sound in a single record groove, the groove carried both lateral and vertical modulation and employed a cutting head based on the design shown in Fig. 13-23. With the development of the 45/45-degree cutting head and pickup by Charles C. Davis of Westrex Corp., almost all stereo recordng system have adopted this latter system. Stereophonic cutting heads are discussed in Questons 13.217 and 14.2.

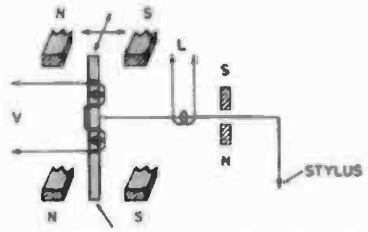


Fig. 13-23. Magnetic structure for a monogroove pickup or recording head. Sound waves are recorded in both the lateral and vertical directions simultaneously.

13.24 What is a variable-pitch recorder?—A disc recorder designed in such a manner that the lines per inch may be varied as the recording progresses. When the program material is of low amplitude, the speed of the cutting-head carriage is slowed down, thereby increasing the number of lines cut per inch. For high amplitudes, the carriage is speeded up to reduce the number of lines per inch. In this manner, greater amplitudes may be recorded and a longer recording time secured.

In the early days of motion picture recording when disc records were used, a 10-minute, 16-inch disc was reduced to a 12-inch disc in this manner. This type recording is also called *margin recording* and many present day high-quality, long-playing records are made using margin recording with a hot stylus, resulting in superior recordings.

13.25 What is a lead screw?—A threaded shaft used to move the cuttiug-head carriage across the face of the recording blank on a disc or cylindrical recorder. (See Question 13.2.)

13.26 What is cutting rate?—The number of lines per inch the lead screw moves the cutting head carriage across the face of the recording blank. Before the advent of microgroove recording, the lines per inch for coarse-pitch recording were 96, 104, 112, 128, 136 or greater, increasing in multiples of eight lines per inch. For microgroove recording, 200 to 400 lines per inch are used. 265 lines being average.

13.27 How are the lines per inch changed in a recorder?—In nonprofessional recorders, the lines per inch are fixed. In professional equipment, the lead screw is changed or the gears that drive it. In a variable-pitch machine as described in Question 13.24, the lead screw is rotated by a separate motor and the lines per inch are increased or decreased by controlling the speed of the driving motor.

13.28 In what direction are disc recordings recorded?—From the outside in and rotated in a clockwise direction as viewed from the side being reproduced. In the early days of making broadcast transcriptions (before diameter equalization), if the program material consisted of several records, they were recorded to start outside-in on the first record, with the next record recorded inside-out to compensate for the loss of high frequencies at the smaller diameters. This procedure was followed to the last record, and the listener was often unaware of the loss of high frequencies as the program changed to the next record; however, the *increase* in high-frequency response was quite noticeable as the larger diameters of the second record were approached.

13.29 What does the term coarse-pitch recording mean?—To distinguish recordings using less lines than that used for microgroove recording, they are referred to as coarse-pitch recordings. The lines for coarse-pitch recording vary between 96 and 150 lines, while those of a microgroove recording range from 200 to 300 lines per inch or more, the average being about 265 lines. Variable-pitch techniques may be applied to either type recording.

13.30 What is the angle of the recording stylus relative to the surface of the recording disc?—This subject has been uuder discussion for considerable time and to date no standard has been set for the angle of the recording stylus. The Westrex 3C and 3D cutting heads are operated at an angle of 23 degrees; however, different recording activities may set them to a different angle. Experimental work by Bauer indicates that increasing the angle of the cutting head beyond 23 degrees will reduce the discrepancy between the recorded groove and that of the pressing, and result in a recorded groove of 15 degrees. The principal cause of the discrepancy in the groove is the spring-back of the lacquer, and the recording stylus.

In the normal recording of records the groove walls form an augle of 90 degrees, which is particularly important in recording stereophonic records

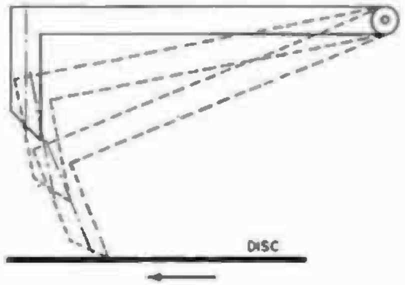


Fig. 13-30A. Because of the stylus suspension in a 45/45-degree stereophonic cutting head, the stylus inscribes an arc in the vertical mode because of the tilt from vertical.

using the 45/45-degree system. For records cut before the advent of stereo, lateral records were cut using an 87-degree *included angle stylus*, on the assumption that after processing the groove grew to 90 degrees. This was controlled in the finished product by the recording engineers, experienced with lacquer-disc recording.

Because of the nature of the stylus suspension in a stereophonic cutting head, the stylus describes an arc in the vertical mode, and because of this motion the cutting face is tilted back from the vertical (Fig. 13-30A). This results in an increase of the included angle of the groove (Fig. 13-30B). In view of this fact, and that the reproducing *stylus rides on the wall of the groove and not the bottom*, the waveform is distorted and increases with an increase in the recorded level. For the Westrex stereo system, the recording angle is specified to be 23 degrees; however, several activities have studied the angle problem and recommended that the angle of the recording stylus be changed. It is hard to predict the final angle, since factors other than stylus angle inject themselves into the study. Among these factors are: fit of the recording stylus in the recording head, stylus shank material, and the bonding of the

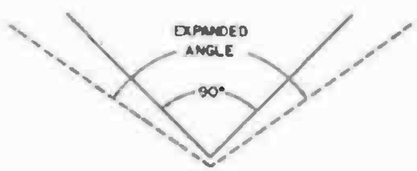


Fig. 13-30B. Increase in the groove included angle because of the tilting action of the recording stylus.

jewel in the stylus shank. Any shifting of the jewel tip because of the bonding will cause resonance and also affect the vertical angle. The jewel and stylus shank must act as an integral part.

When reproducing a record cut at a 23-degree angle (or greater), the reproducer stylus angle is set to an angle of 15 degrees to reduce the distortion caused by the difference in the angle of the recorded groove. It has been agreed by the RIAA, NAB, and the manufacturers of sound-reproducing equipment in both the United States and most of Europe that the angle for the reproducing stylus is to be 15 degrees. At the present time, this leaves the record manufacturer with the problem of establishing the correct angle for the recording stylus to produce a 15-degree groove in the finished product. Certain recording activities using coarse-pitch and microgroove techniques still use the 3-degree to 5-degree drag angle, discussed in Question 13-32. (See Questions 16.58 and 16.59.)

13.31 How is a recording stylus aligned laterally? — By observing the reflection of the recording stylus on the surface of the recording blank as shown in Fig. 13-31. When the reflected image is in perfect alignment with the actual stylus shank, the alignment is correct. (See Question 23.78.)

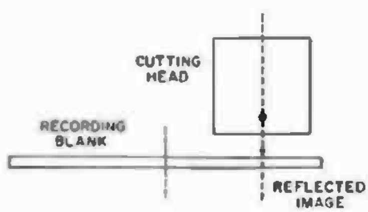


Fig. 13-31. Vertical alignment of stylus by observing its reflection on the surface of the disc.

13.32 What is the drag angle of a recording stylus?—In coarse-pitch recording used before the advent of microgrove recordiug and hot-stylus techniques, the recording stylus was given a slight angle relative to the surface of the recording disc (Fig. 13-32). The recording head was tilted backward about 3 degrees to 5 degrees against the motion of the disc. This was to prevent the stylus from digging into the recording blank aluminum base, as might occur

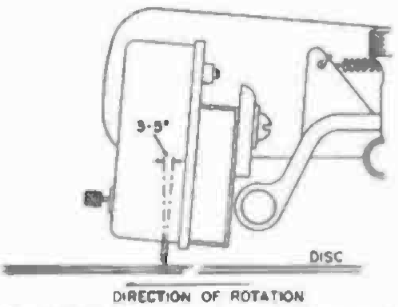


Fig. 13-32. Drag angle of early recorder cutting heads. The stylus angle is determined by observing the image of the stylus on the surface of the disc.

if the stylus had been set vertical. For stereophonic microgroove recording using a hot stylus, the stylus is set to about 23 degrees perpendicular to the surface of the disc surface. The stylus in the reproducing pickup is set with a drag-angle of 15 degrees. (See Question 13.30.)

13.33 What is flutter?—A waver in the reproduction of a tone or group of tones, caused by the irregularities in the speed of a turntable which results in frequency modulation of the tones. (See Fig. 13-33.) High rates of flutter are on the order of 10 Hz. Low rates of flutter occur one per revolution and are often referred to as once-arounds, or wow, because of the frequency characteristic. Typical resulting flutter waveforms are shown in Fig. 23-149.

13.34 How is flutter measured?—With a flutter bridge as described in Question 22.41.

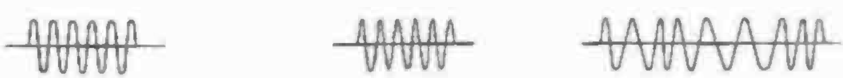
13.35 What is the standard for wow and flutter in a recorder turntable?—The average deviation measured over a range of 0.5 to 200 Hz, from the mean speed of the turntable while recording, shall not exceed 0.04 percent of the mean speed. The average deviation is measured, using a meter having the same dynamic characteristics of a standard VU meter. The term "average" is defined to mean the average of the measuring device rather than the period of time over which the measurement is made.

The measurement is conducted using a stroboscope disc illuminated by either a 50- or 60-Hz lamp operating from the normal power source. For 33⅓ rpm, the stroboscope is to have 216 bars or spots in 360 degrees; for 45 rpm, 160 bars or spots are required. For a recorder, not more than 7 bars or spots may pass a given reference point (either direction) in one minute.

13.36 What is the standard for wow and flutter in a reproducer turntable?—The average deviation from the mean speed of the reproducing turntable when reproducing shall not exceed 0.1 percent of the mean speed. The term "average" is defined to mean the average of the measuring device rather than the period of time over which the measurement is made.

The measurement is conducted using a stroboscope disc illuminated by a lamp operated form the normal 50- or 60-Hz power source. For 33⅓ rpm, the stroboscope disc is to have 216 bars or spots in 360 degrees; and for 45 rpm, 160 bars or spots. For a reproducing turntable, not more than 21 bars are to pass a given reference point in one minute, for either direction.

13.37 Describe a magnetically supported turntable.—The turntable about to be described is manufactured by Stanton Magnetics Inc., and is unusual in its design because the turntable is supported entirely by *magnetic repulsion* at the main bearing (Fig. 13-37A). The repulsion or suspension system consists of two flat circular magnets which are charged so that their force-fields are in the *vertical* plane. One magnet is placed around the bearing well and the other on the under side of the turntable, surrounding the shaft. The magnetic force-field provides an *air suspension;* in other words, the turntable rides on air. Therefore, the turntable has no mechanical contact with the rest of the structure, the turntable shaft is only used for rotational guidance. The turntable is driven through a soft idler wheel, which in turn is driven by a capstan on the



(a) *Undistorted sine wave.* (b) *High - frequency flutter.* (c) *Wow (once around)*

Fig. 13-33. The effects of irregular speed on a waveform.

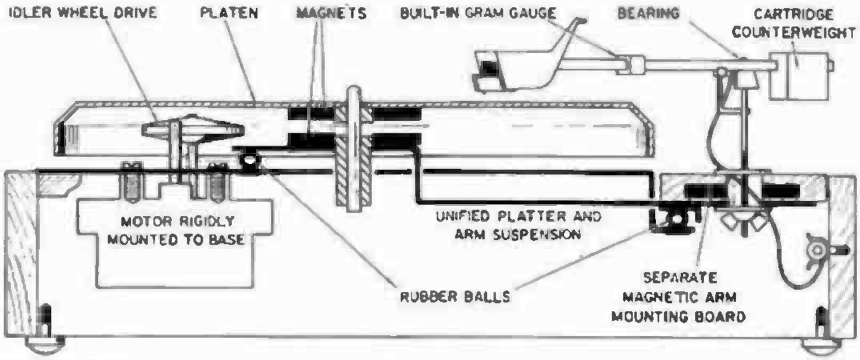


Fig. 13-37A. Cross-sectional view of Stanton Magnetics Inc. Model 800B magnetically suspended turntable.

motor shaft. This structure aids materially in the reduction of transmission of mechanical vibration to the turntable. The motor is a four-pole synchronous type, with sufficient power to quickly bring the platen up to speed.

The pickup arm is mounted on a separate structure, isolated by rubber balls, which affords acoustic isolation between the turntable and the pickup arm. The pickup arm is supported on a single pivot bearing for both vertical and lateral motion. The motor board section supporting the pickup arm is held in place magnetically and aligned with guide pins. This permits easy removal of the arm for experimental work. The arm resonance is below 10 Hz; flutter is less than 0.15 percent; rumble is minus 43 dB below the standard reference of 7 centimeters per second. A view of the complete turntable is shown in Fig. 13-37B.

13.38 What is the effect of flutter on recorded program material?—The music and voices will waver, becoming more noticeable at the higher frequencies. A piano will take on the quality of a harp. Flutter is, in reality, frequency modulation of the program material because of the irregular turntable speed.

13.39 What causes flutter in a disc recorder?—Some of the causes of flutter in a disc recorder are: slippery driving pucks, binding in the drive system, off-center hole in the disc, cutting-head carriage bind on the guide bars, chip brush dragging, and high spots in the drive gears.

13.40 Define the term "once-around."—It is a variation in the sound that occurs once per revolution of the turntable. (See Question 13.33.)

13.41 What causes a pattern or spokes in a disc record?—Vibration in the recorder or excessive alternating-current hum in the amplifier. The pattern resembles wheel spokes when observed under a light. During reproduction, patterns cause a low-frequency rumble and high background noise.

13.42 What is a moiré pattern?—A shadow pattern on a disc record caused

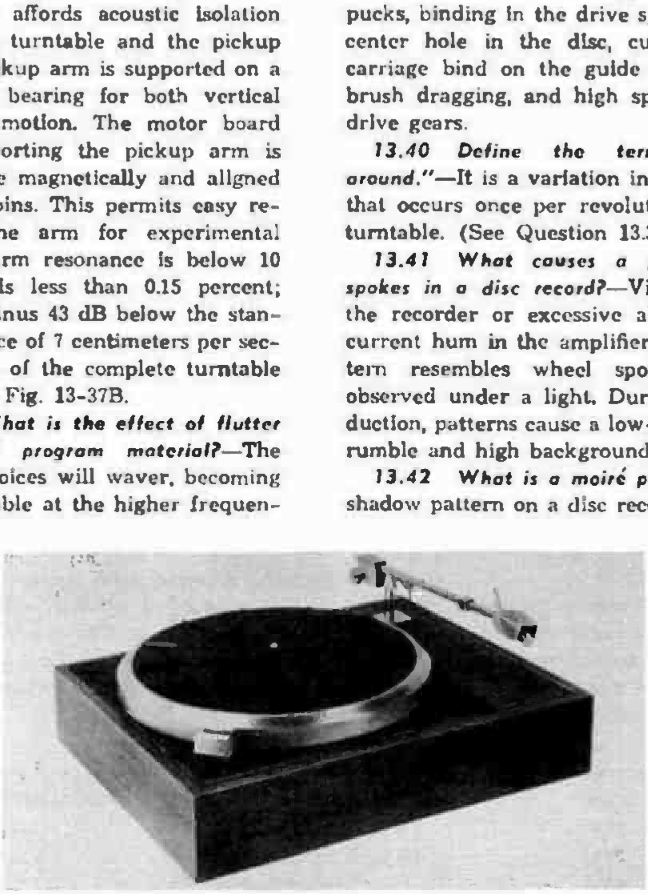


Fig. 13-37B. Stanton Magnetics, Inc. Model 800B turntable using magnetic repulsion to support the platen.

by vibration. It is so named because of its resemblance to moiré cloth.

13.43 What is the cause of rumble in a recording?—Rumble is generally a low-frequency sound caused either by the reproducing turntable drive mechanism, or it was recorded in the original master recording. Low-frequency noise and rumble is sometimes induced in the recording lathe by earth noises, drive gears, and the cutting-head carriage rails. (See Question 23.76.)

13.44 How does turntable rumble affect the reproduction of stereophonic records?—Turntable rumble has a pronounced effect since most of the rumble is in the vertical plane. Turntables employed for the reproduction of stereophonic recordings must have a low rumble factor, at least 10, and preferably 20 dB below that of a turntable used to reproduce laterally-cut records.

13.45 What is a drive hole?—An offset hole in a recording blank for setting over a pin in the recorder turntable to prevent slippage.

13.46 What is a recording medium?—Any material used for the purpose of recording a sound track. Common materials are: photographic film, acetate discs, magnetic tape, and plastics.

13.47 What does the term groove velocity mean?—The amount of disc surface passing under the stylus tip in one second. For constant-angular recording, the groove velocity is continually changing with the diameter of the record groove, decreasing as the center or smaller diameters are approached. For constant-groove velocity recording, the groove velocity is the same at any point on the disc, regardless of the diameter. (See Question 13.22.)

13.48 What are the average groove velocities for different recording speeds?—See Fig. 13-48.

13.49 Describe the effect of a hot and cold stylus on the frequency response of a microgroove recording.—The results from a study by Jackson of this effect are tabulated below. With a reference level of 5 centimeters per second, unfiltered output measurements were taken of the signal frequencies from 30 Hz to 15,000 Hz, using both a hot and cold stylus. The tests were conducted by turning the stylus heater current on and off, and each frequency recorded from an 8-inch diameter to 7.5 inches on the same disc. There was no noticeable difference from 30 Hz to 1000 Hz. The first noticeable change occured at 2000 Hz, with larger differences occuring as the frequency was increased up to 15,000 Hz. The hot-stylus recording measured an increased output over the cold stylus as tabulated below.

2000 Hz + 0.2 dB	10,000 Hz + 1.3 dB
4000 Hz + 0.2 dB	12,000 Hz + 1.6 dB
6000 Hz + 0.5 dB	15,000 Hz + 2.0 dB
8000 Hz + 0.6 dB	

Signals of 50,000, 100,000, 10,000 and 15,000 Hz were recorded, each on a separate disc at various diameters, using both a hot and a cold stylus. The results of these tests are given in the chart in Fig. 13-49. It will be noted, even at 50,000 Hz and at diameters of 4 inches, that there is a slight increase in the output. As in the preceding measurements, the output level was either the same or slightly higher. However, an exception to this last statement occurred only at 4-inch diameter, where 50,000 Hz, 100,000 Hz and 15,000 Hz read plus with a cold stylus, while 10,000 Hz read plus with a hot stylus. All these measurements were repeatable.

The tests tabulated were made using a Westrex Corp. 3C stereophonic recording head, a Cappscoop 90-degree prewired stylus, and a Scully recording lathe. The discs were manufactured by the Audiotape Corp.

13.50 What effect does changing the diameter and groove velocity have on a 78.26 rpm record?—Very little, the losses at the smaller diameters are negligible compared to the same frequency at a diameter of 11.5 inches. (See Question 13.49.)

Recording-speed rpm	*Recording-blank diameter*	*Groove velocity inches per second*
33⅓	16-inch coarse pitch	10 to 25
33⅓	12-inch microgroove	8.5 to 20
45	7-inch microgroove	9 to 19
78.26	12-inch coarse pitch	16 to 48

Fig. 13-48. Average groove velocities for recording speeds.

Diameter	50,000 Hz	100,000 Hz
11.5 inch	Same hot and cold	Same hot and cold
8 inch	Same hot and cold	Same hot and cold
6 inch	Same hot and cold	Same hot and cold
4 inch	+ 0.2 dB cold	+ 0.1 dB cold
	10,000 Hz	15,000 Hz
11.5 inch	+ 0.2 dB cold	+ 0.9 dB hot
8 inch	+ 1.0 dB hot	+ 1.8 dB hot
6 inch	+ 2.5 dB hot	+ 2.0 dB hot
4 inch	+ 4.7 dB hot	+ 2.3 dB cold

Fig. 13-49. Effects of hot and cold stylus on frequency response.

13.51 What are the diameter losses for 45-rpm recordings?—See Question 13.49.

13.52 What is the formula for calculating linear velocity?

$$\text{Velocity} = \lambda \times f$$

where,

λ (lambda) is the wavelength in inches,
f is the frequency in hertz.

13.53 How are lacquer discs manufactured?—Basically, the disc consists of an aluminum base, which after visual inspection for defects and flatness, is cleansed and washed in a detergent solution. The disc is then passed through the coating machine, using a modified curtain-coating method. A carefully controlled film of lacquer is laid on the disc surface as it passes through the curtain. The disc is then dried by passing through a heated tunnel, followed by a cure for a specified period of time. The process is then repeated on the second face, followed by a controlled temperature cure. Finally, quality grading is done through 100-percent visual inspection of both sides. They are then packaged for shipment. Throughout the manufacturing process, white-room conditions must prevail.

The final quality of the disc is dependent on the extreme flatness of the aluminum substrate. The manufacturer must be concerned with minute pits or pimples on the aluminum surface, which result in defects of equal degree in the lacquer coating regardless of the thickness. In addition, during the coating process, the temperature and humidity must be closely controlled and the air supply filtered to remove 99 percent of particles of 0.30 microns or larger. The viscosity of the lacquer must be kept within 5 percent of the coating viscosity, and the temperature to within 2 degrees Fahrenheit, with the pressure of the lacquer entering the coating head absolutely constant.

13.54 What is an instantaneous recording?—A recording made on an acetate disc. See Question 13.53.

13.55 Define the term, "coarse-pitch recording."—Recordings made using 96 to 120 lines per inch are termed *coarse-pitch*, contrasted to *microgroove* recordings made using 250 to 400 lines per inch. To differentiate between these two recording techniques, dimension and equipment used in the first category are termed coarse-pitch and will be referred to as such throughout this work.

13.56 What is the cosmetic effect in a lacquer disc?—The aluminum used for making recording discs is stretched to obtain as near as flat a surface as possible, which causes a *cosmetic effect*. Since the lacquer is not completely opaque, certain cosmetic effects on the surface of the aluminum disc manifest themselves by two flashes per second as the disc revolves. To the recordist unaware of this effect, the disc may be rejected as unsuitable for recording. Discs of this nature are quite satisfactory for recording if they meet the other requirements for a good recording disc.

13.57 What are the recommended disc sizes for recording masters?—See Fig. 13-57. At least one *unmodulated groove* is to be recorded at both the start and end of the recorded area. The diameter of the innermost modulated groove is to be not less than 4¾ inches for 33⅓-rpm discs and not less than 4½ for 45-rpm discs. The center hole must be concentric with the recorded groove spiral within 0.005 inch and has been standardized as follows:

Nominal	Finished Disc (pressing or instantaneous)	Outer Modulated Groove Diameter (outside start)
12 inches	11⅞ ± 1/32 inches	11 7/16 inches maximum
10 inches	9⅞ ± 1/32 inches	9 7/16 inches maximum
7 inches	6⅞ ± 1/32 inches	6 11/16 inches maximum

Fig. 13-57. Recommended disc sizes for master recordings.

33⅓ rpm	0.286 inch + 0.001, − 0.002 inch diameter
45 rpm	1.504 inches ± 0.002-inches diameter

The warp of the disc, when measured using a surface indicator, is to be not greater than 1/16 inch and within any 45-degree segment not greater than 1/32 inch. Following the innermost recording groove, a leadout spiral and a concentric locking groove are provided.

13.58 What is the cause of surface noise using cold-stylus recording techniques?—The causes of surface noise in cold-stylus recording are many and varied. Among them are dull recording stylus, improper alignment, hard recording blank, and many of the effects noted in Questions 13.65 to 13.68. Surface noise may also be induced in the reproduction because of vertical pressure of the pickup, improper tracking angle, pickup stylus at an improper angle with respect to the disc surface, and pinch effect.

13.59 What causes noisy pressings?—Generally speaking, a pressing can be no quieter than the original recording; however, finger prints on the master and poor compounds used for pressing can cause a considerable amount of surface noise.

13.60 How does the signal to noise vary for hot- and cold-stylus recording?—Studies made by Jackson show the signal-to-noise ratios for a master disc in Fig. 13-60A. Older type discs will measure between 2 and 4 dB less signal-to-noise ratio.

A disc with a once-around, warp, bump, or unevenness produces a swinging-type noise, and the signal-to-noise ratio under these conditions may measure as given in Fig. 3-60B.

The measurements were made using a constant velocity of 5.5 cm/sec at 1000 Hz, using a Westrex 3C recording head, a Cappscoop prewired stylus, Scully recording lathe, and Audiotape Corp. recording disc. The signal-to-noise measurements as given in Figs. 13-60A and B apply to both 45 and 78.26-rpm recording, assuming a hot stylus is used. (See Question 13.49.)

13.61 Is it permissible to play back a master to be processed?—No. Masters are never played back for fear of damaging the sound track. As a rnle, two discs are cut simultaneously, a master and a playback for checking quality and cues. The master is checked by visual inspection only.

13.62 How does the hardness of an acetate disc affect the frequency response?—For cold-stylus techniques, the high-frequency response is increased at the expense of background noise. Generally speaking, all discs regardless of manufacture have the same frequency characteristics at the larger diameters. However, this is not true for smaller diameters, as high frequencies fall off at different rates, depending on the degree of hardness.

In addition, the situation is further complicated by the diminishing groove velocity at the smaller diameters. This may be partially compensated for by the use of a diameter equalizer or a hot stylus. (See Questions 13.111 and 13.153.)

Diameter	12-inch	10-inch	8-inch	6-inch	5-inch	4-inch
Hot stylus	−68.5 dB	−69 dB	−69.5 dB	−69.5 dB	−70 dB	−70 dB
Cold stylus	−67.0 dB	−66 dB	−63 dB	−60 dB	−57 dB	−53 dB

Fig. 13-60A. Signal-to-noise ratios for a master disc.

Diameter	12-inches	10-inches	8-inches	6-inches	4-inches
Hot stylus	−66/58 dB	−66/58 dB	−66/60 dB	−67 dB	−67 dB

Fig. 13-60B. Signal-to-noise ratios for a deformed master disc.

13.63 Is there an appreciable difference between recording discs of different manufacture?—There is very little difference among those of good manufacture. Experience indicates the frequency response of a particular disc is the combination of cutting-head frequency response, length of the recording stylus, diameter of the recording area, the electronics of the recording system, and if the recording is made using a cold or hot stylus. Using hot-stylus recording techniques, the resistance of the disc to the stylus is practically nonexistent with regard to frequency response. (See Question 13.49.)

13.64 What is white noise?—When the noise spectrum is uniform over a wide frequency band, it is referred to as white noise because it is analogous to white light. (See Question 1.140.)

13.65 What is orange peel?—A term applied to the surface of a recording blank which looks similar to orange peel. Such surfaces have high background noise.

13.66 What is a buzz?—A rasping noise heard in the background of a record and is generally caused by hum or noise in the system.

13.67 What is chatter?—A rattle in the reproduction caused by improper seating of the reproducing stylus in the sound track.

13.68 What is a chip?—The material removed from a record during the recording process.

13.69 What is an unmodulated groove?—A groove cut in a record without modulation. Such grooves are used at the start and ending of a recording and also to measure the signal-to-noise ratio.

13.70 What is an undulation?—A rising or falling appearance of the groove.

13.71 What is a dry cut?—A bad groove in a disc record. The chip is kinky and brittle because of a dull or improperly aligned recording stylus. It may also be caused by recording blanks which have become hardened with age. This effect is noticed when using cold-stylus recording techniques.

13.72 What is a cutover?—A breaking through the wall of one groove into the wall of the next groove. It is caused by overmodulation.

13.73 What causes a stylus to whistle when recording?—The stylus may be dull or out of alignment. Also, the disc may be old and dry. This effect is related to cold-stylus recording.

13.74 What causes a gray cut?—A cold stylus that is dull or a disc that is dry and hard.

13.75 What is the depth of cut?—The distance from the surface of the recording blank to the bottom of the groove. (See Question 13.150.)

13.76 What is a spiral out?—A groove at the inside end of a recording for the purpose of tripping an automatic record changer.

13.77 What is a locked groove?—A spiral groove at the inside end of a recording, closed by a concentric groove, for the purpose of tripping an automatic record reproducing machine. The locked groove provides a groove for the pickup stylus to ride in until the mechanism trips.

13.78 What is twinning?—When two grooves overlap or touch as a result of overmodulation. Twinning is also caused by improper groove-to-land ratios or a defective lead screw. (See Question 13.25.)

13.79 What is land?—The space between adjacent recording grooves.

13.80 What causes an echo effect in a blank groove?—A sound from an adjacent groove caused by overmodulation. Although the groove wall is not damaged, it is deformed by the excessive modulation in the adjacent groove. If the relationship is just right, with respect to the original modulation, what appears to be an echo is heard.

13.81 What is skating?—A condition existing when a reproducing stylus is pinched out of the groove and moves rapidly across the face of the record. Skating is also caused by the reproducing turntable not being level. It is particularly important, for microgroove reproduction, that the turntable be level in all directions.

13.82 What are horns in a sound track?—Raised edges on the sides of the groove (Fig. 13-82), caused by the

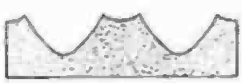


Fig. 13-82. Cross-sectional view of a sound track in a disc record showing the small horns raised at the upper edges by the recording stylus, caused by cold flow in the recording medium.

cutting stylus. The horns are small in size and very delicate; therefore, they are easily damaged by rubbing the surface of the record. The use of the hot stylus technique (see Question 13.153) eliminates this difficulty.

13.83 What is an advance ball?— A rounded sapphire support mounted on the cutting head, either beside or ahead of the stylus, and bearing on the surface of the record. Its purpose is to maintain a constant depth of cut regardless of irregularities in the surface of the record.

Advance balls were originated for soft-wax recording, before the advent of the acetate disc, and are still used in some instances for acetate-disc recording. However, with the use of automatic depth control and hot-stylus recording, advance balls are not used to the extent they were once used. If the recording disc has been stored for several hours at a temperature of 130-degrees Fahrenheit, it can be scored by the advance ball. In this instance, a hard deposit adheres to the advance ball, and if not removed, will score all succeeding discs. Lacquer discs should not be stored in a temperature exceeding 100-degrees Fahrenheit. (See Question 13.49.)

The placement of the advance ball in relation to the cutting-head stylus is shown in Fig. 14-21.

13.84 What is the formula for calculating the wave length of a recorded sound wave?

$$\text{Wavelength} = \frac{V}{F}$$

where,

V is the velocity of the record groove in inches per second at a given diameter,

F is the frequency of the recorded wavelength.

13.85 What is the standard signal-to-noise ratio for monophonic record reproduction? — For monophonic record reproduction the low-frequency noise voltage generated by the turntable, pickup, and equalizer, when playing an essentially rumble-free silent groove, shall be at *least* 40 dB below a reference level of 1.4 centimeters per second peak velocity at 100 Hz. (See Question 23.130.)

The frequency response of the pickup and its equalizers shall conform to the NAB standard reproducing curve, as shown in Fig. 13-95. The amplifier and indicating meter shall have uniform frequency response, within plus-minus 1 dB between 10 and 250 Hz, with the 500-Hz response 3 dB below the 100-Hz response, and an attenuation rate of at least 12 dB per octave at frequencies above 500 Hz (this may be obtained by means of a low-pass filter). The amplifier and indicating meter shall decrease at a rate of at least 6 dB per octave below 10 Hz.

The meter is to have the same dynamic characteristics as a standard VU meter USASI (ASA) C16-1961. (See Question 10.3.) If the meter fluctuates, the maximum values are used. In making these measurements, the pickup-arm resonance must fall outside the prescribed passband or be sufficiently damped to remove any possibility of affecting the measurement. The results are the electrical measurement, and not the aural annoyance value of low-frequency noise, such as rumble at frequencies below audibility. A strong low-frequency noise such as rumble, can cause severe intermodulation distortion with the low-frequency response of modern sound reproducing systems. The inaudible noise can be more serious than the indicated noise.

High-frequency noise is measured with the same meter characteristics when reproducing on a flat velocity basis, over a frequency range of 500 to 15,000 Hz, and shall be at least 55 dB below the level obtained under the same conditions of reproduction, using 1000 Hz recorded at a peak velocity of 7 centimeters per second. The frequency response at 500 Hz shall be 3 dB below the response at 1000 Hz, and shall fall off at a rate of at least 12 dB per octave below 500 Hz (using a high-pass filter). The response of the system at 15,000 Hz shall be 3 dB below the response at 1000 Hz and shall fall off at a rate of 12 dB per octave above 15,000 Hz.

13.86 What is the standard signal-to-noise ratio for stereophonic record reproduction? — For stereophonic record reproduction, the low-frequency noise voltage in each channel generated by the turntable, pickup, and equalizer—when playing an essentially rumble-free silent groove—shall be at least 35 dB below a reference level of 1 centimeter per second peak velocity at 100 Hz in the plane of modulation (vertical or lateral). The frequency response of the pickup and preamplifier shall conform to the NAB standard reproducing curve (Fig. 13-95), and the test equipment as stated in Question 13.85. The reference level of 1 centimeter per second peak velocity at 100 Hz corresponds in amplitude to 5 centimeters per second peak velocity at 500 Hz as it is in the constant-amplitude portion of the recording characteristic.

The high-frequency noise is measured with the same meter characteristics when reproducing on a flat velocity basis, over a frequency range of 500 to 15,000 Hz, and shall be at least 50 dB below the level obtained under the same conditions of reproduction, using 1000 Hz and fall-off rate of at least 12 dB per octave below 500 Hz. The response of the system at 15,000 Hz shall be 3 dB below the response at 1000 Hz and fall off at a rate of at least 12 dB per octave above 15,000 Hz.

13.87 What are the standard recording reference levels used for monophonic and stereophonic recording?—The NAB standard specifies that for monophonic microgroove recording, the recorded program level is to produce a reference level (using a standard VU meter) equivalent to a 1000-Hz tone, recorded at a peak velocity of 7 cm per second.

For stereophonic program material, the reference level is to be equivalent to a 1000-Hz tone, recorded at a peak velocity of 5 cm per second, as indicated on a standard VU meter. This is approximately 3 dB below the level for monophonic recording.

13.88 What percentage distortion is permissible for microgroove recording at 33⅓ rpm?—The distortion at 400 Hz should not exceed 1 percent total rms harmonic distortion. The intermodulation distortion should not exceed 3 percent using 100 and 7000 Hz mixed in a ratio of 4:1. Intermodulation measurements are discussed in Question 23.113.

13.89 What percentage distortion is permissible for microgroove recording at 45 rpm?—The distortion at 400 Hz should not exceed 1 percent total rms harmonic distortion. The intermodulation distortion using 100 and 7000 Hz should not exceed 4 percent mixed in a ratio of 4 to 1.

13.90 What are the recommended land-to-groove ratios?—For other than microgroove recording the ratio is set by the width of the groove, which is generally less than 4 mils and not greater than 5 mils, with a land-to-groove ratio of 60:40.

For 33⅓-rpm microgroove recording, the groove width is approximately 2.75 to 3.0 mils, with a groove-to-land width ratio of 70:30. For 45-rpm microgroove recording, it is the same. If variable-pitch recording is used, the groove-to-land ratio is constantly changing.

13.91 What is the appearance of a vertical-recorded groove? — Its appearance is as shown in Fig. 13-91. In vertical recording the depth of the cut and its width are constantly changing with the percentage of modulation. The sound track is engraved in the bottom of the groove. The unmodulated groove width is maintained at 0.003 inch for 125 to 150 lines per inch.

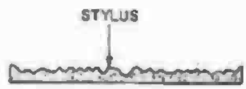


Fig. 13-91. Cross-sectional view of a vertical-cut sound track.

13.92 What is the appearance of a lateral-recorded groove?—It is as shown in Fig. 13-92. The groove depth and width remain almost constant with the grooves moving in a lateral direction or from side to side. The amount of lateral movement is dependent on the percentage of modulation of the audio signal.

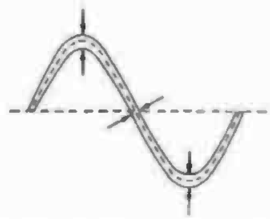


Fig. 13-92. Modulated lateral groove showing how the reproducing stylus is pinched by the narrowing of the groove. The dotted line indicates the center line of the groove.

13.93 Why was the NAB standard changed from a recording characteristic to a reproducing standard?—In the original Standard (1949 and 1953) only the recording characteristic was specified. However, it was found from experience that it was more practical to supply the manufacturer of record-reproducing

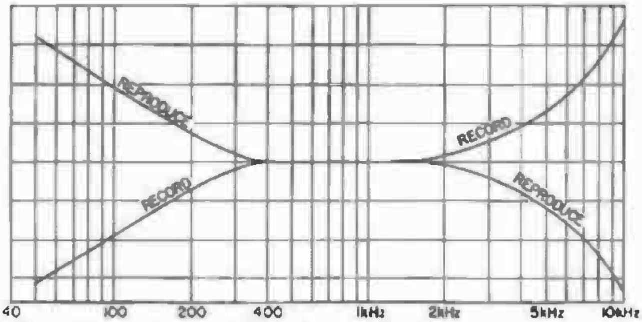


Fig. 13-94. Recording and reproducing characteristics for vertical records. This is the standard characteristic used in 1953, when such records were still in use. Each major vertical division is 5 dB.

equipment a test record with a standard reproducing characteristic than to specify the recording characteristic for the record manufacturer. Therefore, in the NAB Standard, March, 1964, only the reproducing characteristic is specified, with the understanding that the record manufacturer will adjust his recording characteristic to an inverse characteristic of the reproducing standard. On this basis, a reproducing standard was specified.

13.94 What is the recording and reproducing characteristic for vertical recording?—Although it is recognized that vertical recording is no longer used, the characteristics for both recording and reproducing are given in Fig. 13-94 for reference. It will be noted that the turnover frequency is 400 Hz.

13.95 What is the standard reproducing characteristic for disc recordings?—It is shown graphically in Fig. 13-95. This characteristic is known as the RIAA standard and was originally adopted by the disc recording industry in June, 1953, and was reaffirmed in March, 1964, by both the RIAA and NAB. It is used for the reproduction of 33⅓-, 45-, and 78.26-rpm recordings. The recording characteristic is an inverse of the reproducing characteristic, and it is used for both monophonic and stereophonic reproduction. This characteristic may be obtained by the use of three networks, consisting of an L/R parallel network having a time constant of 3180 microseconds, a series R/C network having a time constant of 318 microseconds, and an R/C parallel network of 75 microseconds.

13.96 What is the standard reference level for monophonic recording?—For monophonic recording the recorded program level shall produce the same deflection on a standard VU meter as that reproduced by a 1000-Hz tone recorded at a peak velocity of 7 centimeters per second. A margin of 10 dB is required between the sine-wave load handling capabilities of a system and the level of the program material mea-

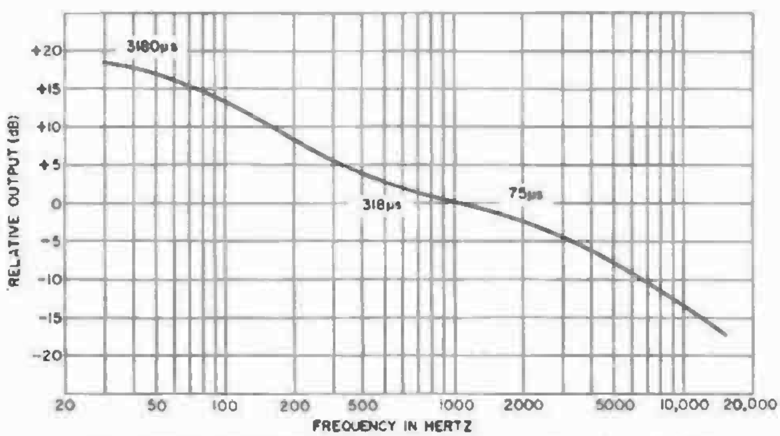


Fig. 13-95. NAB (RIAA) standard reproducing characteristic.

sured on a standard VU meter. (See Question 10.3.)

13.97 What is the standard reference level for stereophonic recording?—For stereophonic recording, the reference program level shall, for each channel in its plane of modulation, produce the same reference deflection on a standard VU meter as that produced by a 1000-Hz tone recorded at a peak velocity of 5 centimeters per second. This is approximately 3 dB below the level of the monophonic reference level.

Experience indicates that at least a 10-dB margin is required between the sine-wave load handling capabilities of a system and the level of the program material, measured on a standard VU meter. This would then contemplate program material running as high as a velocity of 21 centimeters per second and is believed to be the maximum velocity that can be traced without excessive distortion at the smaller diameters of a disc turning at 33⅓ rpm.

13.98 What is a minigroove recording?—A 78.26-rpm recording made using microgroove techniques, including hot-stylus and variable-pitch recording. Although the number of lines per inch is greater than is normally used for 78-rpm recording, it is not enough so that the record may be called long playing, or extended play. The frequency characteristic used is that of Fig. 13-95, known as the RIAA recording characteristic.

13.99 What is an orthocoustic recording?—It is a recording characteristic introduced by RCA Victor several years ago for recording 16-inch transcriptions at a speed of 33⅓ rpm. This characteristic, shown in Fig. 13-99, had been used by the broadcasting and recording industries for many years until the advent of the RIAA recording characteristic.

The orthocoustic characteristic makes use of pre-equalization in the recording circuits and post-equalization in the reproducing circuits as described in Question 6.11. The principal advantage gained by the use of such equalizers is the increased signal-to-noise ratio above 2000 Hz.

13.100 What is an orthophonic recording characteristic?—A reproducing characteristic introduced by RCA Victor some years ago; it is similar to the RIAA characteristic. The orthophonic recording and reproducing characteristic employs both pre- and post-equalization, discussed in Question 6.11.

13.101 Define constant-amplitude recording.—Constant-amplitude recording indicates a mechanical recording characteristic for a fixed amplitude of a sine-wave signal; the resulting recorded amplitude is independent of frequency. This subject is discussed in Question 14.5. As a rule, this method of recording is used with nonprofessional recording equipment employing a crystal cutting head, as it requires no equalization in either the recording or reproducing system.

13.102 Show the original Audio Engineering Society reproducing characteristic.—This characteristic is shown in Fig. 13-102. The original charactreistic has now been modified and corresponds to the RIAA curve shown in Fig. 13-95.

13.103 What is the RIAA reproducing characteristic? — A reproducing characteristic adopted by the Record Industry Association of America as a standard for reproduction of disc records. It is a modification of the original

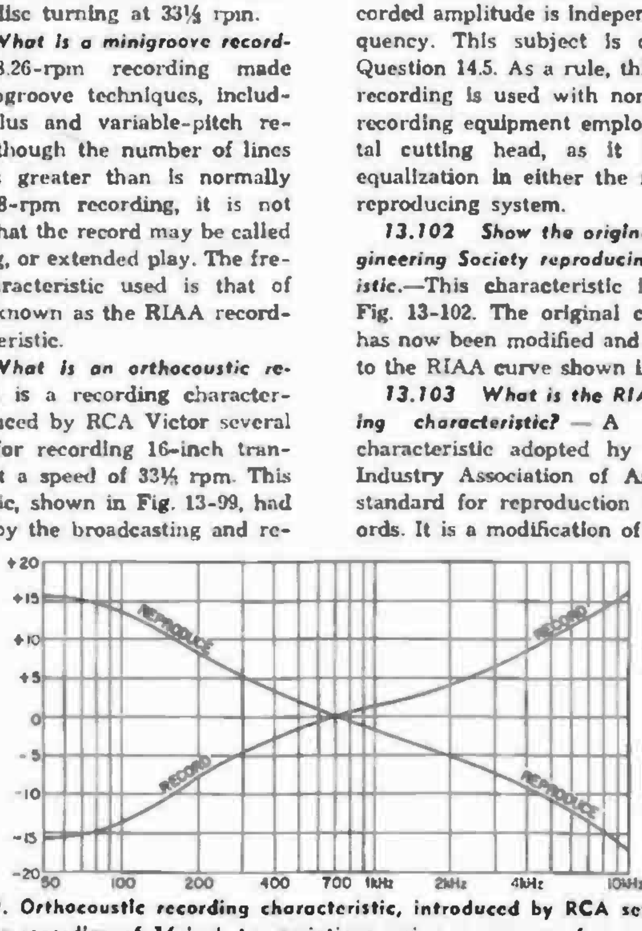


Fig. 13-99. Orthocoustic recording characteristic, introduced by RCA several years ago for the recording of 16-inch transcriptions, using a crossover frequency of 700 Hz; now superseded by the RIAA-NAB standard.

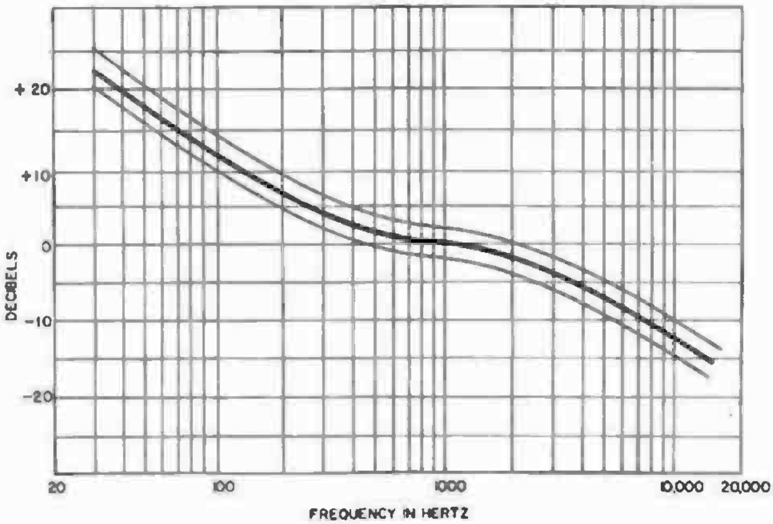


Fig. 13-102. The original AES reproducing characteristic, now superseded by the RIAA characteristic.

reproducing characteristic recommended by the Audio Engineering Society (Fig. 13-102). The present standard is shown in Fig. 13-95.

13.104 What is the standard groove shape for a monophonic pressing?—The groove shape shall have an included angle of 90 degrees, plus-minus 5 degrees, with a top width of not less than 0.0022 inch. For these groove dimensions, it is recommended that the reproducing stylus have a tip radius of 0.001 inch, plus 0.0001 inch, minus 0.0002 inch, with an included angle of 40 to 50 degrees. (See Question 13.180.)

13.105 What is a flat recording characteristic?—One which uses no pre-equalization in the recording circuits. Equalization is used in the reproducing equipment to compensate for the constant-amplitude constant-velocity characteristics of the recording head. This method of recording is seldom used, except in home-recording systems, because of the low signal-to-noise ratio above 2000 Hz.

13.106 What is the purpose of a low-frequency equalizer in a reproducer circuit?—To compensate for the constant-amplitude characteristic of the recording head below the turnover frequency. The equalizer has an inverse-frequency characteristic to the recording characteristic.

13.107 What is the turnover frequency?—The frequency at which a cutting head changes from a constant-velocity characteristic to a constant-amplitude characteristic. This is illustrated in Fig. 14-9.

13.108 What is the purpose of recording disc masters at half-speed?—Because of the high levels recorded on present-day disc records, certain recording activities in both the United States and Europe are recording their masters at half the normal recording lathe speeds. Reduced speeds decrease

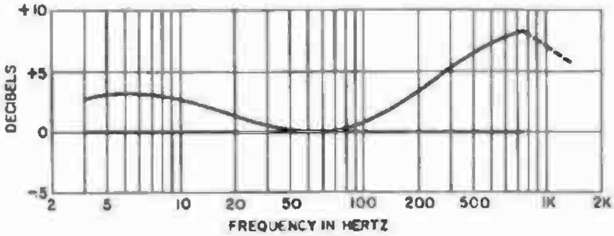


Fig. 13-108. Equalization characteristic for transferring tapes recorded at 15 ips and reproduced at 7.5 ips for recording disc masters at half-speed.

the load on the cutting-head stylus and reduce the elastic spring-back characteristics of the acetate recording blank, thus reducing the distortion. The result of recording at half-speed (or one octave lower) is the cutting-head system response extends to only one-half the normal frequency range, or about 7500 Hz. The additional octave is retrieved by adjusting the equalization in the negative-feedback circuit of the cutting-head amplifier circuit.

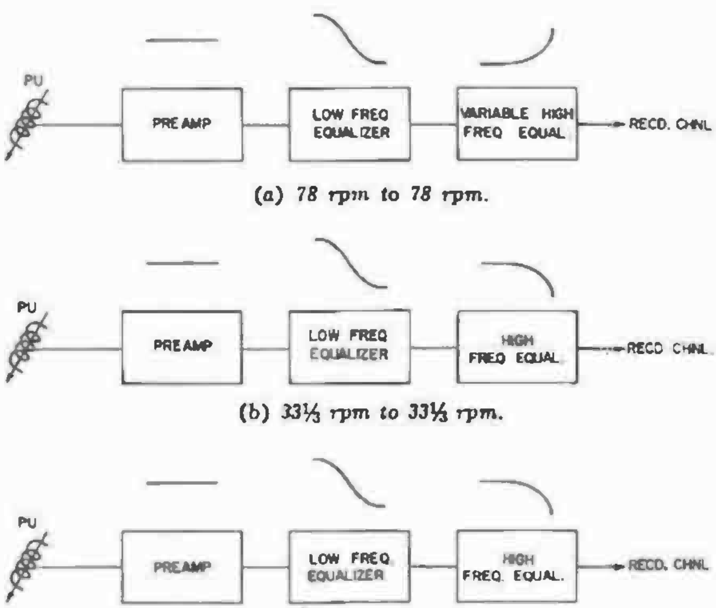
Reducing the recording lathe speed by one-half shifts the pre-equalization characteristics by one octave. To restore the final recording to the RIAA standard special equalization is required between the master tape and the disc-recording amplifier, so that when the disc is played back at its normal speed, it has the RIAA standard characteristics. To accomplish this, the following conditions must prevail:

> Tape recorder equalized for NAB standard characteristic at both 7.5 and 15 ips.
>
> Disc-recording system equalized for RIAA recording characteristic.
>
> Tape originally recorded at 15 ips, using standard NAB characteristic, played back at 7.5 ips.
>
> The recording lathe is rotated at 16⅔ rpm for recording 33⅓-rpm masters, or 22½ for 45-rpm masters, and 39.13 for 78.26 rpm.

Although this method of recording is generally confined to 45-rpm recordings, it may be used successfully with any system if due consideration is given to the intervening equalization.

Pressings made from half-speed recordings are played back at their normal speed of 33⅓ rpm or 45 rpm. A typical equalization curve suitable for connection between the master tape machine and the disc-recording channel is given in Fig. 13-108. (See Question 17.228.)

13.109 What are the methods used for recording from one characteristic to another?—They are as shown in Fig. 13-109. At (a) the pickup is shown feeding a preamplifier with a fixed low-frequency equalizer and variable high-frequency equalizer at its output. This circuit may be used to rerecord one 78-rpm recording to another 78-rpm recording. The low-frequency equalizer is set for the proper low-frequency compensation and the high-frequency equalization is used to boost or attenuate the high frequencies as required. If possible, the high frequencies should



(a) 78 rpm to 78 rpm.

(b) 33⅓ rpm to 33⅓ rpm.

(c) 33⅓ rpm to 78 rpm.

Fig. 13-109. Various methods used for transferring disc records of one characteristic to another.

be increased from 4 to 6 dB, surface noise permitting.

A 33⅓-rpm recording being transferred to a similar type recording is shown at (b). The regular equalization used for reproducing such records is used. The recording equalization is that which is regularly used for such type recordings.

A 33⅓-rpm recording being transferred to a 78-rpm disc is shown at (c). Again, the regular reproducing equalizers are used; however, the equalization in the recording circuits is that required for normally recording 78-rpm records.

An important point to remember is that the original recording is reproduced in a normal manner, except for a slight increase of high frequencies to compensate for transfer losses. The recording circuits are normal for the type recording being made.

13.110 What are the actual values in decibels of equalization used for the RIAA standard reproducing characteristic?—The actual values in decibels are tabulated below and are values from which the curve of Fig. 13-95 is plotted. The recording characteristic is the *inverse* of this curve. The original standard specified the frequency response of the recording system, but later it was found to be more practical to specify the reproducing characteristic. (See Question 13.93.) The characteristics of this curve may be obtained by the use of three networks having 3180, 318, and 75 microseconds each.

Hz	*dB*
30	−18.61
50	−16.96
70	−15.31
100	−13.11
200	−8.22
300	−5.53
400	−3.81
700	−1.23
1,000	0.00*
2,000	+2.61
3,000	+4.76
4,000	+6.64
5,000	+8.23
6,000	+9.62
7,000	+10.85
8,000	+11.91
9,000	+12.88
10,000	+13.75
11,000	+14.55
12,000	+15.28
13,000	+15.95
14,000	+16.64
15,000	+17.17

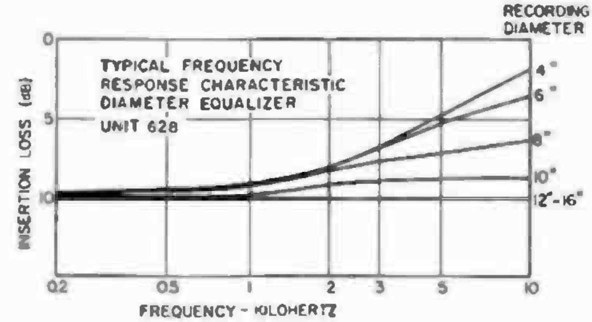
*Reference frequency.

13.111 What is diameter equalization?—The use of a variable equalizer in the recording circuit to compensate for the loss of high frequencies at the smaller groove diameters caused by the decreasing groove velocity. The equalizer consists of a bridged-T network of constant insertion loss as described in Question 6.15. The shaft of the equalizer attenuator is mechanically connected to the cutting-head carriage on the recorder as shown at R in Fig. 13-2. The loss of the attenuator is controlled by the position of the cutting-head carriage relative to the diameter of the recording surface. Thus, the correct amount of equalization is inserted for a particular diameter.

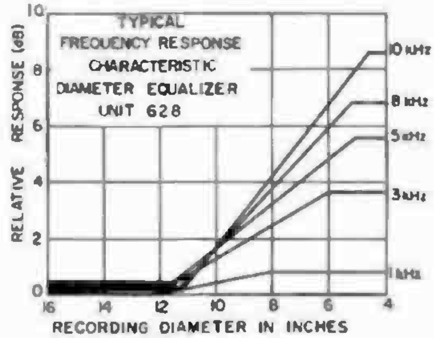
Generally, for 16-inch transcriptions, the equalization begins to take effect at a diameter of 12 inches and continues in steps of 0.5 dB for a total of 6 to 8 dB to a diameter of 5 inches. The resonant frequency of the equalizer depends on the bandwidth of the recording. Diameter equalization is not practical at diameters less than 6.5 inches, as the intermodulation and tracking problems in the finished record increase quite rapidly below 7 inches.

A typical diameter equalizer-attenuator, with its mounting bracket and control pulley, is shown in Fig. 13-111A. A spiral spring under the pulley returns the attenuator to its normal posi-

Fig. 13-111A. A diameter equalizer-attenuator control manufactured by Cinema Engineering Co. The cord is attached to the cutting-head carriage.



(a) *Insertion loss versus frequency.*



(b) *Recording diameter versus the relative response at various frequencies.*

Fig. 13-111B. Diameter equalizer characteristics.

tion when the recording-head carriage is returned to the outside diameters. The network elements are mounted in a nested shielded case separate from the attenuator pot. Diameter equalization should not be confused with pre- and post-equalization as they are two separate networks. Typical frequency characteristics for diameter equalization are shown in Fig. 13-111B.

With the advent of hot-stylus recording techniques, diameter equalization is not always used. However, with cold-stylus recording it is essential. In using diameter equalization, one important factor must be borne in mind. As the high-frequency equalization is increased, so are the high-frequency distortion products. Therefore, the distortion in the amplifier system must be kept to a low value.

13.112 Where is a diameter equalizer connected in a recording channel?—A typical monophonic recording channel is shown in Fig. 13-112, with both pre- and diameter equalization, and the

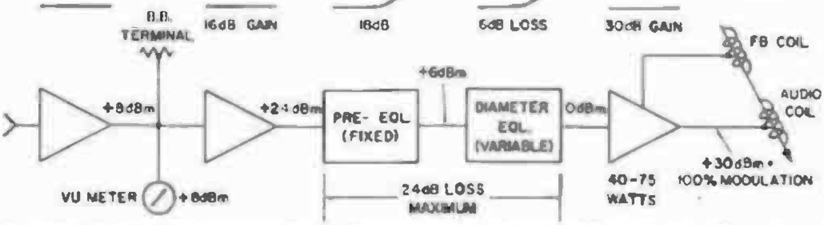


Fig. 13-112. Block diagram for a disc-recording channel showing the placement of pre- and diameter equalization with the losses and gain for an average recording system. The levels shown are in reference to 400-Hz or a lower frequency, depending on the design of the equalizers.

approximate operating levels. The signal from the mixer console or tape machine is applied to a line amplifier which, in turn, drives a bridging bus. From the bus, the signal goes through a bridging amplifier to a pre-equalizer having such characteristics that the playback characteristic conforms to the industry standard (Fig. 13-95).

From this equalizer, the signal passes to the diameter equalizer, which is variable, and automatically operated from a mechanical connection on the recording lathe. (See Question 13.111.) This is followed by a power amplifier of 40- to 75-watts output for driving the cutting head. Although the circuit diagram indicates a negative-feedback cutting head, the same circuitry is used with a nonfeedback head.

The approximate gain and frequency characteristic is shown above the various components. In lining up such a channel for recording, either 400 Hz or some other frequency may be used, depending on the frequency characteristics of the equalizers. Because of the characteristics of the equalizers, the exact same frequency must be used each time the channel is lined up. This is accomplished by applying, for example, 400-Hz signal to the line amplifier and setting the bus level to plus 8 dBm, and making a test cut which should measure 100-percent modulation of the cutting head.

The components shown in the diagram are not of any given manufacture or design but are only representative of typical characteristics and placement in the circuit. A complete magnetic tape transfer channel to disc is discussed in Question 17.228.

13.113 What power is recommended for the bridging amplifier shown in Fig. 13-112?—It should be capable of producing one watt or more of power to prevent the distortion from becoming excessive when the full equalization is being used.

13.114 What output power is recommended for driving a cutting head?—The minimum for satisfactory results is 40 watts; 75 is desirable, and 100 to 150 watts is not uncommon. The amplifier output should have a damping factor of at least 20, and higher if possible. The harmonic and intermodulation distortion should not exceed 1 percent at full power output.

13.115 What is the formula for calculating the playing time of a disc record?

$$T = \frac{NS}{rpm}$$

where,

T is the playing time in minutes,
S is the recorded width in inches,
N is the number of lines per inch,
rpm is the turntable speed in revolutions per minute.

13.116 What is a stroboscopic disc?—A circular disc containing a number of black and white bars, which is used for checking the speed of turntables and other rotating machines. (See Fig. 13-116.) The disc is placed on the turntable and the bars observed under a light source fed from the normal ac lighting circuits. When the speed of the turntable is correct, the black bars will appear to stand still. If the table is turning too fast, the bars speed up and drift in the direction of rotation. When running slow, the reverse takes place. Stroboscopic bars may be painted around the rim of a turntable and a 115-volt neon light mounted close by for constant observation.

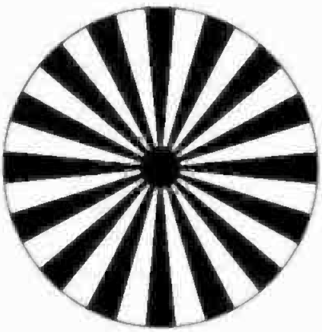


Fig. 13-116. *A stroboscope disc used for checking the rotational speed of a turntable.*

13.117 What is the formula for calculating the bars on a stroboscopic disc?

$$\text{Bars} = \frac{F \times 2 \times 60}{rpm}$$

where,

F is the frequency of the light used to observe the bars,
rpm is the speed of the turntable in revolutions per minute.

The number of bars required in a stroboscopic disc using 60-Hz lighting current is:

rpm	*bars*
16	450
33⅓	216
45	159
78.26	92

The design of stroboscopic discs for use on projectors and cameras is discussed in Question 3.87.

13.118 What is a light pattern?—A test record recorded to measure the frequency response of a disc-recording channel. Frequencies of constant amplitude are sent into the channel and recorded, using the normally employed equalizers. The finished record is then viewed under a light held parallel to the surface of the record. The pattern seen is the frequency response of the recording channel including all components and the cutting-head characteristics. The subject is further discussed in Question 23.75.

13.119 What is a Christmas tree pattern?—A light pattern similar to that described in Question 13.118.

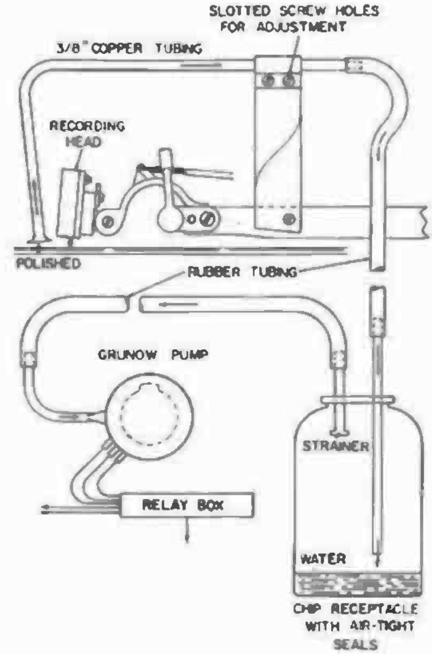


Fig. 13-120. A typical vacuum system for removing the chip from a disc record while recording. (Courtesy, Audio Devices Inc.)

13.120 What is a vacuum system and its purpose?—A suction system used for removing the chip from a disc record as the recording head moves across the face of the disc. A typical vacuum system is shown in Fig. 13-120.

13.121 How is a recording blank held by vacuum to a recording turntable?—The vacuum system used for removing the chip from the record while recording is connected to a group of holes in the periphery of the turntable, thus holding the blank in place. (See Question 13.2.)

13.122 What does the term "cut" mean?—To stop or cut off. It is also used to designate a sound track in a disc record.

13.123 What is a copper master?—A thin sheet of copper which is electroplated onto the surface of an acetate or soft wax master disc. When removed from the master, it is a negative of the original.

13.124 What is a nickel master?—A nickel-plated copper master; also a negative.

13.125 What is silver spraying?—It is a method of metallizing the lacquer master with silver, using a dual spray nozzle, wherein ammoniated silver nitrate and a reducer are combined in an atomized spray to precipitate the metallic silver.

13.126 What is gold sputtering?—A method of gold plating a master disc record by the use of a water-cooled vacuum chamber. A metal plug is inserted in the center hole of the record and connected to a source of positive dc high voltage. A gold button connected to the negative high voltage is placed near the center of the record. The high voltage causes the molecules of the gold button to pass to the surface of the disc, depositing a layer of gold approximately one molecule in thickness on the disc. This method is used in place of electroplating. This method is now obsolete.

13.127 What is a stamper?—The metal negative of the original master used in a hydraulic press for stamping the finished record. A typical record press is shown in Fig. 13-127. Hydraulic pressure combined with heat supplied by live steam passing through the record press molds the plastic biscuit into a record. Cold water is circulated through the blocks in the press for a few moments before releasing the pressure and removing the record.

13.128 What is a matrix?—A stamper. The image is negative.

Fig. 13-127. Disc record press as used in the RCA Victor record pressing plant at Camden, N.J.

13.129 What is a mother?—A nickel-plated positive of the original record used for making stampers.

13.130 What is a pressing?—A commercial phonograph record. It may be formed with or without heat.

13.131 What is a metal master?—A copper master. (See Question 13.123.)

13.132 What is a back plate?—A metal or wood backing to which a copper or negative master is attached to facilitate handling.

13.133 What is a master recording?—The original recording which may be a disc, film, or a magnetic tape.

13.134 What is a platen?—A flat circular plate used in the hydraulic press to hold the stamper.

13.135 What is a reverse copy?—A metal copy of the nickel master made by means of electroplating. It is an exact duplicate of the original recording. After mounting on a back plate it is called a master.

13.136 What is a shellac pressing?—A commercial record made some years ago, and so called because the compound contained shellac, which provided a hard surface; now obsolete.

13.137 What is a binder?—A substance used to bind the basic materials in a processed record together.

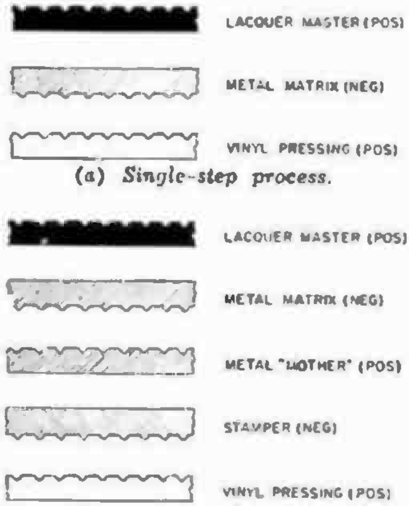
13.138 What is a filler?—A substance added to the basic material used in processed records to provide color and weight.

13.139 What is a biscuit?—A dough-like material used in the manufacture of phonograph records. (See Fig. 13-127.)

13.140 What is a wax record?—A recording blank made of soft wax and composed of soap, styric acid, and other ingredients. It is used for recording the original master and is processed in a manner similar to any other master. Soft wax was the original method used for making disc records; however, it is now obsolete, having been replaced by the acetate disc master.

13.141 What is vinylite?—A plastic material used in the manufacture of phonograph records because of its low surface noise and its resistance to breakage.

13.142 Describe the procedures for processing disc records.—Two methods may be used—the single-step and the three-step process—as illustrated in Fig. 13-142. The single-step method is used when less than 200 pressings are to be made and consists of a lacquer master, metal matrix, and a vinylite pressing. The three-step method is used



(a) *Single-step process.*

(b) *Three-step process.*

Fig. 13-142. Record-processing procedures. (Courtesy, Audio Devices Inc.)

when more than 200 pressings are required and the master must be reused.

13.143 What method is recommended for storing disc records?—They should be stored edgewise.

13.144 How should masters be packed for shipment?—They should be packed as shown in Fig. 13-144.

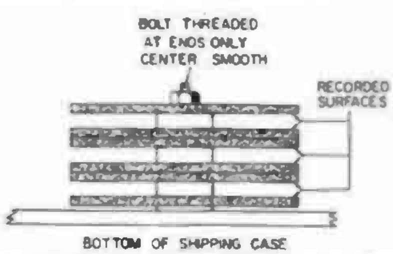


Fig. 13-144. The suggested method for shipping masters. (Courtesy, Audio Devices Inc.)

13.145 What is an air check?—A recording made from a radio broadcast by means of a radio receiver.

13.146 How should recording blanks be handled?—They should be handled by the edges only. Never permit the hands to come in contact with the recording surfaces as the oil from the fingers causes noisy spots in the processed disc.

13.147 What precautions should be taken when recording a master to be processed?—It should be visually inspected for defects in the surface before recording. After recording, it should again be inspected with a magnifying glass for breakthroughs in the walls of the sound track. A playback record cut at the same time the master was recorded is used for checking for quality and cues. Discs to be processed are never played back, as damage to the sound track may result.

13.148 Describe a compatible recording.—In the reproduction of disc records, if a stereophonic recording is played on a reproducer employing a lateral or monophonic pickup, the sound track modulations will be damaged because of the nonvertical compliance of the pickup motor mechanism. It has long been the aim of recording engineers to produce a sound track that would play equally well usiug either a monophonic or stereophonic pickup. Experiments seemed to indicate that it was the low-frequency vertical signal that caused most of the wear. If the low-frequency signal was removed from the vertical portion of the sound track, the low frequencies would appear in the lateral component which represents the sum of the two channels, because the vertical information represents the stereo information equal to the difference between the two channels. Recordings made in this manner show considerably less wear when reproduced using a lateral pickup. However, further study indicates that stereo information is carried by the low frequencies and leads to the degradation of the quality of reproduction.

Experience indicates that at the present time *stereophonic* recordings should not be played back using a *monophonic* pickup. However, *monophonic* recording can be reproduced quite satisfactorily using a *stereophonic pickup.*

13.149 How may acetate discs be cleaned?—With a weak solution of soap in cool water.

13.150 How may the sound-track groove depth be calculated?

For an included angle of 87 degrees:

$$\text{Groove depth} = 0.5269 \times A - 0.4527 \times R$$

For an included angle of 70 degrees:

$$\text{Groove depth} = 0.7141 \times A - 0.7434 \times R$$

where,

A is the groove width,
R is the stylus tip radius.

The groove width is measured with a calibrated microscope.

13.151 How much does the RIAA high-frequency pre-equalization improve the signal-to-noise ratio?—It will improve the signal-to-noise ratio by 8 dB, resulting in an effective signal-to-noise ratio of 58 dB under minimum conditions. (See Question 13.95.)

13.152 What is a negative-feedback cutting head?—This device is discussed in Section 14.

13.153 What is hot-stylus recording?—The use of a heated stylus during the recording process. This method reduces the pressure on the recording stylus and improves the high-frequency response as well as the signal-to-noise ratio. This subject is discussed in length in Questions 15.60 to 15.71.

13.154 How does turntable rumble affect low-frequency reproduction?—Turntable rumble has a high energy content and causes the cone of the loudspeaker to operate in the nonlinear portion of the speaker characteristic, distorting the low frequencies. If the rumble is great enough, the reproduction sounds as though the cone were breaking up.

13.155 Show how a given waveform is affected by a change in the groove velocity.—A given waveform recorded at a diameter of 12.5 inches and at 5.5 inches is shown in Fig. 13-155. It will be noted the waveform at 5.5 inches is cramped into a space less than half that at 12.5 inches. This complicates tracking and increases the noise and distortion.

13.156 What is the minimum separation specified for stereophonic reproduction?—The separation between recorded and unrecorded channels measured at the output of the reproducer with its equalizer and preamplifier must be at least 26 dB over the range of 100 to 7500 Hz, and the separation shall not fall off at a greater rate than 6 dB per octave. As the values specified may be subject to noise and erratic measurement, it is suggested, if possible, that a tuned voltmeter be used for such measurements.

13.157 What is the standard for balance between stereophonic channels?—Playing back a record, with the output of each channel adjusted to within 0.25 dB at 1000 Hz, the frequency response of each channel shall agree with the standard reproducing curve of Fig. 13-95, within plus-minus 1 dB between 100 Hz and 7500 Hz, and within plus-minus 2 dB above and below these frequencies. Equal modulation for the two sides may be obtained by using a monophonic lateral test record.

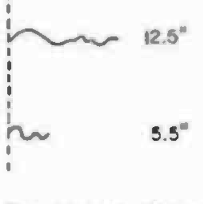
13.158 What is the standard for phasing a stereophonic recording channel?—For recording, the standard states that equal in-phase signals applied to the left and right channel inputs of a stereo disc recorder shall result in lateral modulation of the stereo groove. Conversely, equal antiphase signals produce vertical modulation. For reproducing, the lateral modulation of the groove will produce equal in-phase voltages at the output of the pickup, and, conversely, that vertical modulation will produce equal antiphase voltages. (See Question 13.202.)

13.159 How does the distortion vary for lateral recording?—It varies as the square of the signal amplitude, as the square of the stylus radius, as the square of the frequency, and, inversely as the groove velocity.

13.160 How does the distortion vary for a vertical recording?—It varies directly as the signal amplitude, directly as the frequency, directly as the stylus radius, and directly as the groove velocity.

13.161 Define groove angle.—It is the angle between the two side walls of a groove, measured in a radial plane perpendicular to the disc surface.

13.162 Show the schematic diagram for a cutting-head filter or preequalizer.—A recording filter or equalizer for recording, using the RIAA recording



(a) Waveform recorded at a high groove velocity.

(b) Same waveform recorded at a low groove velocity.

Fig. 13-155. The effect of groove velocity on recorded waveforms.

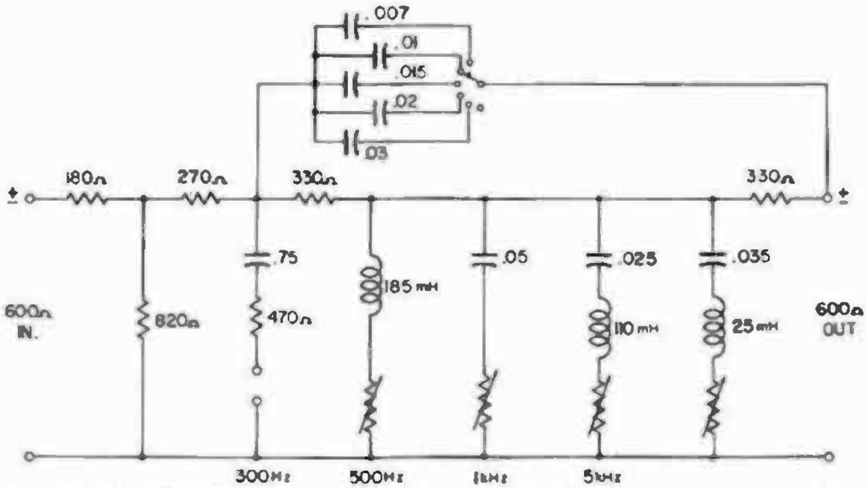


Fig. 13-162A. Recording pre-equalizer for adjusting the frequency response of a cutting head.

characteristic, developed by H. E. Roys of RCA is shown in Fig. 13-162A.

The circuit consists of an input pad, and five adjustable circuits for smoothing out the frequency response of the cutting head. An ideal cutting-head recording characteristic and a practical response curve are shown in Fig. 13-162B, with the RIAA recording characteristic. The practical response is attained by the use of the recording filter shown in Fig. 13-162A. The small peaks and valleys in the final cutting-head response are quite small and may be considered to be flat for all practical purposes. The high-frequency tilt-up of the RIAA characteristic is obtained by the adjustment of the capacitors in parallel with the upper portion of the filter network. The frequency response of the cutting head is adjusted by means of the adjustable circuits in the filter network for the smoothest frequency response by recording a series of light

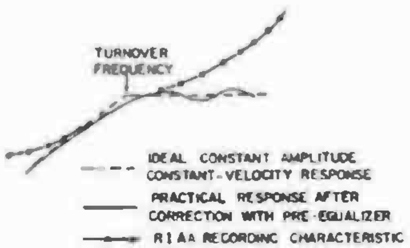


Fig. 13-162B. The ideal cutting head recording characteristic and a practical response curve. Also shown is the RIAA recording characteristic.

patterns or by the use of a cutting-head calibrator, discussed in Question 14.49. (See Question 23.75.)

13.163 How is the recording filter of Fig. 13-162 connected in the recording circuit?—The recording filter or equalizer may be connected in the recording channel in several different ways. The network is connected ahead of the recording amplifier at part (a) in Fig. 13-163A. At part (b) it is connected in tandem with a variable diameter equalizer.

Three disc recorders driven from a bridging bus are shown in Fig. 13-163B. In this instance, the recording filters are designed with a 10,000-ohm bridging input transformer connected across a 600-ohm bus. Separate recording amplifiers are used for each recorder. In this type arrangement a diameter equalizer cannot be used.

13.164 How is 100-percent modulation of a cutting head determined?—It may be acheved by two methods: by having a VU meter connected across the head circuit or by recording an unmodulated groove and measuring its width, using a calibrated microscope. The latter method is preferred, as it takes into consideration the characteristics of the disc material, stylus heat (if used), and any other factors associated with the actual displacement of the cutting-head stylus. The gain control in the recording amplifier should have ½-dB steps to permit accurate adjustment of the 100-percent modulation. The frequency used for lining-up the

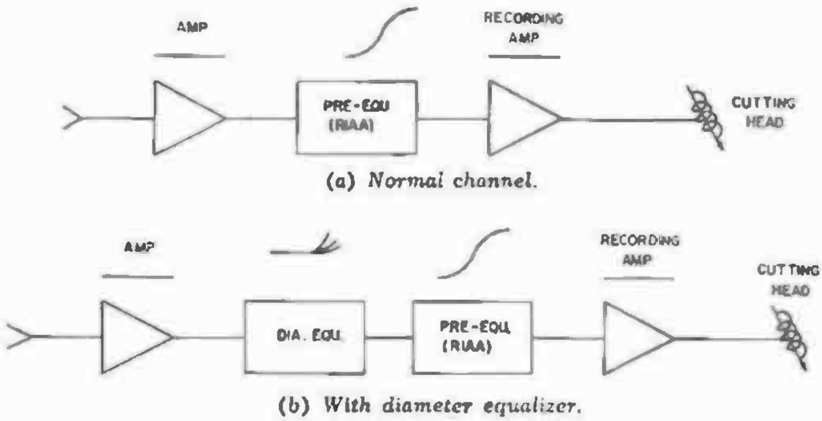


Fig. 13-163A. Position of the recording pre-equalizer in the recording channel.

channel before recording should be at a predetermined frequency, taking into consideration the equalizer characteristics. (See Question 13.112.)

13.165 Define translation loss (playback).—It is the loss suffered in the reproduction of a pressing, whereby the amplitude of the reproducer stylus differs from the recorded amplitudes on the record.

13.166 What is the recommended reproducing characteristic for 16-inch 33⅓-rpm recording?—The same frequency response as for 33⅓-rpm microgroove recording (Fig. 13-95). As this type recording uses coarse pitch, the recording stylus uses an included angle of 88 degrees, plus-minus 5 degrees, with a tip radius of 1.5 mils.

13.167 What is the recommended reproducer characteristic for 78.26-rpm recording?—The same frequency response as for 33⅓-rpm microgroove recording (Fig. 13-95). Because this type recording uses a coarse pitch, the recording stylus uses an included angle of 87 degrees, with a tip radius of 1.5 mils.

13.168 What is the recommended reproducing characteristic for 16⅔-rpm recording? — The frequency response follows the same general characteristic as for the microgroove. However, it may require some correction to obtain a satisfactory finished product. Such records are recorded using microgroove techniques. Recordings such as these are used for books for the blind and other extremely long playing records. (See Question 13.22.)

13.169 What does the abbreviation "RIAA" mean?—It is the initials of the Record Industry Association of America. It is also used to designate the reproducing characteristic adopted by all

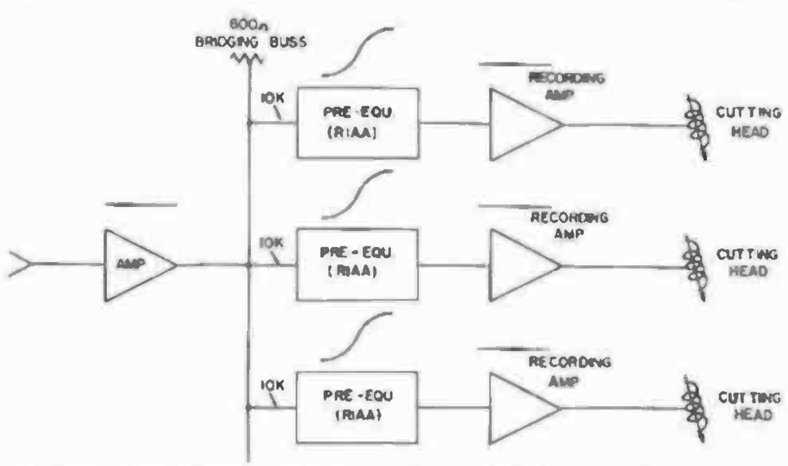


Fig. 13-163B. Three disc recorders driven from a bridging bus. The pre-equalizers have 10,000-ohm bridging input impedance.

the leading recording companies as a standard for record reproduction. It has also been adopted by many of the record manufacturers of Europe. The standard was adopted in June, 1953, and reaffirmed in March, 1964. This characteristic is discussed in Question 13.95, and the actual values of equalization in Question 13.110.

13.170 What is pinch effect?—See Question 16.27.

13.171 Define tracking error—It is the *inability* of the reproducing pickup stylus to properly follow the recorded groove, thus inducing distortion and noise with possible damage to the groove modulations. *Trackability* is now the generally accepted term, rather than tracking error. (See Questions 13.172 to 13.174, and 16.42.)

13.172 What are the most common causes of poor tracking in a reproducer?—Some of the most common causes are pinch effect, groove too shallow or too light, turntable not level, bent recording blank, overmodulation of the sound track, and side walls of the groove broken through.

13.173 What is tracing distortion?—Nonlinear distortion in reproduction of a disc record. Distortion is created because the curve traced by the playback stylus is not an exact replica of the modulated groove. Tracing distortion is caused by the stylus tip which is round and of finite radius. If the wavelengths of the modulations are of the same dimensions as the stylus tip, difficulty will be experienced when attempting to reproduce the sound track.

The effects of tracing distortion become greater as the smaller diameters are approached because of the lower groove velocities. Tracing distortion is created because the playback stylus and the cutting head stylus contact do not occur at exactly the same point. In tracing a groove, the playback stylus contacts the groove wall at a point different from that made by the recording stylus. This difference of position causes distortion in the reproduction. With the advent of the biradial or elliptical reproducing stylus, the effect of tracing distortion has been considerably reduced. (See Questions 13.30 and 13.32.)

13.174 What is tangent error?—When a laterally recorded disc record is recorded, the cutting head is carried across the face of the recording disc at right angles to the direction of the disc motion. However, when reproduced, the pickup is never at right angles to the direction of motion, except at one point, because the pickup arm is pivoted in such a manner that it swings across the face of the disc in an arc as shown in Fig. 13-174. Point A is the only place where the stylus point is at right angles to the direction of motion.

The constant change of angle by the playback stylus causes a strain on the sidewalls of the record groove, which eventually tears out the high frequency modulations, leading to increased noise and distortion. (See Question 16.52.)

13.175 Define constant-velocity recording.—Constant-velocity recording indicates a mechanical recording characteristic wherein, for a fixed amplitude sine wave, the resulting recorded amplitnde is inversely proportional to frequency.

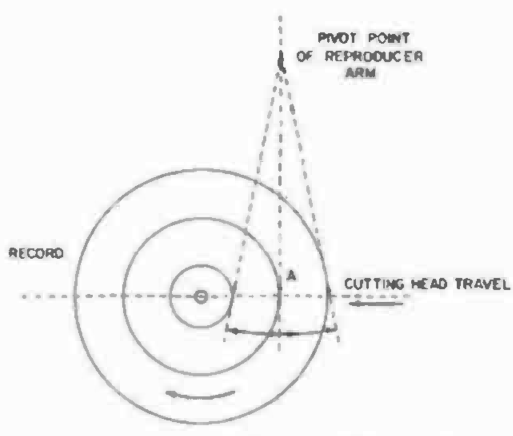


Fig. 13-174. Tangent error in a reproducing arm. The error is zero at point A only.

13.176 What is a sweep-frequency record?—A test record on which have been recorded a series of constant-amplitude frequencies. Each frequency is repeated 20 times per second starting at 50 Hz and continuing up to 10,000 Hz or higher.

The record is played back and its output observed by means of an oscilloscope connected at the output of the playback unit. Marker frequencies appear at intervals to identify the frequencies. The use of these records is discussed in Question 23.141.

13.177 What is the nomenclature for recording and reproducer styli?—This subject is discussed in Section 15.

13.178 What are the included angles used with recording and reproducing styli?—This information is given in Section 15.

13.179 What are the characteristics of a magnetic cutting head?—This subject is discussed in Section 14.

13.180 What are the standard groove characteristics for stereophonic recordings?—The plane of modulation in a 45/45-degree stereophonic disc groove shall have orthogonal modulation planes inclined at 45 degrees to a radial line on the surface of the disc and at the intersection of the modulation planes and be normal to the radial lines. The outer wall of the groove shall contain the *right-hand channel* information and the inner wall the *left-hand information*. The phase relationship between channels shall be such as to result in *lateral* groove displacement when the stereo-recording system is driven with equal amplitude and in-phase signals, and the groove displacement shall be *vertical* when the recording system is driven by equal amplitude signals in antiphase or 180 degrees. (See Figs. 13-180A and B, and Question 13.104.)

13.181 What is the standard groove shape for stereophonic pressings?—The groove shape shall have an included angle of 90 degrees, plus-minus 2 degrees, with a top width of not less than 0.001 inch, and a bottom radius not greater than 0.0002 inch. For these groove dimensions it is recommended the reproducing stylus have a tip radius of 0.0005 to 0.0007 inch, with an included angle of 40 to 45 degrees.

It will be noted that the groove dimensions mentioned above apply to the *finished product* rather than to the recording stylus. In some instances, the groove dimensions may depart slightly from those of the recording stylus; however, these variations can generally be controlled in the processing plant. In the event that it is necessary to play back both monophonic and stereophonic discs with the same reproducer, the use of a 0.007-inch stylus is desirable.

13.182 What is the lines-per-inch rate for starting spirals?—Eight grooves per inch, plus or minus 2.

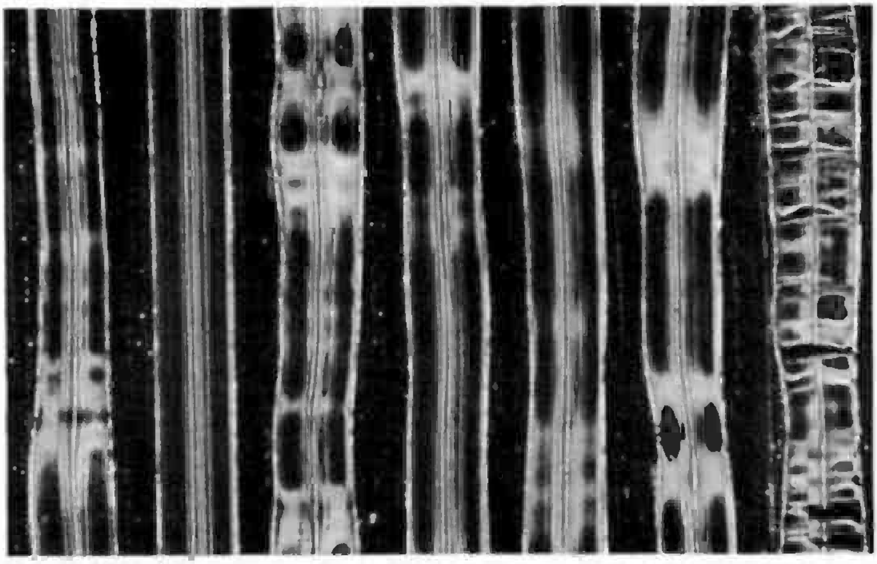


Fig. 13-180A. Photomicrograph of 45/45 grooves recorded with stereophonic program material.

13.183 How many grooves are recommended before the first modulation occurs?—Not less than two nor more than four complete revolutions.

13.184 What are the smallest recommended diameters to be recorded?—

For 16-inch transcriptions at 33⅓ rpm using a reproducing stylus of 2.3 mils, 7.5 inches.

For 78.26-rpm recordings, 3¾ inches.

For 45-rpm recordings, 4¼ inches.

For 33⅓-rpm microgroove recordings, 4¾ inches.

13.185 What is the standard for run-in grooves and the smallest groove on the inside?—There shall be at least one unmodulated groove at the recording pitch before the recording, and one at the end of the recording. The last modulated groove shall not be less than 4¾ inches in diameter for 33⅓-rpm discs, and 4¼ inches for 45-rpm discs.

13.186 What is the standard for concentricity of the center hole?—It shall be concentric with the recorded groove spiral within 0.005 inch, with a diameter of 0.286 inch plus 0.001 inch, minus 0.002 inch for 33⅓-rpm recording. For 45-rpm discs, the hole is 1.504 inches, plus-minus 0.002 inch. Warping of the disc shall not exceed a total indicator reading of the surface in excess of 1/16 inch, within any 45-degree segment, with a total reading of the indicator not to exceed 1/32 inch.

13.187 What is the recommended groove width for coarse-pitch recording?—For lateral recording, the finished groove should have an included angle of 88 degrees, plus or minus 5 degrees, for a top width of not less than 4.0 mils for records to be reproduced with a stylus having a tip radius of 2.3 mils.

13.188 What is the average dynamic range of a disc recording?—For high quality recordings, the dynamic range is:

Shellac pressings, lateral cut	32 to 40 dB
Microgroove vinylite	40 to 55 dB
Vertical cut	40 to 45 dB
Special systems, microgroove	45 to 60 dB
16-inch coarse-pitch, lateral cut	45 to 60 dB

13.189 What is the recommended groove width for vertical recordings?—For an included angle of 88 degrees, plus or minus 5 degrees, the bottom radius should be 2.0 to 2.3 mils, with a top width of not less than 4.0 mils.

13.190 What is grouping?—Uneven spacing of the grooves. The grooves should be so spaced that at no one point does the pitch deviate from the mean by more than 5 percent for constant-groove space recording.

13.191 What is flash?—It is the excess compound generated at the edge of a pressing, during the compression moulding of a disc record. The edge is later buffed smooth.

13.192 What is the procedure for recording standard test records?—Standard procedures for this type recording have been set forth in the National As-

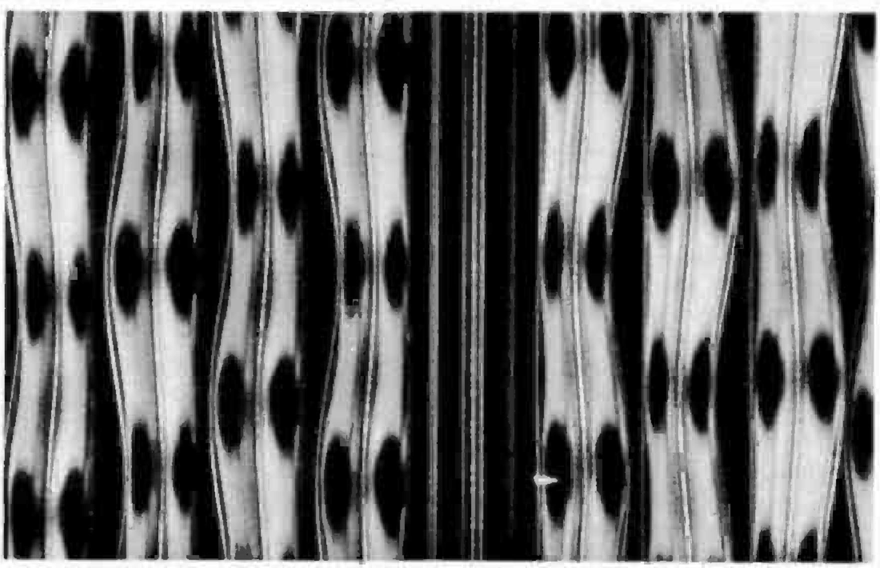


Fig. 13-180B. Photomicrograph of single-channel 45/45 stereophonic record groove.

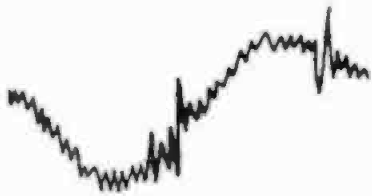


Fig. 13-193. Modulation noise superimposed on a fundamental frequency.

sociation of Broadcasters (NAB) Standard, "Disc Recording and Reproduction," March, 1964.

13.193 What is modulation noise? —Noise created by the signal. The signal is not included as a part of the noise. A typical modulation noise pattern is shown in Fig. 13-193. With the use of hot-stylus techniques, modulation noise has ceased to be a problem.

13.194 What is a poid?—The curve traced by the center of a sphere when it rolls or slides over a surface having a sinusoidal profile. It is the path traced by the reproducer stylus of a vertical pickup.

13.195 What is shaving a record?— In the days when soft wax was used for recording masters, the recording blank was rotated at a high speed and a thin shaving was removed from the surface to obtain a high polish.

13.196 How does tracing distortion differ in a vertical recording?—When a sine wave is reproduced from a vertical recording, the curve traced by the stylus tip is a poid. (See Questions 13.173 and 13.194.)

13.197 What is the transition frequency?—It is the frequency where the recording characteristic departs from constant velocity to one of constant amplitude. It is also called the crossover or turnover frequency. (See Question 14.8.)

13.198 Define the term "off-set angle."—In lateral-disc reproduction, the off-set angle is the smaller of the two angles between the projections into the plane of the disc of the vibration axis of the pickup stylus and the line connecting the vertical pivot (assuming a horizontal disc) of the pickup arm, with the stylus point. (See Questions 13.30, 13.32, and 16.60.)

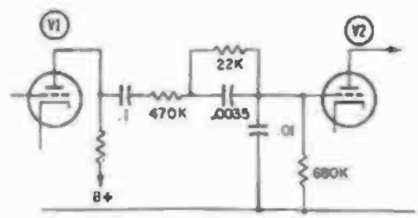
13.199 Describe a typical pickup and reproducing system test record.— Several different test records are available for making individual tests on pickup and for testing complete reproducing systems. Records for the testing of pickup characteristics include frequency response measurements with frequency announcements from 20 to 20,000 Hz, separation tests (crosstalk), pickup-arm resonance, stylus wear, wavelength loss, and compliance and phasing. A standard reference level is also included (7 cm per second). Other records for the measurement of harmonic and intermodulation distortion are also available.

Records designed for complete system testing generally include identification of left and right channels, stereo balance, loudspeaker phasing, rumble test, stylus wear, pickup-arm resonance, flutter, RIAA frequency-response section, and a standard reference level. (See Question 13.192.)

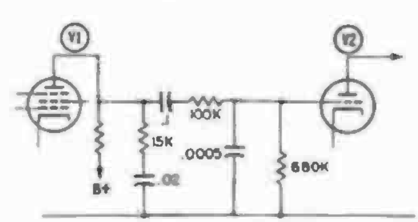
13.200 What is side thrust?—The radial component of force on a pickup arm caused by stylus drag.

13.201 Show an RC equalizer circuit suitable for reproducing the RIAA reproducing characteristic. — Two such circuits are shown in Fig. 13-291. Solid-state circuits are given in Section 12.

13.202 Show the relationship between the coils of a stereophonic cutting head and those of a pickup.—Referring to Fig. 13-202, the connections for the cutting head coils are shown at part (a) and those of the pickup are shown at part (b). Because of the mechanical design of the cutting head, the left- and right-hand sides of the program mate-

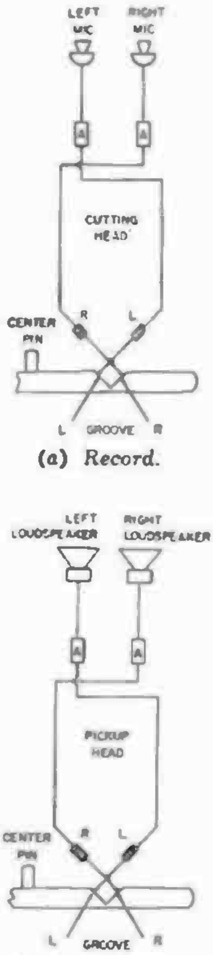


(a) *Triode.*



(b) *Pentode.*

Fig. 13-201. Equalizer circuits for RIAA reproducing characteristic. (Courtesy, Radio Corporation of America.)



(a) *Record.*

(b) *Reproduce.*

Fig. 13-202. Cross-sectional view of stereophonic record groove showing how the program material is recorded and reproduced. Because of the mechanical design of the pickup, the right- and left-hand sides of the grooves are reproduced by the coils on the opposite sides of the record groove.

rial are reproduced by the coils on the *opposite* sides of the record groove.

When the program material is recorded, as shown at part (a), the left-hand side is recorded on the *left*, or on the *inside wall of the groove*. The right-hand side of the program material is recorded on the *right, or outside of the groove*.

When the record is reproduced, the *right-hand pickup coil is connected to the left-hand amplifier and loudspeaker*. The *left-hand pickup coil is connected to the right-hand amplifier and loudspeaker*. This procedure is necessary because of the mechanical design of the pickup.

The RIAA standard for stereophonic reproduction is that when two identical signals are fed to the cutting head actuating the coils in-phase, only a lateral sound track is recorded. Conversely, equal antiphase signals will produce only vertical modulations. It is assumed that the reproducing system after leaving the pickup is properly phased. (See Question 13.158.)

13.203 What are the general characteristics of an embossed sound track? —In an embossed sound track the material is not removed from the disc but is pushed aside, creating tiny horns at the upper edges of the sound track. The frequency response of an embossed sound track is rather limited and generally falls between 70 and 5500 Hz. The signal-to-noise ratio will vary with different types of recording materials.

The harmonic and intermodulation distortion is affected by the material of the record, the amplitude of the signal, and an effect known as cusping in the sound track. A fairly good average of distortion is 6 to 10 percent at 400 Hz when recording and reproducing on the same machine. Crystal or magnetic cutting heads may be used; however, the crystal is preferred as it may be coupled to the recording amplifier to produce a constant-amplitude recording characteristic. The output stage should have at least 12 dB of negative feedback. (See Question 13.32.)

13.204 What is the frequency characteristic for an embossing recorder amplifier?—It is as shown in Fig. 13-204. It will be noted the upper midrange frequencies have been pre-equalized to obtain a greater signal-to-noise ratio and to add presence to the reproduction. The amount of pre-equalization will vary with the recording media, stylus, and cutting head.

13.205 Can embossed records be played back from the recorder using the same head and stylus?—Yes, but with a loss of signal-to-noise ratio. Using the same stylus has the disadvantage that the same included angle is used to reproduce as that used to record. This permits the stylus to ride in the bottom of the groove increasing the noise in the reproduction. The signal-to-noise ratio for this type operation is about 20

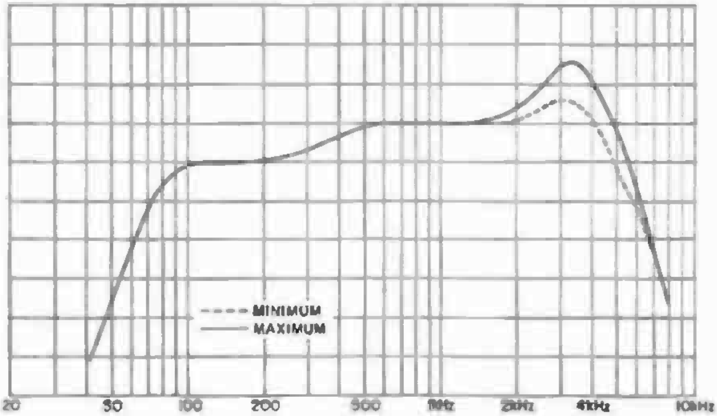
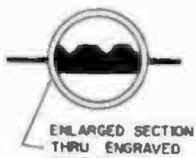


Fig. 13-204. Recording characteristic for embossing recorder.

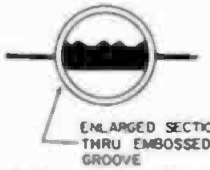
dB. Playing the same record back using a greater tip radius will increase the signal-to-noise ratio 10 dB.

13.206 What is the relationship of a constant-groove velocity to recording time?—The recording time is proportional to the area of the recording. Since the area of a circle is proportional to the square of its radius, it becomes possible to record long periods of playing time using 300 lines per inch.

13.207 Show a cross-sectional view of an engraved and an embossed sound track.—A cross-sectional view of an engraved sound track is shown at part (a), and an embossed sound track is shown at part (b) in Fig. 13-207. In the engraved sound track the material is removed from the disc as the recording progresses. In the embossed sound track the material of the disc is displaced and shoved upward, appearing as two small ridges at the edges of the sound track.



(a) *Engraved sound track.*



(b) *Embossed sound track.*

Fig. 13-207. Enlarged sections of record grooves.

13.208 What is the average pressure developed in an embossing system of recording?—About 20,000 pounds per square inch. To reduce the friction, the recording blank is impregnated with a lubricant and the surface treated with wax.

13.209 What type medium is recommended for embossed sound tracks?—As a rule, cellulose acetate about 0.15 inch in thickness is used; however, in some instances vinylite is used.

13.210 What is acoustic recording?—The original method of recording by causing the sound waves to actuate a diaphragm to which is attached a stylus. The stylus bears on the recording medium and mechanically engraves a sound track corresponding to the impressed sound waves. This is the method used before the advent of electrical recording. Acoustic recording is also called mechanical recording.

Because the output of most of the instruments used for recording is low and a considerable amount of energy is required to obtain a satisfactory level on the record, horns were attached to the string instruments to reinforce their acoustic output. A typical recording session in the early 1920s is pictured in Fig. 13-210.

The sound mixer consisted of a ball of yarn in a tube attached to the large horn at the left. The position of the ball of yarn in the tube was varied to regulate the volume of sound fed to the recording diaphragm and stylus.

13.211 Describe the geometry of a 45/45-degree stereophonic record groove cut with a 90-degree included angle stylus.—A cross-sectional view of a 45/45-

Fig. 13-210. A recording session in the early twenties at the RCA Victor plant in Camden, N.J.

degree record groove for four limiting conditions is shown in Fig. 13-211. A recording stylus with an included angle of 90 degrees is used.

The plot at the upper left illustrates the type of groove that will be recorded when the signal is fed to the left-hand channel only. The right-hand wall of the groove will be a slant line varying in depth. The right-hand edge will be smooth and without modulation. The left edge of the groove will be varied in accordance with the signal.

The plot at the upper right of the diagram depicts the reverse condition when the signal is fed to the right-hand channel only. The plot at the lower left of the diagram shows the type of groove recorded when two identical signals in phase are fed to the two cutting-head driving coils. For this condition, a vertical recording results.

The plot at the lower right shows the groove recorded when the two signals are out of phase at the driving coils. For such a condition, a lateral recording will result. During an actual stereophonic recording session, all four of the

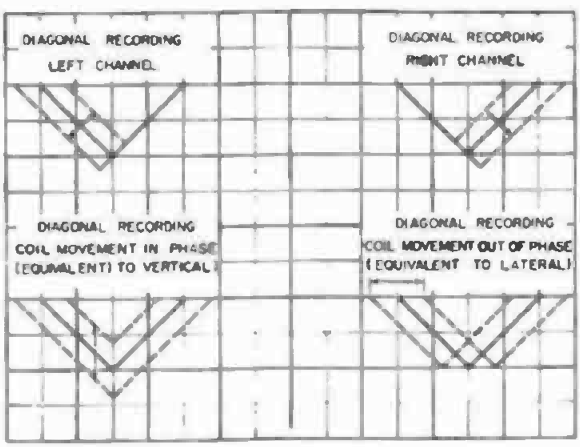


Fig. 13-211. Cross section of 45/45 grooves for four limiting conditions.

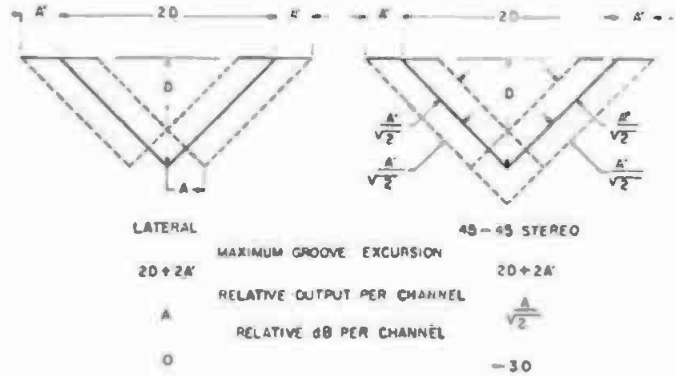


Fig. 13-212. Comparison of 45/45 stereophonic groove with standard lateral groove.

described conditions are taking place under very complex conditions of the stylus motion.

13.212 Compare a monophonic groove to that of a 45/45-degree stereophonic groove, showing the maximum permissible groove excursion. — This comparison is shown in Fig. 13-212 for the conventional lateral recorded disc and for the lateral motion in a stereophonically recorded disc. The maximum excursion for either type of groove is:

$$2\,D + 2\,A'$$

The maximum modulation for the conventional lateral groove is:

$$A = A'$$

In the 45/45 groove, the maximum modulation for either channel is:

$$\frac{A'}{\sqrt{2}}$$

Thus, each channel for a stereophonically recorder disc will have a 3-dB lower output level with respect to a conventional lateral recording.

The maximum amount of lines recommended for stereophonic recording by the 45/45-degree method is 225 lines per inch, with a minimum groove width of 1 mil and a maximum bottom radius of 0.2 mil.

13.213 What are the limiting factors relative to the maximum modulation that can be applied to a 45/45-degree groove?—Refer to the cross-sectional view of a 45/45-degree stereophonically recorded groove in Fig. 13-213. The specified minimum depth is shown at D, and the maximum modulation for either cutting-head coil is indicated at A. For these conditions, the maximum horizontal excursion will be:

$$2\,D + 4\,\sqrt{2\,A}$$

The maximum depth of the groove is:

$$D + 2\,\sqrt{2\,A}$$

13.214 What is the average harmonic and intermodulation distortion for stereophonic recording and reproducing systems?—It has become the policy of most manufacturers not to quote figures for harmonic or intermodulation distortion, as the value of distortion depends on many factors. However, intermodulation distortion at 3.5-cm velocity for

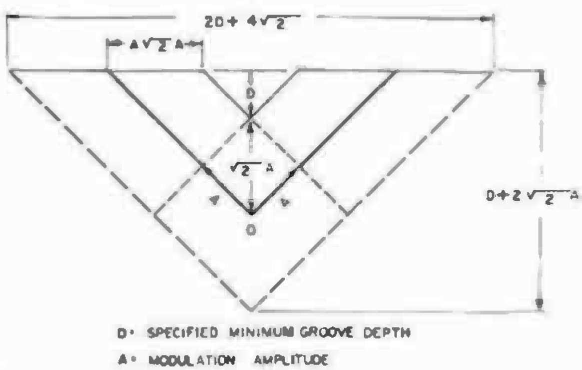


Fig. 13-213. Cross section of 45/45 groove for maximum groove excursion.

each channel will not be more than 3.5 to 4 percent for a diameter of 11 inches. As the diameter is decreased. the distortion will rise to about 5 to 6 percent. This is based on the use of 60 and 6000 Hz in a ratio of 4:1. For monophonic recording, using 7-cm velocity and 400 Hz, the distortion generally does not exceed 2.5 percent at a diameter of 6 inches.

It should be taken into consideration that these measurements include the distortion contributed by the test oscillator, recording amplifiers, cutting head and cutting-head stylus angle, and the disc record. On the reproducing side, measurements include the distortion of the disc record, pickup, preamplifiers, and any distortion induced by equalization in either the recording or reproduction circuitry. Considering the many factors involved in disc recording and reproduction, this is not a high figure of distortion. No doubt the greatest distortion factor is the angle of the stylus to the surface of the disc. (See Question 13.30.)

13.215 Describe the RCA dynagroove stereophonic disc record.—The dynagroove system developed by RCA-Victor is a combination of many factors. It is a correlation between the artist and engineer to achieve a record that will more nearly approach concert-hall reproduction in the average home by the use of several devices in the recording system. It has been found by extensive measurements that 90 percent of the consumers listen to record reproduction in their homes at a peak level of 70 to 90 dB, with the average about 80 dB. The peak level of a full symphony orchestra is about 100 dB. Thus, it may be seen that the peak level in the home is considerably lower than the concert hall. The principal reason for this lower level is that the tolerable peak level in a small room is lower than that of a concert hall, due to the shorter path of travel and the faster growth in the smaller enclosure; hence, a lower peak level of listening. Six factors enter into the realistic reproduction of sound in the home; the peak sound level, loudness versus the loudness level, frequency response of the human ear, and the reverberation characteristics of the enclosure.

When recording the original master tape in the studio, several factors are taken into consideration. such as the studio acoustics, microphone placement and directivity, frequency response, distortion of the amplifier system, and the record speed and signal-to-noise ratio. It has been found that operating the recorder at 30 inches per second, the random signal-to-noise ratio is increased 3 dB, with a reduction in flutter and wow of 50 percent, as compared to a speed of 15 ips.

After the original magnetic tape is recorded, it is transferred to a submaster tape using a Dynamic Spectrum Equalizer to correct for differences in the listening conditions and to that of an average dwelling. The design of this equalizer is such that its frequency response is altered continuously as the program is recorded, and is a function of the program amplitude and differences in the types of musical combinations and selections. Typical frequency-response curves for this equalizer are shown in Fig. 13-215.

When the level is low, the low frequencies are accentuated. At medium levels, only a slight amount of accentuation is used in the low frequencies and in the presence range (2000 to 6000 Hz), with a reduction in the response in the region between 400 and 1000 Hz. At high sound levels, the presence region is accentuated and the range below 1000 Hz is reduced. The whole object is to increase or decrease the sound level in the appropriate frequency bands so the program material may be appreciated under the existing ambient noise level and acoustic environment. Properly handled, the dynamic range is not disturbed and reproduction in a small room enhanced.

In the recording of both the master tape and the submaster, they are carefully monitored using peak-indicating volume-indicator meters. After the submaster tape has been recorded, it is transferred to a master disc, using devices termed, "Dynamic Styli Correlators," in the left and right channels. This device corrects for the discrepancies between the angle of the recorder and reproducer stylus used by the consumer and reduces the tracking distortion at the smaller diameters by a factor of 6:1.

In addition to the factors mentioned, certain changes have been made in the record-processing procedure to improve

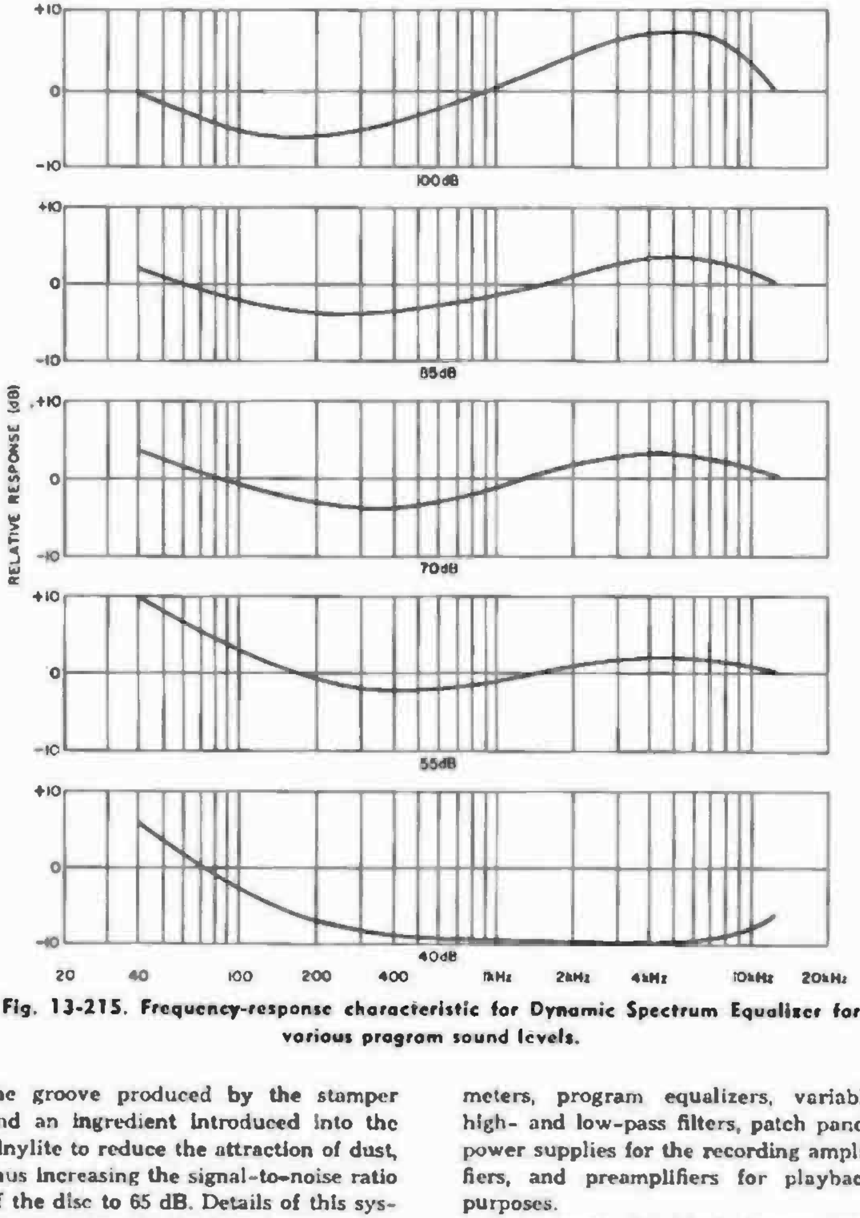


Fig. 13-215. Frequency-response characteristic for Dynamic Spectrum Equalizer for various program sound levels.

the groove produced by the stamper and an ingredient introduced into the vinylite to reduce the attraction of dust, thus increasing the signal-to-noise ratio of the disc to 65 dB. Details of this system are given in the literature.

13.216 Describe a monophonic and stereophonic magnetic tape or film to disc transfer channel.—Pictured in Fig. 13-216A is a Westrex magnetic tape- or film-transfer channel for monophonic or stereophonic disc mastering. The channel shown is complete except for the monitor speakers, disc recording lathe, magnetic tape or film reproducer, playback turntable, and pickup.

Two cabinet-type racks house two RA-670 amplifiers, two RA-1574D 75-watt recording amplifiers, gain controls, two 40-watt monitor amplifiers, VU meters, program equalizers, variable high- and low-pass filters, patch panel, power supplies for the recording amplifiers, and preamplifiers for playback purposes.

Referring to the block diagram Fig. 13-216B, at the upper left the input signals from the left and right channels are applied to the input of two RA-670 limiter amplifiers. These amplifiers have extremely fast attack time (50-microseconds) and a variable release time. Leaving the limiter, the signals pass through high- and low-pass filters, then to the variable program equalizers. At the output of the equalizers is a gain control for balancing the gain of each channel and a two-gang attenuator for adjusting the gain of the two sides simultaneously.

The signals are now applied to a monitor switching panel where the monitoring may be taken from one of several different points in the system—the left or right cutting-head circuits, the recording channel immediately preceding the recording amplifier, the tape reproducer, or the phono jack. Two line amplifiers feed through a record on-off panel and then to two RA-1574D power amplifiers, which drive a Westrex 3-D cutting head on the recording lathe.

At the lower portion of the diagram are shown the 40-watt monitor amplifiers, preceded by a two-gang gain control and two repeat coils. At the lower left is a stereo preamplifier for playback purposes, and pickup.

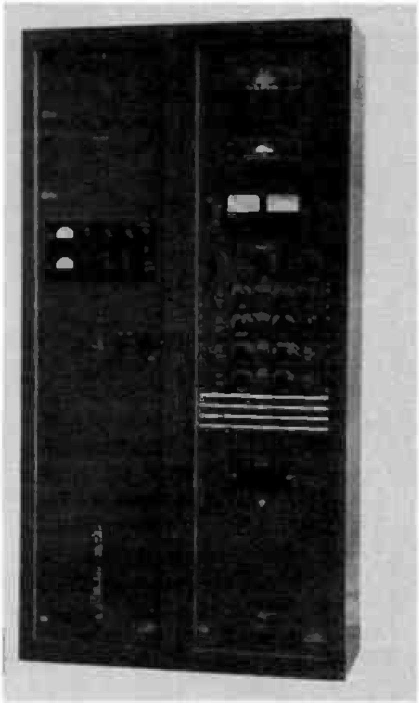


Fig. 13-216A. Westrex Corp. Model 2300 StereoDisc magnetic tape to disc record transfer channel.

Referring to the schematic diagram in Fig. 13-216C, the RA-1574D StereoDisc recording amplifier, the incoming signal is fed to a 600-ohm impedance input transformer T1 which drives a cathode follower V1A that functions as a constant-voltage generator for the plug-in equalizer. The output from this unit is amplified by V2A and B and has a low-frequency boost for the RIAA recording characteristic. (See Question 17.162.) The signal from V2B drives V3B, then phase-inverter V3A which in turn drives power stages V4A, V4B, V5, and V6. The output transformer T2 secondary is designed for a load of 5 to 10 ohms.

A negative-feedback circuit is returned from the feedback coil in the cutting head to potentiometer P1 in the control grid of tube V7B at the lower left. The feedback signal is amplified by tubes V7B, V7A, and V1B. The output signal from V1B injects feedback voltage into the cathode of V3B and provides a high-level signal to pin A of the monitor-amplifier equalizer C-98461 at the lower center of the diagram. The signal at the output of the equalizer drives tubes V8B and V8A feeding output transformer T3, which provides a 600-ohm output at a level of plus 4 dBm, for feeding the monitor power amplifier and speaker. The plus 4 dBm develops for a peak velocity of 3.54 cm/sec recording level at the cutting head stylus tip. Tubes V9 and V10 are employed for voltage regulation for the screens of V5 and V6.

The circuitry for the equalizers is given in Figs. 13-216G and H. The Westrex 3-D cutting head is discussed in detail in Question 14.2.

Two power supplies are required, one for each amplifier. The power supply is unregulated, regulation being supplied by the circuitry in the RA-1574D amplifiers. Referring to the schematic diagram in Fig. 13-216D, the power transformer T1 has three primary taps for line voltages of 105 to 120 volts. Full-wave high voltage is obtained through rectifiers CR4 and CR5, consisting of two units in parallel, feeding a choke input filter L1. Negative-bias voltage is generated by rectifier CR1 and is adjusted by control P1. The normal voltage is around 38 volts. The dc heater voltage is obtained from rectifiers CR2A and CR2B operating as voltage doublers. The output voltage is adjusted by control R1. Pilot-light voltage is supplied from the same winding. Transformer T2 provides isolation and heater voltage for regulator tube V9. A delay of 50 seconds in applying the 600 volts dc to the amplifier is obtained by the use of a delay unit VS3. Interconnections between the RA-1574D amplifier and power supply are shown at the lower left of the diagram.

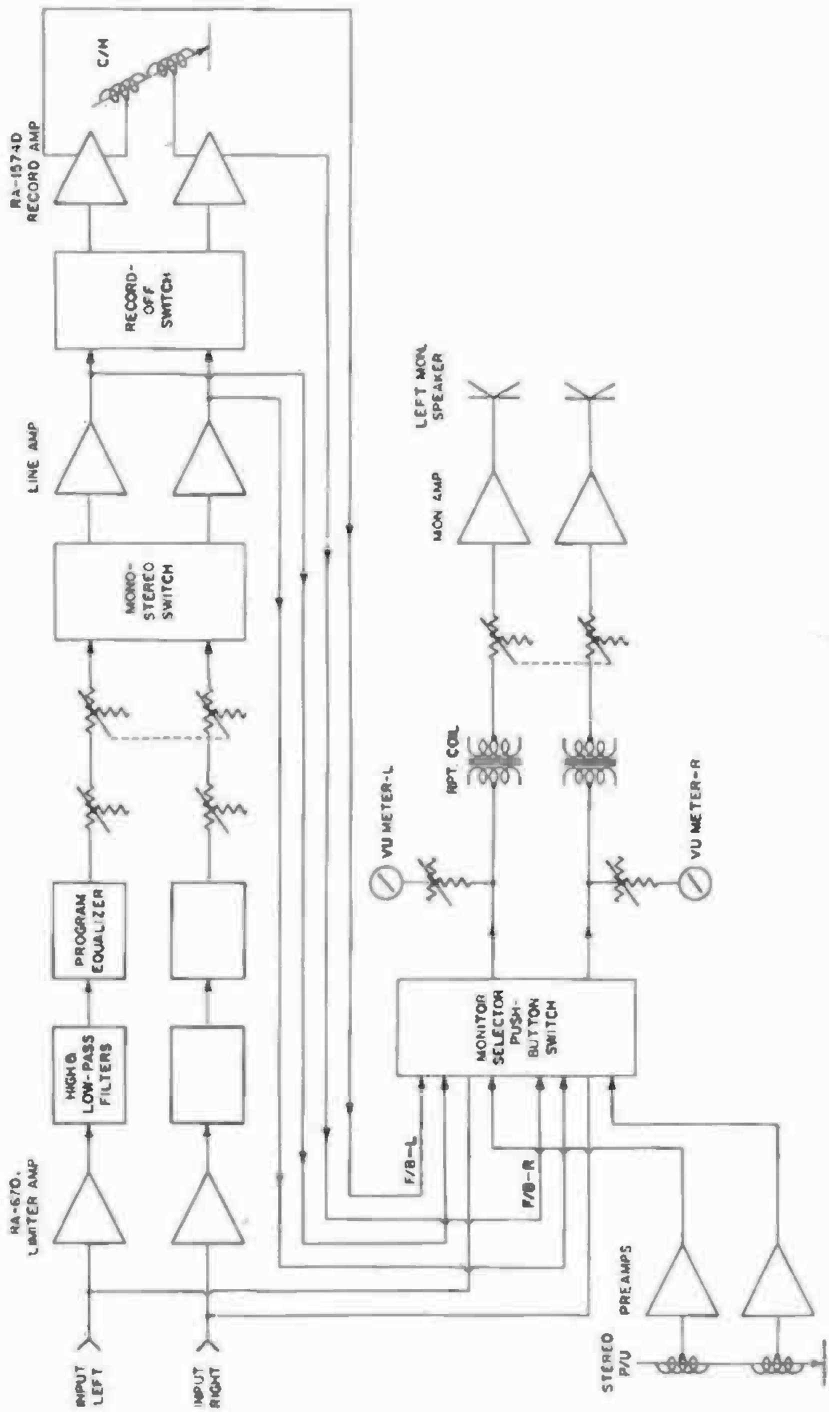


Fig. 13-216B. Block diagram of the Westrex Corp. Model 2300 StereoDisc magnetic tape to disc record transfer channel.

Since the monitoring signal is taken from the negative-feedback circuits in the cutting head, a network (C-98461) is required having an *inverse frequency characteristic* to the RIAA characteristic (Fig. 13-216H). This creates the proper listening response. Equalizer C-98451 is connected between tubes V1A and V2A to supplant the negative-feedback loop characteristics and pro-

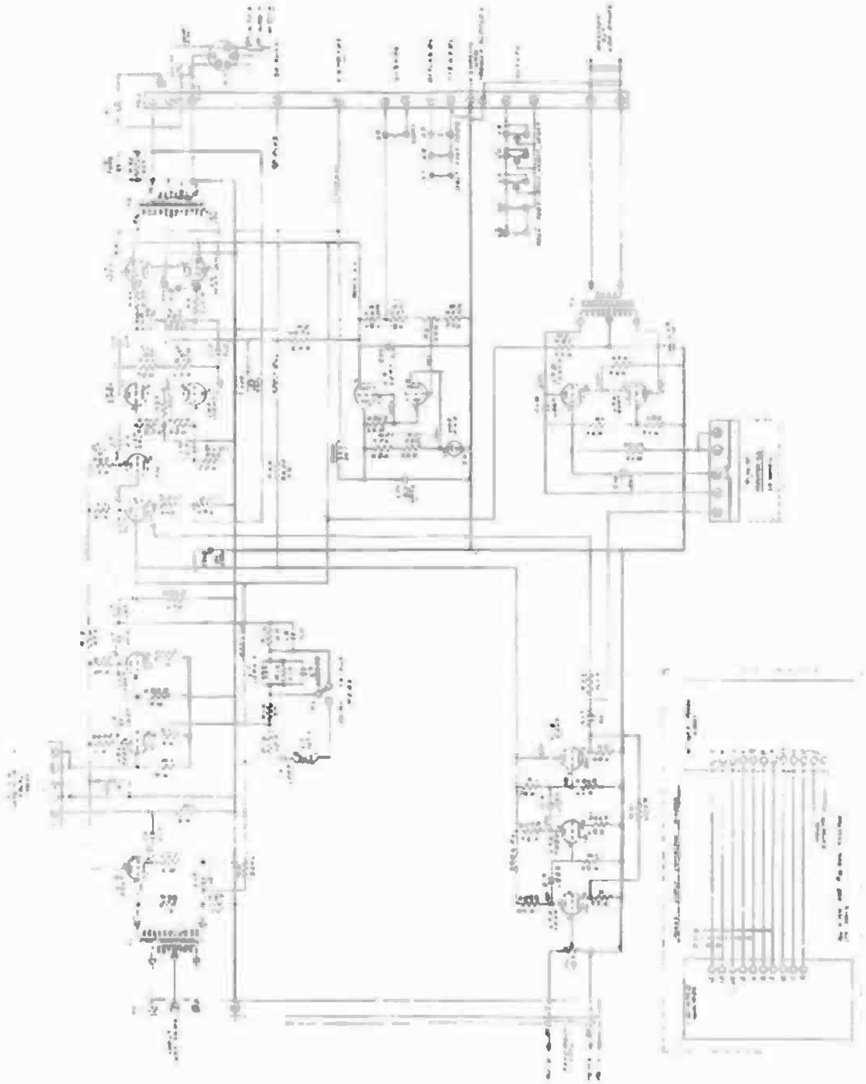


Fig. 13-216C. Schematic diagram for Westrex Model RA-1574D stereophonic recording amplifier.

vide an RIAA recording characteristic from 30 to 1000 Hz, and then constant-velocity from 1000 to 15,000 Hz (Fig. 13-216G). The insertion loss for this network is approximately 18 dB.

The statistics for the RA-1574D amplifiers are: Input impedance 600-ohms. Cutting head output 10 ohms for driving coils, and 11 ohms for negative-feedback coils. Sensitivity 14 dBm for 3.54 cm/sec peak recording velocity at 1000 Hz for both channels. Adjnstable feedback up to 29 dB. Power output 75 watts continuous for 1 percent THD. Output noise 3 millivolts across 10 ohms. Monitor output level for a peak velocity of 3.54 cm/sec is plus 4 dBm.

13.217 Describe the difference between a lateral-vertical and 45/45-degree stereophonic recording system.—The basic principles of 45/45-degree recording heads have been discussed in Questions 13.211 to 13.213 and will not be repeated here. Both of the above methods of recording stereophonic sound cut *identical* grooves. In the 45/45-degree system, the driving coils of the cutting head exert their forces at an angle of 45 degrees with respect to the recording disc surface. In the lateral-vertical method, a *matrixing network* composed of two transformers employing double secondaries convert the left and right channel information into *sum of vectors* for each channel lateral and vertical. The resulting signal then becomes identical with the 45/45-degree system of recording.

The principal difference, in the two systems is the *initial* alignment of the left and right sides. Separation in the 45/45-degree system is fixed and depends on the mechanical construction of the cutting head. In the lateral-vertical system, this parameter is a function of balance of *gain* between the two

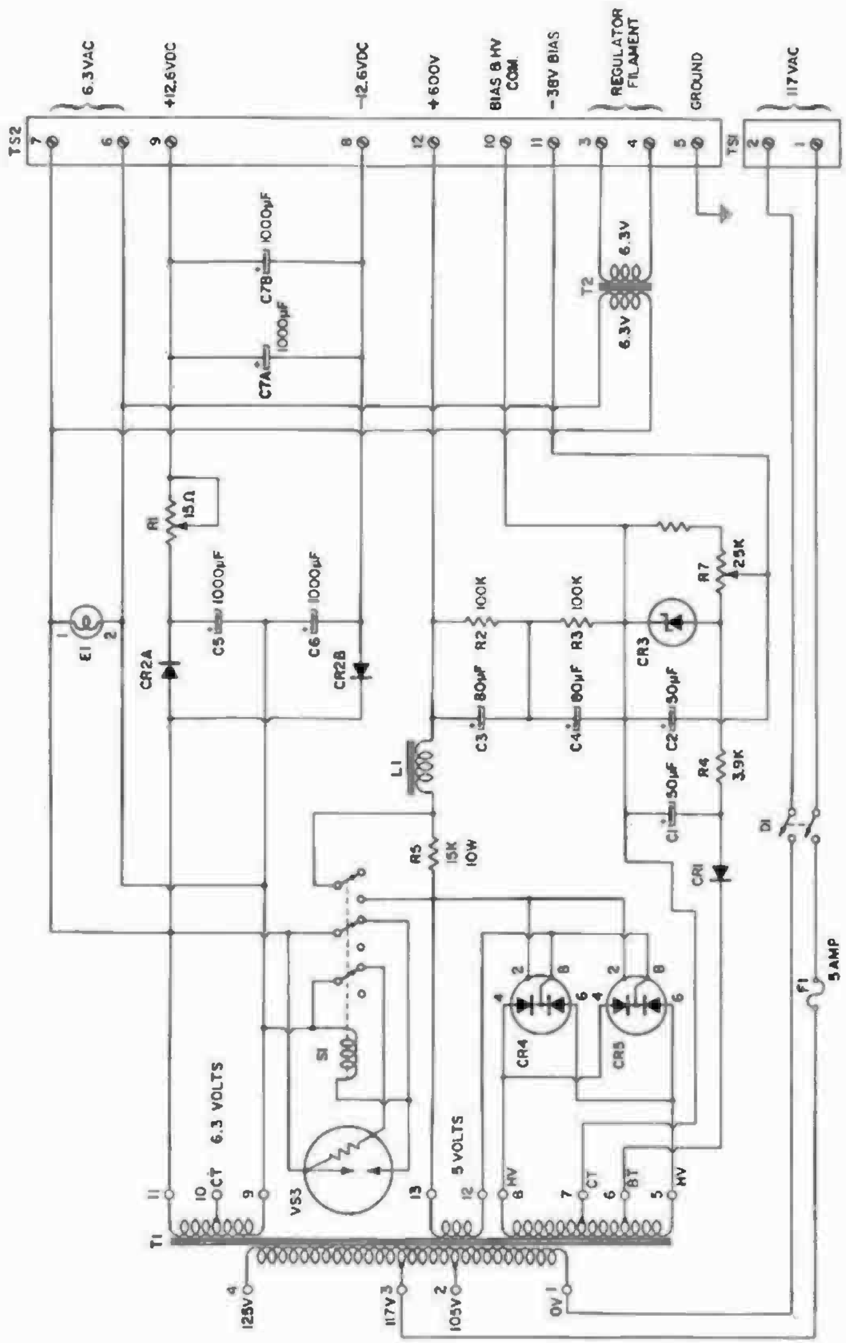


Fig. 13-216D. Westrex Corp. RA-1567 power-supply unit for RA-1574D power amplifier.

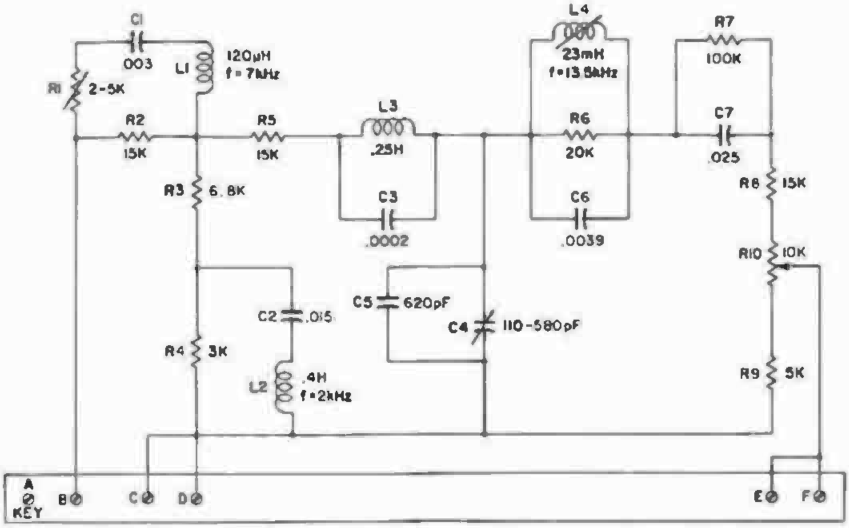


Fig. 13-216E. Schematic diagram for Westrex Corp. C-98451 equalizer (RIAA) for use with the RA-1574D StereoDisc Recording Amplifier.

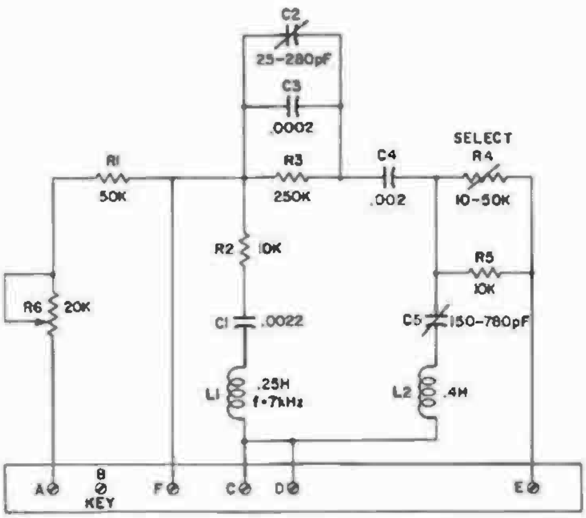


Fig. 13-216F. Schematic diagram for Westrex Corp. C-98461 equalizer (Monitor) for use with RA-1574D StereoDisc Recording Amplifier.

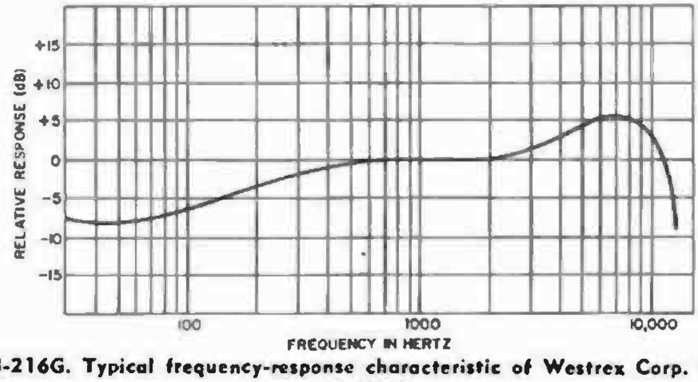


Fig. 13-216G. Typical frequency-response characteristic of Westrex Corp. C-98451 equalizer.

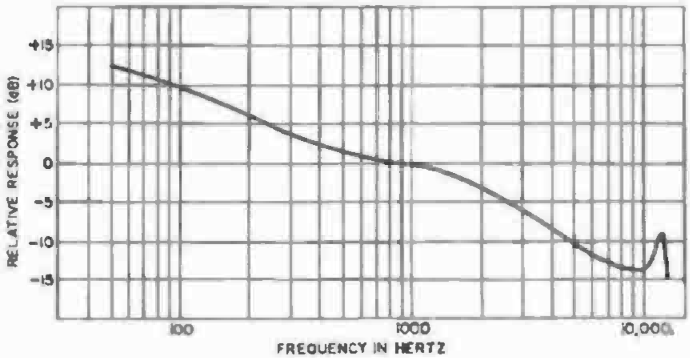


Fig. 13-216H. Typical frequency-response characteristic of Westrex Corp. C-98461 equalizer.

sides of the system. Because the gain is fixed by the matrixing network, an absolute balance is not possible.

Since a lateral-vertical recording depends on two vector forces produced simultaneously by the lateral and vertical sides, varying the gain of one channel results in a changing of the resulting driving forces, thus reducing cross-talk ratio between the two sides. Also, if the two sides are not of identical frequency response, the effect is that of a change in gain in one side. Because of the above disadvantages, the 45/45-degree method of recording has been adopted by most recording organizations.

Recording heads manufactured by Holtzer, Westrex, Neumann, and Ortofon are the 45/45-degree variety, while those of Fairchild employ the lateral-vertical design.

The lateral-vertical system is aligned initially by adjusting the amplifier gain of each side at 1000 Hz for maximum channel separation, then again adjusted for a uniform frequency response and separation simultaneously.

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Section 14

Cutting Heads

Because angles, mechanical mounting and alignment, and electronics are critical, an understanding of physics, mechanics, and electronics is useful in the understanding of disc cutting heads (recording heads).

This section discusses both monophonic- and stereophonic-type recording heads, and their associated equipment and recording techniques. Crystal, ceramic, and various types of magnetic cutting heads are discussed, as well as cutting head calibration. Hot- and cold-stylus recording techniques are also discussed.

14.1 What is an electromechanical transducer?—A device which transforms electrical energy into mechanical energy, or vice versa. A recording or cutting head is a typical example of this principle.

14.2 Describe the basic principles of a cutting head and the details of construction.—Recording or cutting heads, as they are commonly called, are electromechanical transducer devices that translate electrical waveforms into mechanical motion and are used to engrave a sound track on a plastic or nitrocellulose disc record. Cutting heads may be designed to record in the lateral or vertical direction or a combination of both for stereophonic recording. The essential parts for a lateral-type head are shown in Fig. 14-2A. A permanent magnetic field is supplied by a magnet, indicated by north and south poles N and S. A soft-iron armature A is supported and centered in the magnetic field on a knife-edge bearing B and connected to a damping spring at the rear (not shown). Two coils C surround the ends of the armature and carry the audio-frequency currents to be recorded. These currents induce a magnetic field in the armature that is continuously changing polarity. This causes the armature to be attracted or repelled by the permanent magnet in accord-

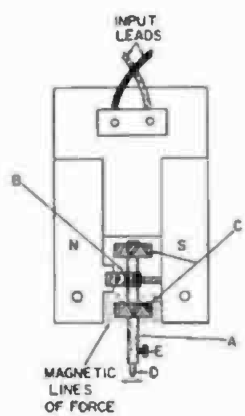


Fig. 14-2A. Mechanical and magnetic structure of a magnetic cutting or recording head.

Fig. 14-2B. Neumann monophonic cutting head Model ES-59-020. The stylus heating coil for hot-stylus recording is shown wrapped around the recording stylus. (Courtesy, Gotham Audio Corp.)

Fig. 14-2C. Exterior view of Neuman Model SX-45 stereophonic cutting head. (Courtesy, Gotham Audio Corp.)

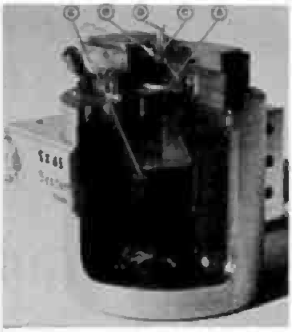


Fig. 14-2D. Cutaway view of Neumann Model SX-45 stereophonic cutting head showing the interior construction. (Courtesy, Gotham Audio Corp.)

ance with the strength of the induced audio current; thus, the stylus D mounted in chuck E enscribes the electrical waveform applied to the coils C in a disc record. Although cutting heads can be obtained similar to the basic construction, present-day commercial recording heads generally make use of negative-feedback coils within the heads, as such design reduces distortion, increases the signal-to-noise ratio, and has a more uniform frequency response.

A commercial cutting head, manufactured by Neumann of West Germany is shown in Fig. 14-2B. This head is designed for monophonic lateral groove recording and employs a negative-feedback moving-coil structure, having a resonant frequency of 55,000 Hz, thus permitting a large value of negative feedback to be used. Its frequency characteristic is linear over a range of 30 to 24,000 Hz, without equalization. The armature is made in the shape of a hollow cone, the apex forming the stylus holder. Near the tip of the cone are two feedback coils, with the driving coil mounted at the opposite end or open end of the armature. The feedback voltage is a true representation of the stylus motion. The audio signal, negative feedback, and the stylus heating-coil connections are brought to the back of the head on a 10-pin subminiature connector. A sapphire recording stylus mounted in an aluminum shaft is inserted by means of a stylus wrench in

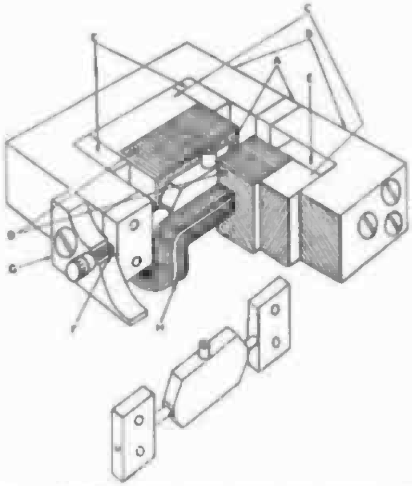


Fig. 14-2E. Inverted interior view of balanced-armature, moving-vane magnetic cutting head, using negative feedback, for monophonic recording.

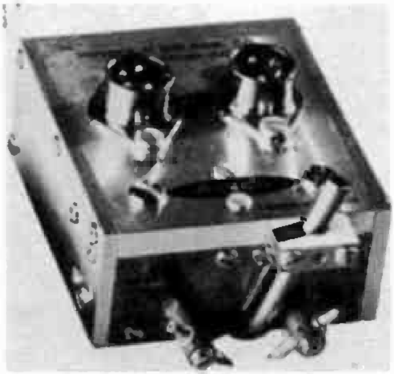


Fig. 14-2F. Holzer Audio Engineering Co. Model SC-1 stereophonic cutting head. It is of moving-iron feedback design requiring no linkage or suspensions.

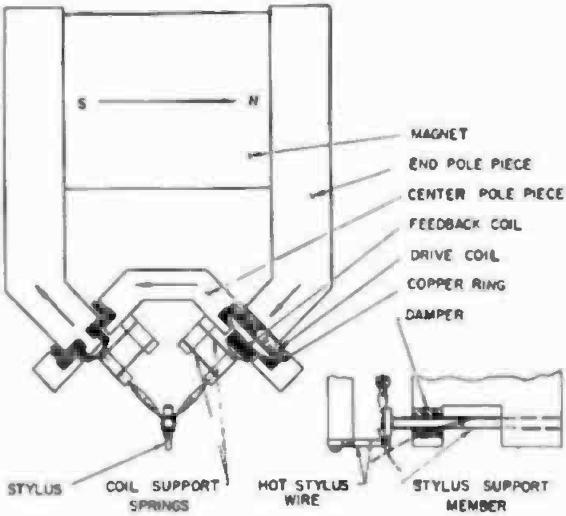


Fig. 14-2G. Simplified cross-sectional view of Westrex Corp. stereophonic cutting head.

the stylus chuck. The stylus heater wires are connected by means of spring-loaded binding posts. A special amplifier having a continuous output power of 75 watts is required. The impedance of the feedback coils is 150 ohms, and of the drive coils, 6 ohms.

In Fig. 14-2C is an exterior view of a stereophonic recording head, showing connections to the stylus heating coil. The interior construction is shown in Fig. 14-2D. The moving-coil structure A is somewhat similar to its monophonic counterpart in Fig. 14-2B, using an aluminum oxide armature, on which are wound two drive coils B for the left and right channels. The feedback coils are mounted at right angles to drive-coils directly at the stylus shank C. This type of construction permits both drive and feedback coils to be wound on a common armature and has the advantage of assuring that the feedback voltage is an actual reproduction of the cutting stylus excursions. The driving-coils are so wound (onto the armature) as to result in a direct 45/45-degree motion *sum and difference formation or matrixing*. The diaphragm structure suspension has the property of infinite stiffness in the lateral plane and complete elasticity to vertical and rotational movements. The stylus and heating coil are shown at D, and the permanent magnet structure is at E. Two special power amplifiers are required for operation. The frequency range is 30 to 16,000 Hz. Channel separation at 100 Hz is greater than 20 dB, at 1000 Hz, 30 dB, and at 10,000 Hz, 20 dB. Motional feedback is approximately 47 dB. Maximum current capability is 500 milliamperes.

A moving-vane balanced armature stereophonic-type cutting head is one of the simplest in construction, as it employs no springs or balancing mechanism. An interior view of its constructional features appears in Fig. 14-2E. The ends of the armature A, both in the assembly and below, are clamped between two U-shaped steel yokes B, within which lie two magnets C and

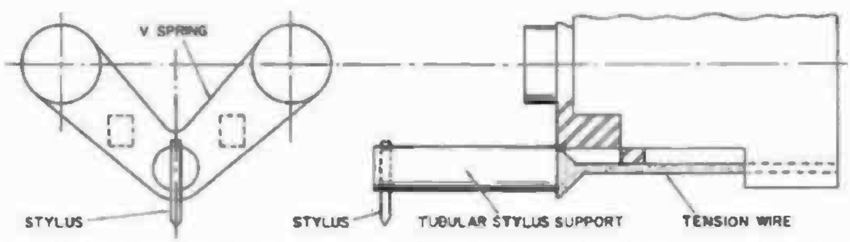


Fig. 14-2H. Stylus support member for Westrex Corp. Model 3D stereophonic recording head.

laminated pole pieces D, with brass clamping blocks E. The metal at the bottom of the slots is shaped to form torsion bars which support the active center position of the armature on the gap between the pole pieces. To prevent nonaxial deflections, these torsion bars are made as short as possible, consistent with reasonable stress at maximum excursion. Through the center of one torsion bar passes the long shank of the cutting stylus clamping screw F; the thread is carried in an external block G. The shank has high torsional compliance so that the presence of the clamping screw does not add appreciably to the mechanical impedance of the armature.

The coil H lies in slots in the faces of the pole pieces, within the main winding is a second coil connected to the cathode of the second stage of the driving amplifier to provide the negative-feedback voltage. The mechanical resonance of the armature, which is around 10,000 Hz, is damped out by the use of silicon damping fluid in the air gaps between armature A and pole pieces D. Silicone oil has the property of maintaining a constant viscosity over a wide temperature range.

A stereophonic cutting head developed by R. T. Speiden and manufactured by Holzer Audio Engineering Co., (HAECO) is shown in Fig. 14-2F and has similar constructional features. The impedance is 16 ohms, with a constant-velocity frequency response plus-minus 1.5 dB from 1000 Hz to 15,000 Hz. It is recommended that 10 dB of negative feedback be introduced at 1000 Hz. For a flat frequency response down to 30 Hz, the feedback is increased to 16 dB. The sensitivity is such that 1 watt of power (plus 30 dBm) will produce a 7-centimeter cut from 1000 to 15,000 Hz. Adding the RIAA pre-equalization characteristic of 13.75 dB at 10,000 Hz requires approximately 20 watts for full modulation. The design is such that at no time does the feedback voltage of the cutter drop to less than 90 degrees,

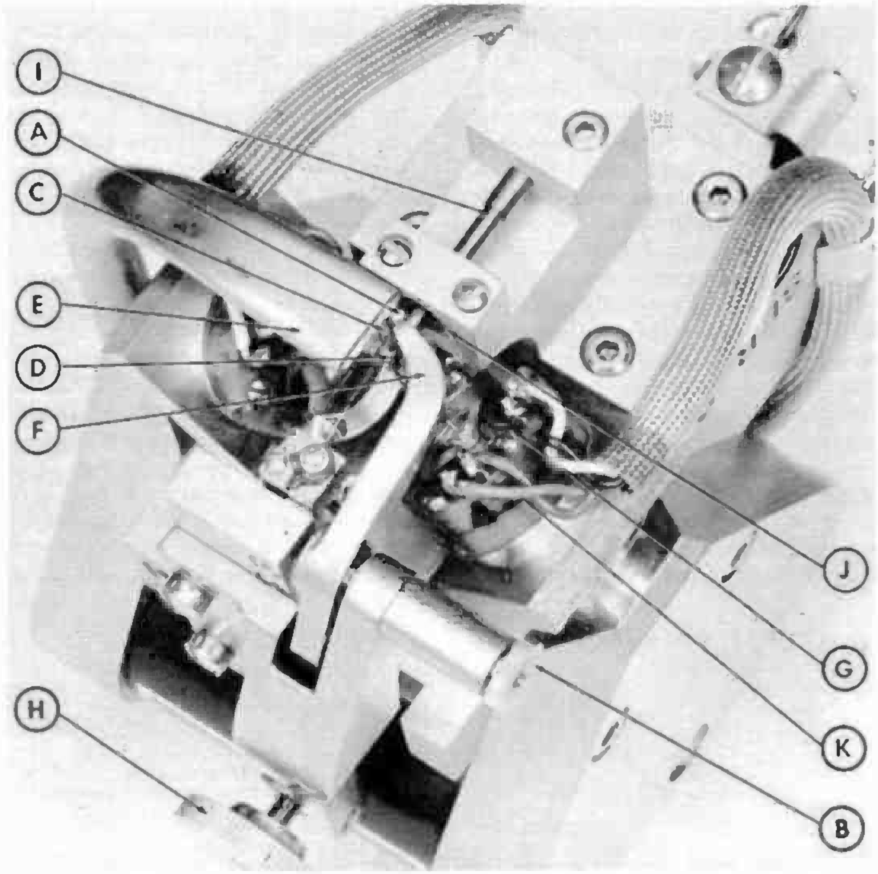


Fig. 14-2I. Underside view of Westrex Corp. StereoDisc recording head.

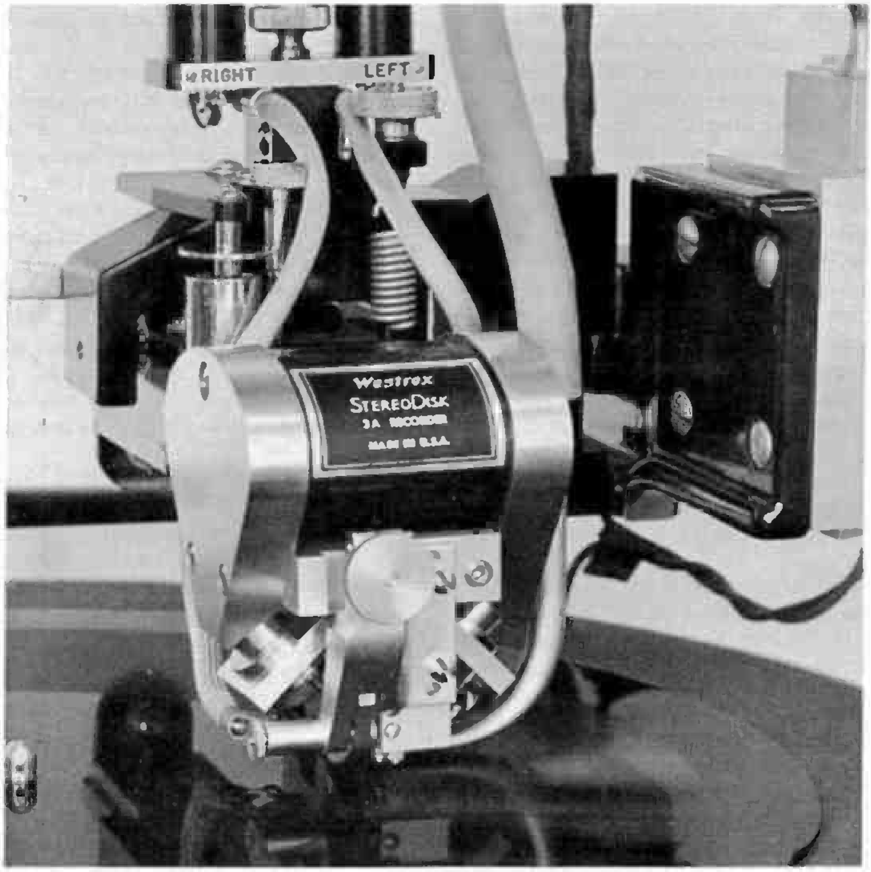


Fig. 14-2J. Westrex StereoDisc cutting head mounted on recording lathe.

or increase to more than 270 degrees from the signal input to the driver system.

A simplified constructional diagram of the Westrex Corp. Model 3C and 3D StereoDisc recording heads (Westrex refers to cutting heads as recording heads) is given in Fig. 14-2G. The following discussion holds true, generally, for both the 3C and 3D heads, except for certain modifications mentioned in the discussion. The cutting-head assembly contains two magnesium coil form assemblies, each associated with a given recording channel. Each coil assembly contains a driving coil and a negative-feedback coil located in separate pole pieces with annular gaps. The pole pieces are attached to a single Alnico V dc magnet. The magnetic gaps of the driving and feedback coils are arranged in a series-parallel fashion to insure equal flux densities in the corresponding gaps.

Copper slugs or shields are used to reduce cross talk between the driving and negative-feedback coils. These shields or slugs can be seen in the shaded areas near the coils. The springs supporting the coils are made of beryllium copper and are V-shaped to maintain the alignment of the coils.

The coil assemblies are attached to the stylus holder through links which are stiff longitudinally, but flexible laterally. These links are braced in the center to prevent excessive lateral compliance. This structure results in a stiff forward driving system with a high compliance in the lateral direction.

The supporting member for the Model 3C stylus consists of a tubular cantilever spring (Fig. 14-2G), with the support used in the Model 3D shown in Fig. 14-2H. This type of support was selected because the compliance of a cantilever spring will be the same for all directions of motion. This compliance permits the stylus to present a uniform impedance to complex motions in any direction in the vertical plane. This uniform impedance is particularly true for those frequencies in the recording spectrum where the negative-feedback voltage exercises little control over the stylus motion. The use of a cantilever spring also reduces the tendency of cross talk between the two channels because of the rotational compliance of the stylus. The damping material at the front of the stylus support has little or no effect on the recording at frequencies below 10,000 Hz and little effect above 10,000 Hz. Its purpose is to smooth out the peaks and valleys in the monitor output.

The system damping is not affected by temperature because damping is supplied by the negative-feedback coil in the driving amplifier system. The driving coil impedance is 10 ohms, and the negative-feedback coil impedance is 11 ohms.

The stylus for the 3D head is somewhat different from that of the 3C head, in that it can be replaced without removing the head assembly from the recording lathe. The stylus diameter is 1.5 times that of the 3C stylus and has a flattened face 5-mils deep and a ground flat of 39 mils, ground the full length of the stylus shank. These details are discussed in Question 15.73. The stylus heating coil for both the 3C and 3D heads consists of 7½ turns of 0.005-inch resistance wire having a resistance of 32 ohms to the foot. The coil is wound on a mandrel 0.038 inches in diameter or on the shank of a No. 62 drill. The completed coil is slipped over the stylus shank and held in place by the natural spring tension of the wire.

The heating current for the coil can be taken from a 6.3-volt transformer with a resistance in series with the coil or from a Variac in the primary winding.

Fig. 14-2I is an underside view of the 3B head, with its principal components called out. They are: (A) advance ball, (B) advance ball lateral adjustment, (C) stylus tip, (D) heater coil, (E) chip suction pipe, (F) stylus linkage, (G) coil assembly, (H) groove depth, (I) stylus support, (J) damping member, (K) driving and negative-feedback coil leads.

In Fig. 14-2J is pictured a Westrex Model 3A recording head mounted on a recording lathe. This method of mounting the head assembly is quite similar for all Westrex heads. Vertical adjustment screw A and lateral screw B are for adjusting the angle of the recording stylus. Because of the wide difference of opinion relative to the recording angle, Westrex recommends a cutting angle of 23 degrees, established with the Model 3C recording head. Distortion versus the effective vertical-cutting angle is an extremely complex problem involving such factors as frequency, peak stylus velocity, groove spacing, groove diameter, depth of cut, and the properties of the recording blank. (See Question 13.30.)

The negative-feedback circuit in both the 3C and 3D heads displays a

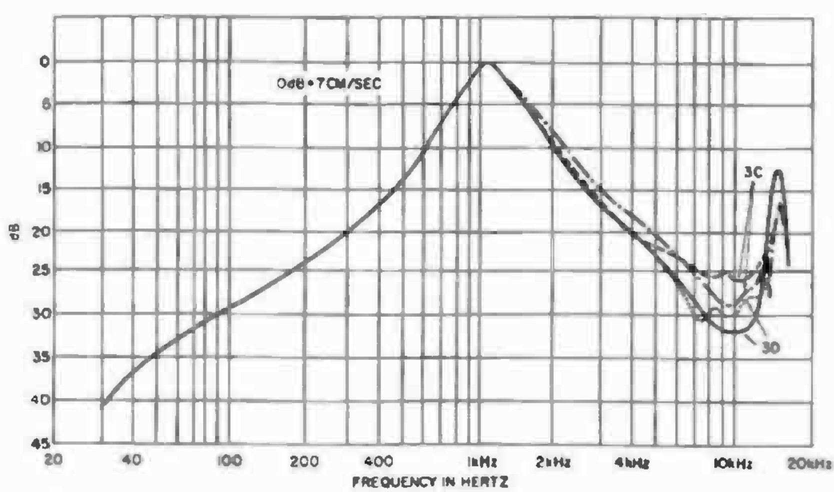


Fig. 14-2K. Resonant-frequency characteristic, without feedback, of Westrex Corp. Model 3C and 3D negative-feedback recording heads.

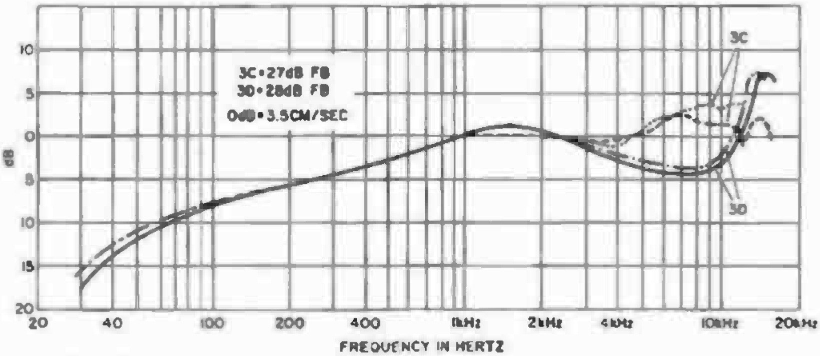


Fig. 14-2L. Frequency characteristics, with feedback, of Westrex Corp. Model 3C and 3D negative-feedback recording heads.

resonant peak around 1300 Hz (Fig. 14-2K). Applying 27 dB negative feedback to the 3C head or 28 dB to the 3D head, the frequency response appears as given in Fig. 14-2L. After connecting the head to a special amplifier (RA-1574D described in Question 17.228) with its equalizers, the frequency response at both the low and high ends is flattened out to within plus-minus 1 dB, 50 to 12,000 Hz, and plus-minus 2 dB, 12,000 to 15,000 Hz. RIAA equalization is provided from 30 to 1000 Hz. The 3D head has an overall increase in sensitivity of about 25 percent, as compared to that of model 3C.

14.3 What are the normal frequency characteristics of a magnetic cutting head?—Constant velocity as shown in Fig. 14-3. It will be noted that, using 500 Hz as a reference frequency, each time the frequency is doubled the amplitude is halved. If the frequency is halved, the amplitude is doubled. This is a true constant-velocity characteristic. However, this characteristic is not suitable for commercial recording, as frequencies below 250 Hz would be overmodulated or overcut and the high frequencies would be too low in amplitude. To prevent overcutting at the low frequencies, cutting heads are designed to record a modified, constant-amplitude constant-velocity characteristic as shown in Fig. 14-6A.

14.4 What does the term "constant velocity" mean?—It means that the stylus will travel the same distance in a given time, regardless of the frequency. To better illustrate the term, refer to Fig. 14-3 and assume the amplitude at 1000 Hz is one inch. At 2000 Hz the frequency has been doubled but the amplitude is half that at 1000 Hz; therefore, the distance traveled by the stylus in one second is still one inch. At 100 Hz, the amplitude is 10 times that at 1000 Hz; however, the frequency is 1/10

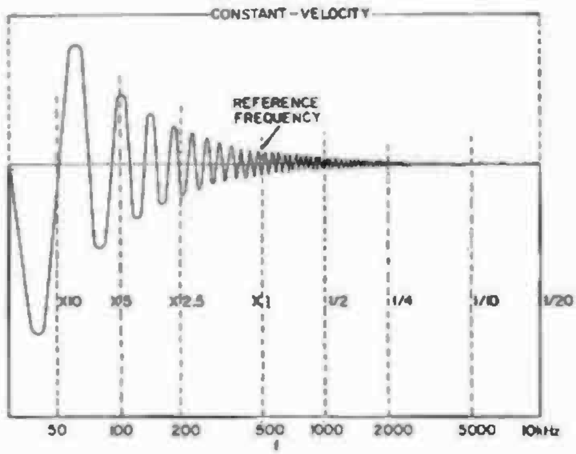


Fig. 14-3. Frequency characteristic of a magnetic cutting head operating as a constant-velocity device.

that at 1000 Hz. Therefore, the same distance is traveled by the stylus. The same reasoning holds true for any frequency within the operating range of the head.

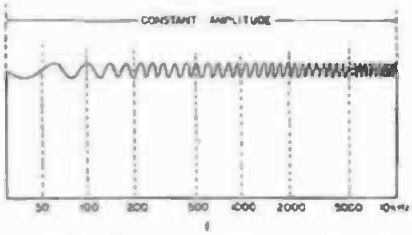


Fig. 14-5. Frequency characteristic of a crystal cutting head operating as a constant-amplitude device.

14.5 What does the term "constant amplitude" mean? — That for a constant signal amplitude to the head, the recording characteristic will be of a constant amplitude as shown in Fig. 14-5. A crystal cutting head is a good example for describing constant-amplitude recording. Because the impedance of a crystal cutting head decreases as the frequency increases, the crystal impedance will be high with respect to the coupling circuit over the entire operating range. Therefore, it may be operated as a constant-amplitude device. It will be noted that, regardless of frequency, the amplitude swing of the stylus is constant. This is the normal recording characteristic of a crystal cutting head. Crystal cutting heads are not used professionally, but have been used on home recording equipment quite extensively.

14.6 What is the recording characteristic used with magnetic cutting heads?—A modified constant-amplitude constant-velocity characteristic as is shown in Fig. 14-6A. It will be noted that a frequency, called the turnover frequency, separates the characteristic into two parts. The portion to the left of the turnover frequency is called the constant-amplitude portion and that to the right the constant-velocity portion. The constant velocity portion behaves as described in Question 14.3.

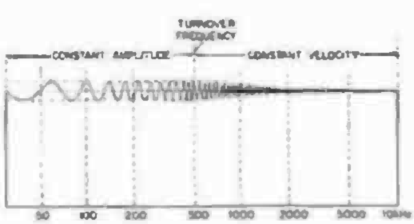


Fig. 14-6A. Frequency characteristic of a magnetic cutting head adjusted for a constant-amplitude, constant-velocity recording characteristic.

To achieve the constant-amplitude portion of the characteristic, the cutting head is so designed that regardless of frequency, the amplitude swing below the turnover frequency is constant.

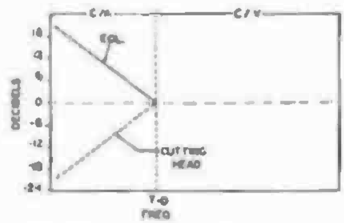


Fig. 14-6B. Equalizer frequency characteristics required to compensate for the constant-amplitude characteristics of the cutting head when reproducing.

To secure a constant amplitude below the turnover frequency, each time the frequency is halved the amplitude is reduced by half, or 6 dB. This will result in a constant-amplitnde characteristic for a constant-amplitude signal across the bead from the turnover frequency to the lowest frequency.

Because this type characteristic reduces the amplitude of the frequencies below the turnover frequency approximately 6 dB per octave, it is necessary to employ an equalizer in the reproducer circuit which has an inverse characteristic to the constant amplitude portion of the recording characteristic. (See Fig. 14-6B.)

14.7 What are the advantages of a constant-amplitude constant-velocity recording characteristic? — It permits a higher recording level at the higher frequencies because of the constant-amplitude characteristic below the turnover frequency; thus, the signal-to-noise ratio of the frequencies in the constant-velocity portion are increased.

If reproduced with a magnetic pickup, the frequency response above the turnover frequency is almost uniform. The use of a low-frequency equalizer below the turnover frequency having a rise of 6 dB per octave returns lower frequencies to a flat response. (See Fig. 14-6B.)

14.8 What is the turnover frequency of a magnetic cutting head?—A frequency where the characteristic of the cutting head departs from a constant-amplitude characteristic to one of constant velocity. In the past, the turnover frequency has varied with different manufacturers of cutting heads and with different recording activities. However, in June, 1953, the RIAA placed the turnover frequency at 1000 Hz. Records made before this date used turnover frequencies anywhere between 250 and 1200 Hz. As a rule, the frequency characteristic is specified on the record envelope.

14.9 What effect does the turnover frequency have on reproduction?—If the correct turnover frequency is not used during the reproduction, an improper balance is obtained between the frequencies lying in the constant-amplitude constant-velocity sections of the recording characteristic. It is highly important, when reproducing a record, that the reproducing circuits shall be capable of being adjusted to the correct turnover frequency.

The effect of using a turnover frequency that is too low and one that is too high is shown in Fig. 14-9. It will be noted that if the turnover frequency is too low (A), there is a step downwards in the frequency response. If the turnover frequency is too high (B), the low end starts rising too soon upsetting the balance between the high and low frequencies.

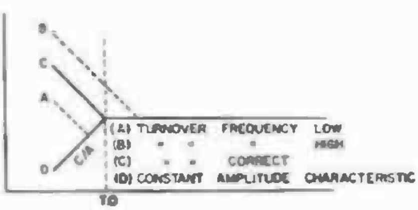


Fig. 14-9. The effect of correct and incorrect turnover frequency in a reproducing circuit. The effect when the turnover frequency is too low is shown at A. The effect when the turnover frequency is too high is shown at B. The correct turnover frequency is illustrated at C, and the constant-amplitude characteristic is given at D.

14.10 What are turnover frequencies used with early cutting heads?—Since the adoption of the RIAA characteristics, the turnover frequency has been standardized at 1000 Hz. (See Fig. 14.10.)

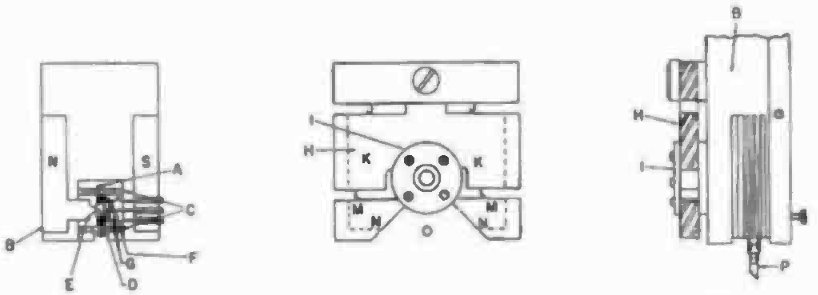
14.11 How are the damping springs and material applied to the mechanical construction of a magnetic cutting head?—Magnetic cutting heads designed for disc recording may be considered to be multisection mechanical bandpass filters terminated in an artificial line. The primary function of a cutting head is to engrave on the disc a faithful reproduction of the applied electrical waveforms.

The cutting head converts electrical energy into mechanical energy; therefore, it is similar to an electric motor or a loudspeaker. For present day requirements, a cutting head must be capable of recording with uniform characteristics up to at least 10,000 Hz. Many recording activities record to 15,000 Hz.

The interior construction for a typical monophonic balanced-armature cutting head is shown in Fig. 14-11. Refer to the front view shown at (a) in Fig. 14-11 for the following explanation. A balanced armature A is placed in a permanent magnetic field supplied by a permanent magnet B. At the upper and lower ends of the armature are placed actuating coils C connected in series. The use of two coils in series cancels the even harmonics and reduces the distortion. The armature A is constructed of laminated steel with a V-shaped saddle D at the center. The saddle rests on a ground knife-edge support E. Steel rod F balances the armature in the exact center of the gap

Columbia and Mercury	78 rpm	300 Hz
Columbia and Mercury	33⅓ rpm	300 Hz
RCA-Victor	78 rpm	500 Hz
Decca (FFRR)	78 rpm	300-500 Hz
Technicord	78 rpm	650 Hz
Vertical recording	33⅓ rpm	400 Hz
Miscellaneous	78 rpm	250-700-1000 Hz

Fig. 14-10. Turnover frequencies used with early cutting heads.



(a) *Front view.* (b) *Rear view.* (c) *Side view.*

Fig. 14-11. Interior view of typical monophonic balanced-armature cutting head, showing the damping springs and damping materials.

between the pole pieces. Two steel rods G with offset loops are used to supply pressure to the outer ends of the saddle mount D, to balance the armature in the center of the magnetic structure.

The rear view of the cutting head is shown at (b) in Fig. 14-11. In this view the rear end of the armature is shown connected to a thermoplastic damper H which removes resonant peaks and helps to smooth out the frequency response. This damping material is sometimes referred to as an artificial line. A metal washer I secures the damping member to the aruuature. A side view showing how the damping material is secured to the magnetic structure ap-appears at C in Fig. 14-11. The cutting-head just discussed is of the *nonnegative feedback* type.

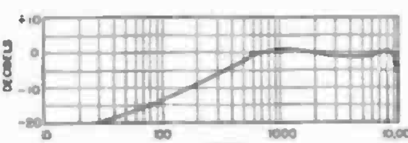


Fig. 14-12. Frequency response of a magnetic cutting head plotted from a light pattern.

14.12 How are the damping material and springs adjusted to secure the desired frequency response?—The adjustment of a cutting head to secnre a desired recording characteristic is both tedious and difficult and is somewhat of a cut-and-try job. The adjustment is started by centering the armature between the pole pieces by means of the balance springs G (shown in Fig. 14-11) while observing the contact between the knife-edge and the saddle on the armature under a microscope. When the contact is uniform, the springs are locked in place.

The next procedure is to adjust the various areas of the damping material which control the frequency response of the cutting head. This is done by removing sections of the damping material shown at (b) in Fig. 14-11. Areas J of the damping material affect the frequency response between 800 and 2000 Hz, while areas K affect the region between 3000 and 5000 Hz. Areas L may be used to control frequencies between 5000 and 6500 Hz. Areas M, N, and O control the frequencies above 6500 Hz.

The calibration should be carried out at normal room temperature, as temperature has a pronounced effect on the final response. The calibration is made by recording light patterns and measuring the response visually by playing back and measuring it electrically, or by means of an fm calibrator described in Question 14.49. A typical frequency response plotted from a light pattern is shown in Fig. 14-12. (See Question 23.75.)

14.13 Will a cutting head calibrated for 33⅓ rpm be satisfactory for 45 rpm?—It is possible to obtain satisfactory calibration for both speeds; however, if the cutting head is of the negative-feedback type, generally the amplifier system is equalized for 33⅓, 45, and 78.26 rpm and will be adjustable within the specified tolerances for the RIAA Standard reproducing response.

14.14 How does temperature affect the calibration of a magnetic cutting head?—If the cutting head is calibrated at normal room temperature and then operated at a low room temperature, the frequency characteristic will change. Certain manufacturers incorporate a small heating unit in the cutting head to maintain it at an even temperature regardless of the room temperature.

When making calibrations of a cutting head, single frequencies should not be applied to the head for long periods of time as the damping material becomes overheated and softens, permitting the calibration to change. If the head has been subjected to overheating, it should be permitted to cool slowly and then should be rechecked for calibration.

14.15 What is the material used for the damping member of a magnetic cutting head?—A thermoplastic called Viscoloid. In some instances tungsten-loaded rubber is used.

14.16 How is a monophonic signal (lateral cut) obtained with a stereophonic cutting head?—Identical signals having the same phase and level produce a lateral (monophonic) signal. Identical signals out-of-phase produce a vertical cut. (See Question 13.202.)

14.17 What is a mechanical transmission line used with magnetic cutting heads?—In the early Western Electric cutting heads, a rubber damping line, composed of three rubber tubes, one within the other, was used to damp out mechanical resonance of the moving parts. The rubber line was attached to the armature by means of a metal fin running the entire length of the rubber line. The mechanical transmission line reduced distortion besides smoothing out the frequency response. This system was used with the lateral coarse-pitch cutting heads, before the use of negative-feedback cutting heads.

14.18 Is the impedance of a magnetic cutting head constant?—No, it varies with frequency. As a rule, the magnetic cutting head will show its stated impedance only between 400 and 1000 Hz, then rise to several thousand ohms at the higher frequencies. Below 1000 Hz the impedance may drop to one-tenth that at 1000 Hz. A typical impedance curve for a 500-ohm head of the nonnegative feedback type is shown in Fig. 14-21A.

14.19 What are the standard values for cutting impedance?—There are *no standards* relative to cutting impedance; the values are optimal with the manufacturer. Typical impedances run from 5.6 ohms for the Westrex 3B stereo-cutting head, up to 600 ohms for others, but generally they are below 200 ohms.

14.20 What is the purpose of connecting an RC network in series with a cutting head?—To prevent the cutting head from loading down the driving amplifier at the lower frequencies because of the decreasing impedance as explained in Question 14.18. It also helps to maintain the 6 dB per octave dropoff in the constant-amplitude portion of the recording characteristic. (See Fig. 14-12.)

14.21 Explain the action of an RC network connected in series with a nonfeedback cutting head.—Fig. 14-21A shows the impedance characteristics for a typical nonfeedback magnetic cutting head, rated 500 ohms at 1000 Hz. As may be seen, this head has an impedance of 90 ohms at 40 Hz, 500 ohms at 1700 Hz, and then rises to over 3000 ohms at 10,000 Hz.

With the cutting head described in the foregoing connected across the output of a driving amplifier, the amplifier output sees 500 ohms at only one frequency (1700 Hz). This means the amplifier will be overloaded at the lower frequencies because of the lowered impedance of the cutting head decreasing

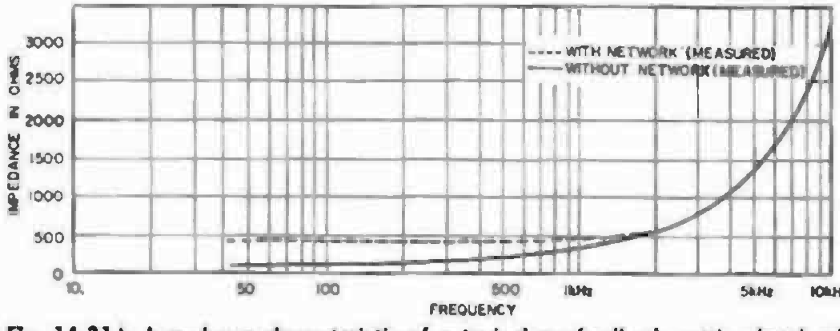
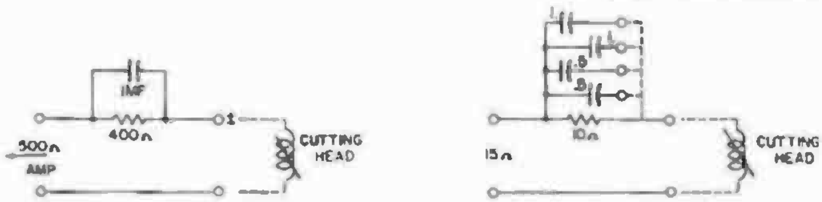


Fig. 14-21A. Impedance characteristic of a typical nonfeedback cutting head, with and without RC network.



(*a*) *500 ohms.* (*b*) *15 ohms.*

Fig. 14-21B. RC networks used with nonfeedback cutting heads to maintain a constant-amplitude, constant-velocity frequency characteristic.

the power output and increasing the harmonic distortion. The impedance rise above 1700 Hz will have little effect on the amplifier as it acts as a bridging load across the output. This statement only holds true if the output stage consists of triodes or pentodes using a considerable amount of negative feedback. To prevent the low impedance of the cutting head from disturbing tbe output stage of the driving amplifier, an RC network is connected in series with the output circuit, as shown in Fig. 14-21B.

An RC network for a cutting head rated 500 ohms impedance consists of a 400-ohm noninductive resistor and a 1-μF capacitor connected in parallel. The parallel impedance of this network is 148.4 ohms at 1000 Hz. When connected in series with the cutting head, the load impedance seen by the amplifier output is the network impedance plus the impedance of the cutting head as shown by the dotted line in Fig. 14-21A. The curves shown are actual impedance curves measured on a commercial cutting head. The variation in impedance between 100 and 1700 Hz is due to the head characteristics and is of little consequence to the over-all recording characteristic. With the RC network described in the circuit, the impedance, as may be seen, cannot fall below 400 ohms which appears in the region of 1000 Hz. The action of the network when connected in the circuit may be explained as follows:

At a frequency of 1700 Hz, the network in series with the impedance of the cutting head presents a load impedance to the output of the amplifier of 500 ohms. As the frequency decreases, the cutting-head impedance decreases. However, the reactance of the capacitor in the network is rising and, at a frequency of 40 Hz, the reactance of the capacitor has reached about 4000 ohms, which leaves practically pure resistance in the circuit.

It will be noted the load impedance below 1000 Hz never rises above 400 ohms. Above 1700 Hz, the reactance of the capacitor in the network drops very rapidly and effectively shorts out the resistor, leaving only the impedance of the cutting head in the circuit.

If such a network as is described in the foregoing is not used, high distortion may be expected at frequencies below 1000 Hz and the frequencies below the turnover frequency will not drop off at the rate of 6 dB per octave. The constants of the RC network will vary with cutting heads of different manufacture and impedance. The cor-

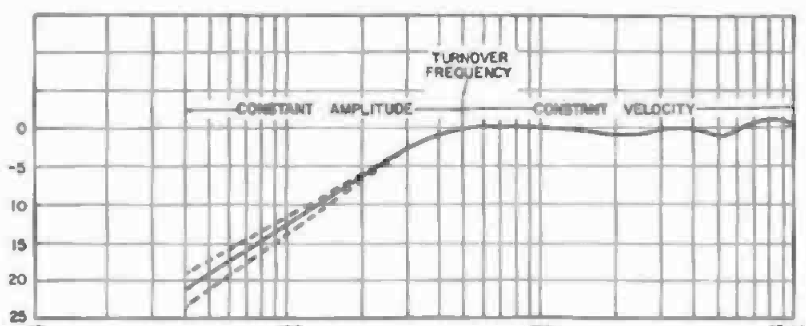


Fig. 14-21C. A typical response for a nonfeedback monophonic cutting head plotted from a light-pattern measurement.

rect value may be obtained experimentally by measuring the impedance of the cutting head and then designing the RC network. Measurements are then made with the cutting head and the network connected in series and plotted as shown in Fig. 14-21A.

Light patterns are also helpful in arriving at the final characteristic. A typical frequency characteristic plotted from a light pattern is shown in Fig. 14-12C. The recording of light patterns is discussed in Question 23.75.

14.22 What is the standard reference level and frequency used for cutting head calibration?—It is 7 centimeters per second for both coarse pitch and microgroove at speeds of 33⅓, 45, and 78.26 rpm. It is also the same for vertical recording. (See Fig. 14-22.)

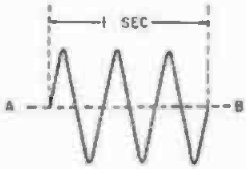


Fig. 14-22. Illustration of a standard reference level. The amplitude is such the stylus travels 7 centimeters from A to B in one second at a frequency of 1000 Hz.

14.23 How is the sensitivity of a cutting head stated?—Cutting heads are rated in watts input for a given velocity in centimeters per second, at a frequency of 1000 Hz. The average power required for driving a modern cutting head is approximately 1 watt (plus 30 dBm). This holds true for both monophonic or stereophonic heads.

14.24 How is the turnover frequency of a nonfeedback cutting head controlled?—By the resonant frequency of the mechanical system, the effective mass of the moving parts, and the stiffness of the centering springs.

Below the resonant frequency of the mechanical system, the cutter mechanism acts as a spring, resulting in a constant armature deflection for a constant input across the head. Thus, a constant-amplitude characteristic is obtained below the turnover frequency.

Above the resonant frequency of the mechanism the system is mass controlled and for a constant input, results in a constant-velocity characteristic. In this manner an overall constant-amplitude constant-velocity characteristic is obtained.

14.25 What limits the maximum amplitude swing of the armature of a cutting head?—The swing of the armature is limited by the mechanical structure and the damping mechanism. The maximum amplitude is also limited by the amount of power that may be safely applied to the actuating coils. If the power is too great, the coils will overheat the damping material and necessitate the recalibration of the head for both sensitivity and frequency response. Overmodulation also increases the noise and distortion.

14.26 What is a vertical cutting head?—A cutting head which engraves the sound track on a disc record in a vertical plane. The sound track varies in depth and width. A cross-sectional view of a negative-feedback cutter with the principal components indicated is shown in Fig. 14-26.

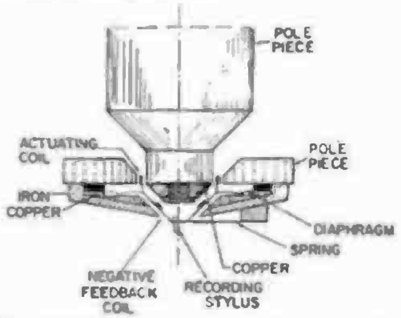


Fig. 14-26. Cross-sectional view of a vertical, negative-feedback, cutting head. (Courtesy, Bell Telephone Labs.)

14.27 What is the turnover frequency of a vertical cutting head?—Approximately 250 Hz for those not using negative feedback. For those employing feedback, the turnover frequency is 400 Hz.

14.28 What is the purpose of using negative feedback with a vertical cutting head?—To obtain a more uniform frequency response and reduce distortion, as the head mechanism has a rather sharp peak around 1000 Hz. Approximately 43 dB of negative feedback is required to obtain a uniform frequency response between 40 to 12,000 Hz. A special power amplifier is required.

14.29 What are the advantages of using negative feedback with a cutting

head?—The use of negative feedback with a cutting head adds approximately 16 dB to the signal-to-noise ratio at 10,000 Hz, at a diameter of 6 inches. Tests made on heads using negative feedback indicate a 69 dB signal-to-noise ratio, using a hot-stylus recording technique. A vibrating coil may be controlled over a wide range of frequencies by the application of negative feedback. The feedback eliminates the need for heavy mechanical damping materials, and reduces the effects of the disc material on the tip of the recording stylus. Negative feedback governs the power rather than absorbing it in the control process. Furthermore, the use of viscous damping offers difficulties as its effectiveness deteriorates over a period of time and its characteristics are affected by temperature and changes resulting from the use of a hot stylus.

Intermodulation distortion measurements made at a velocity of 8 centimeters per second show:

Distortion	*Frequency Range*
0.61%	40-2000 Hz
0.62%	60-2000 Hz
0.61%	100-2000 Hz

The test frequencies were set in a ratio of 4:1, the high frequency being 12 dB lower in amplitude than the lower frequency.

14.30 What is a crystal cutting head?—A cutting head which employs a piezoelectric crystal for the stylus actuating mechanism. The crystal is similar (except heavier) to that used in a crystal pickup. A typical crystal cutting head is shown in Fig. 14-30.

The audio currents cause the crystal to twist in a lateral direction. A recording stylus mounted in a chuck is attached to the crystal slabs for the purpose of engraving the sound track on the recording disc.

Ceramic crystals have to some extent replaced the Rochelle salt crystal; however, because of the greater sensitivity, Rochelle salt crystals find their greatest usage in pickups, microphones, and headphones. Generally, ceramic crystals are treated both electrically and mechanically much the same manner as Rochelle crystals. (See Question 25.103.)

14.31 What are the recording characteristics of a crystal cutting head?—Constant amplitude as shown in Fig. 14-5. If a constant voltage is applied to the crystal slabs, the amplitude motion of the recording stylus will be constant regardless of the applied frequencies. Recordings made with a crystal cutting head require no equalization in either the recording or reproduction circuits if played back with a crystal pickup.

14.32 What is the impedance of an average crystal cutting head at 1000 Hz?—Approximately 20,000 to 40,000 ohms.

14.33 How are crystal cutting heads coupled to the output of an amplifier stage?—Crystal cutting heads may be coupled to an amplifier output stage in a number of ways. The output stage may be single or double ended. The recording characteristic is controlled by

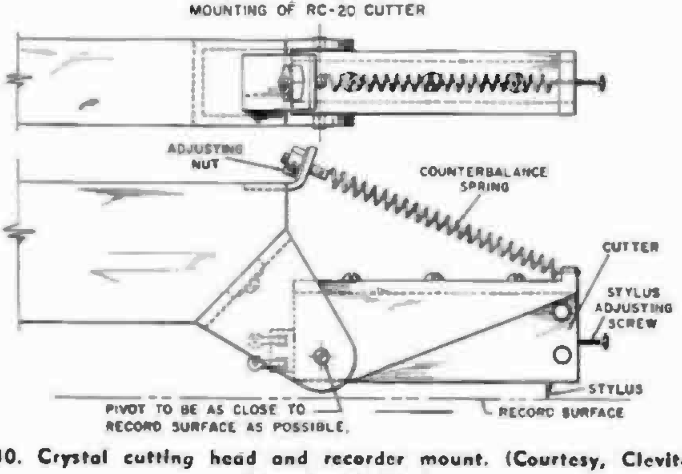


Fig. 14-30. Crystal cutting head and recorder mount. (Courtesy, Clevite Corp., Piezoelectric Div.)

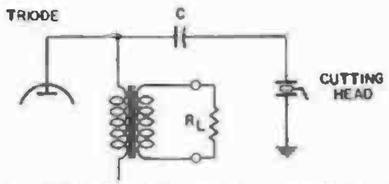


Fig. 14-33A. A coupling circuit for operating a crystal cutting head constant-amplitude. Plate load must not exceed 4000 ohms.

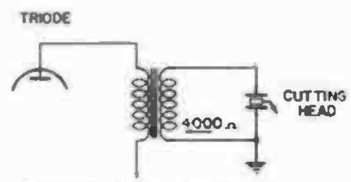


Fig. 14-33B. Method of driving a crystal cutting head as a constant-amplitude recording device.

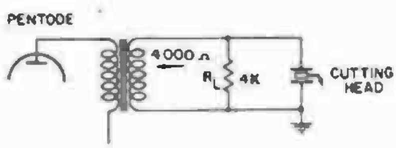


Fig. 14-33C. Crystal cutting head coupled to a pentode or beam-power tube for constant-amplitude recording. The terminating resistor R_L supplies the correct load to the cutting head.

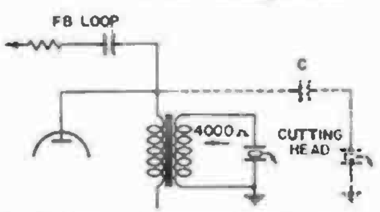


Fig. 14-33D. A crystal cutting head using constant-amplitude recording characteristics coupled to a pentode or beam-power tube using negative feedback.

the output impedance presented to the cutting head.

For constant-amplitude recording the head may be coupled as shown in Fig. 14-33A. Here the crystal is coupled to the output tube, a triode, through coupling capacitor C. In Fig. 14-33B, the cutting head is driven from an output winding, the terminating impedance being the plate resistance of the tube. The impedance presented to the crystal should fall between 3000 and 4000 ohms.

To operate a crystal cutting head from a pentode or beam-power tube without negative feedback, the coupling circuit shown in Fig. 14-33C is used. Beam-power tubes use a load impedance approximately one-fifth the plate resistance, and the distortion without feedback will rise quite sharply. The output transformer is terminated by a resistor R_L to reflect the correct load impedance to both the plate of the tube and the cutting head. For an output stage using negative feedback, the desired output impedance is supplied by the output transformer winding (Fig. 14-33D). It is important that the output impedance seen by the cutting head does not exceed 4000 ohms, including the output impedance as modified by the negative feedback. Coupling circuits for driving crystal cutting heads employing a constant-amplitude constant-velocity recording characteristic are discussed in Question 14.37.

14.34 What is the signal voltage required to drive a crystal cutting head? —For constant-amplitude recording, about 50 volts. For constant-amplitude constant-velocity, approximately 150 volts. This is the rms voltage as measured across the crystal with a vacuum-tube voltmeter.

14.35 What type load characteristic does a crystal cutting head present to the output circuit of an amplifier?—A crystal cutting head presents a capacity load in which the impedance decreases as the frequency increases. For this reason, it is recommended that only class-A triode amplifiers or beam-power amplifiers using at least 12 dB of negative feedback be used for driving crystal cutting heads.

Since a crystal cutter is of the actuated type, the stylus displacement (amplitude) is proportioual to the impressed voltage over practically its entire operating range. Because of this characteristic, constant-amplitude records may be cut without equalization in the recording circuits.

Commercial constant-amplitude constant-velocity recordings may be cut by the use of special coupling circuits which are explained in Questions 14.37 and 14.38.

14.36 What is the internal capacitance of a crystal cutting head?—Approximately 0.007 to 0.10 μF.

14.37 How are crystal cutting heads coupled to an amplifier for a constant-

amplitude constant-velocity recording characteristic? — As shown in Fig. 14-37A. When the crystal impedance is equal to the impedance of the coupling circuit, the freqnency response will be down 3 dB at the frequency where the two impedances are equal. Above this frequency (the turnover frequency) the frequency response will fall off at a rate of 6 dB per octave. In other words, the cutting head will operate on a constant-velocity basis above the turnover frequency and on a constant-amplitude basis below the turnover frequency.

By the selection of the proper component values for the coupling circuit, the turnover frequency can be placed anywhere in the normal frequency range. If the turnover frequency is established between 800 to 1000 Hz, the cutting head will record a commercial constant - amplitude constant - velocity characteristic (also called a modified constant-velocity characteristic). If the turnover frequency is placed at the upper extreme of the frequency spectrum, the cutting head will record a constant-amplitude characteristic as shown by the line at zero amplitude in Fig. 14-37B.

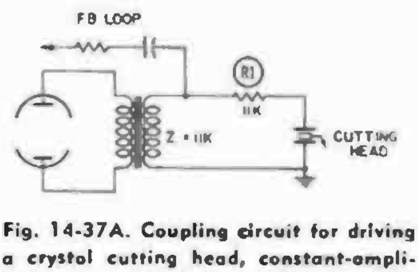


Fig. 14-37A. Coupling circuit for driving a crystal cutting head, constant-amplitude, constant-velocity with a turnover frequency of 1000 Hz.

14.38 How are the coupling circuit values selected for a constant-amplitude constant-velocity recording characteristic?—By the impedance of the coupling circuit and the desired turnover frequency. The frequency response of a crystal cutting head coupled for three different turnover frequencies (250, 500, and 800 Hz) is shown in Fig. 14-37B. The flat portion of the curve is the constant-amplitude section and the right-hand portion the constant-velocity section.

A graph that may be used for selecting the circuit constants for a particular turnover frequency is given in Fig. 14-38. Assume a turnover frequency of 1000 Hz is desired. Following the 1000 Hz line upward to where it intersects the diagonal line then reading at the left, it will be noted a source impedance (output) of 22,000 ohms is required. This impedance is divided equally between the output winding and the series resistor R1, or 11,000 ohms each. As this is rather an odd value of output

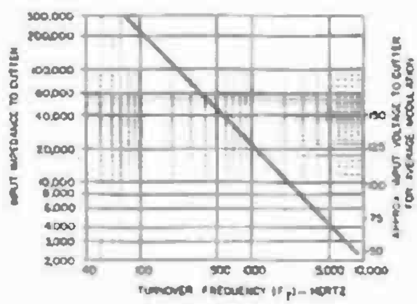


Fig. 14-38. Graph for selecting circuit constants to operate a crystal cutting head constant velocity, or modified constant velocity. (Courtesy, Clevite Corp., Piezoelectric Div.)

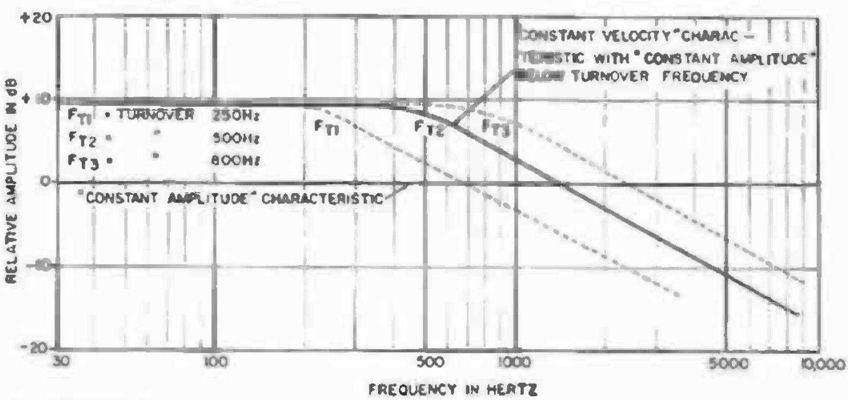


Fig. 14-37B. Frequency characteristics of a crystal cutting head operating constant amplitude and constant velocity. (Courtesy, Clevite Corp., Piezoelectric Div.)

impedance to obtain from an output transformer, a matching transformer may be connected between the regular output transformer and the cutting head with the proper impedance ratio. A 10-percent variation in impedance match is acceptable. The series resistor R1 should never be smaller in value than the output impedance of the coupling transformer.

It will also be noted in Fig. 14-37B that frequencies below the turnover frequency are recorded approximately 10 dB higher in amplitude than for the constant-amplitude characteristic. Also, frequencies above turnover frequency fall off at a rate of 6 dB per octave.

14.39 Why is a lower recording level used for constant-amplitude recording?—To provide easier tracking for the pickup stylus when reproducing. Constant-amplitude recordings can be cut at a higher level, if the high frequency response is restricted to a frequency below 9000 Hz. If a constant-amplitude recording is cut at the same level used for constant-amplitude constant-velocity recording, difficulty may be experienced in obtaining satisfactory tracking at the high frequencies during reproduction.

14.40 Can hot-stylus techniques be used with a crystal cutting head?—No, the heat would damage the crystal. (See Question 14.14.) The techniques of hot-stylus operation are discussed in Questions 15.58 to 15.71.

14.41 What precautions should be taken to prevent damage to a crystal cutting head?—Direct current should never be permitted to come in contact with the crystals. Sudden high-level signals of sine-wave character should not be applied to the head when testing, as the crystal may be shattered. Bring the level up gradually.

14.42 Why are magnetic cutting heads preferred to crystals for commercial recording?—Because magnetic recording heads are more rugged in their construction and will hold their frequency characteristics over long periods of time. Crystal cutting heads are affected by both temperature and humidity. Also, they may be damaged quite easily if the stylus is struck against the side of the turntable.

Because the impedance of a magnetic cutting head is generally on the order of 500 to 600 ohms, it may be operated at a considerable distance from its driving amplifier. Crystal cutting heads, being high-impedance devices, require that they be operated fairly close to the driving amplifier.

14.43 What is the composition of a ceramic crystal? — Ceramic elements used in transducer manufacture are usually ammonium dihydrogen phosphate, lithium sulphate, barium titanate, or several variations of lead zirconate and lead titanate ceramics. The ceramics are polycrystalline in nature and do not have piezoelectric properties in their original state. The magnitude and character of the piezoelectric effect is greatly dependent on the orientation of the applied force or the electric field, with respect to the axes of the material. Slabs are cut from the crystal, rather than from the whole crystal itself. They are cut in a similar manner to quartz crystals for radio use, as described in Questions 25.103 and 25.191.

14.44 What Is an embossing head?—A magnetic or crystal cutting head similar in construction to the cutting head used for standard recording. They may be operated at constant amplitude, or constant amplitude, constant velocity. The stylus is dull-pointed and does not remove any material from the record as it records. The sound track is indented in the recording media and record material displaced rather than removed.

Embossed records are used in dictating machines and similar devices. A typical embossing recorder is shown in Fig. 13-22.

14.45 What is the relationship between the pressure wave at a microphone and the stylus motion?—For a constant-amplitude recording, a constant pressure at the microphone produces a constant-amplitude motion at the recording stylus, if the frequency characteristics of the recording system are uniform. For constant-amplitude constant-velocity recording, the amplitude of the undulations in the record are inversely proportional to the frequency in the constant-velocity range. Below the turnover frequency, they are constant amplitude or follow the frequency characteristic of the cutting head.

14.46 Describe the different methods used for calibrating a cutting head.—The frequency response can be measured by several different methods, by applying a constant voltage to the head and recording a series of frequencies prodncing a light pattern. This pattern is then measured electrically by playing it back or photographing the image and calculating the variations in response. Another method used rarely is the deflection of the stylus tip in a special fixture having a calibrated eye piece or by use of a frequency-modulation calibrator, described in Question 14.49. The preferred method is the light pattern electrically measured. The recording of light patterns is discussed in Question 23.75.

14.47 What is the formula for calculating the frequency response of a light pattern?—The measurements are referred to a reference frequency, generally 1000 Hz. The ratio of the other frequencies to the reference frequency is:

$$20 \operatorname{Log}_{10} \frac{F_1}{F_2}$$

where,

F_1 is the reference frequency,
F_2 is any other frequency.

It should be mentioned the electrical response from a light pattern is not quite the same as for an optical measurement of the same pattern. Generally the light pattern is used to measure the high-frequency response, and the electrical response of the pattern for the low frequencies.

14.48 How is the sensitivity of several cutting heads adjusted for uniform sensitivity?—Assnme the several cutting heads are similar in design and have approximately the same frequency characteristics. An unmodulated groove is cut and the width of the cut is measured and adjusted to normal. A 1000-Hz test signal is applied to the head under test and the gain of the driving amplifier adjusted to produce a 7-centimeter per second cut. The exact power to the cutting head is then noted for future reference.

Each cutting head is measured in a similar manner. The gain of each cutting-head driving amplifier is adjusted to produce the same percentage of modulation within plus or minus 0.5 dB. Thus, compensation is made for the difference in sensitivity for each individual cutting head. After the power level has been established for each head the recording of a test signal may be dispensed with, and only the output power level of the driving amplifier need be measured.

If the cutting heads employ internal heaters, sufficient time must be permitted for the head to come up to its normal operating temperature. Such procedures should be a part of the daily operating routine.

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Section 15

Recording and Reproducing Styli

It has been said that the entire evolution of sound reproduction can be divided into two parts, (1) learning how to make a sound, and (2) the reduction of distortion. This is a thumbnail sketch of the history of recording and reproducing styli. Imperfections induced by either the recording or reproducing stylus will be manifest as distortion. In the 1920s, a cactus or bamboo phonograph reproducing needle was very popular, as it failed to reproduce the distortion of the higher frequencies, as heard when a steel or tungsten needle was used. Work done by Frank and Isabel Capps, Charles Davis, and C. J. LeBel greatly improved the status case of the stylus.

This section covers the materials in general use in the manufacture of styli, geometry involved, stylus pressures, stylus heating-coil design and use, characteristics, and measurements.

15.1 What is a recording stylus?—A sharp-pointed, gouge-shaped instrument for engraving a sound track on a cylindrical or disc record. The tip may be of precious or semiprecious stone, or metal. Commercial recording styli use sapphire tips. (See Fig. 15-1.)

15.2 Describe a reproducer stylus?—A stylus somewhat similar to that used for recording, except the tip is not sharp, but conical. The tip is designed to fit the side walls of the groove but clear the bottom of the groove. The tip may be a sapphire, a ruby, or a diamond. (See Fig. 15-2.)

15.3 What are the essential parts of a recording stylus and the basic principles of manufacture?—The recording stylus is one, if not the most important, component in a recording system, and as far as is known was first used by Edison in 1877. With the advent of present-day high-quality sound, wide frequency range, low distortion, increased signal-to-noise ratio, long play and stereophonic recordings, and improved recording media, a great responsibility is placed on both the recording and reproducing styli. Since its first use, many different materials have been use for their manufacture—steel alloys, sapphire, and diamond. However, over the

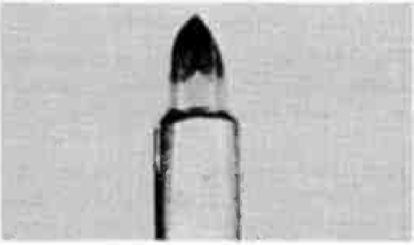
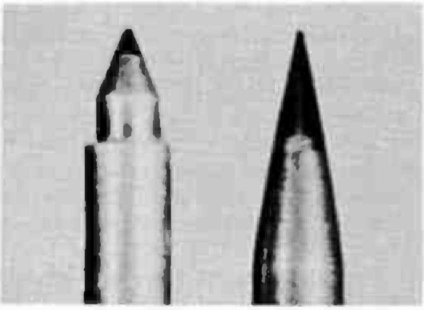


Fig. 15-1. A coarse-pitch recording stylus with 87-degree included angle, 1.5-mil tip radius.

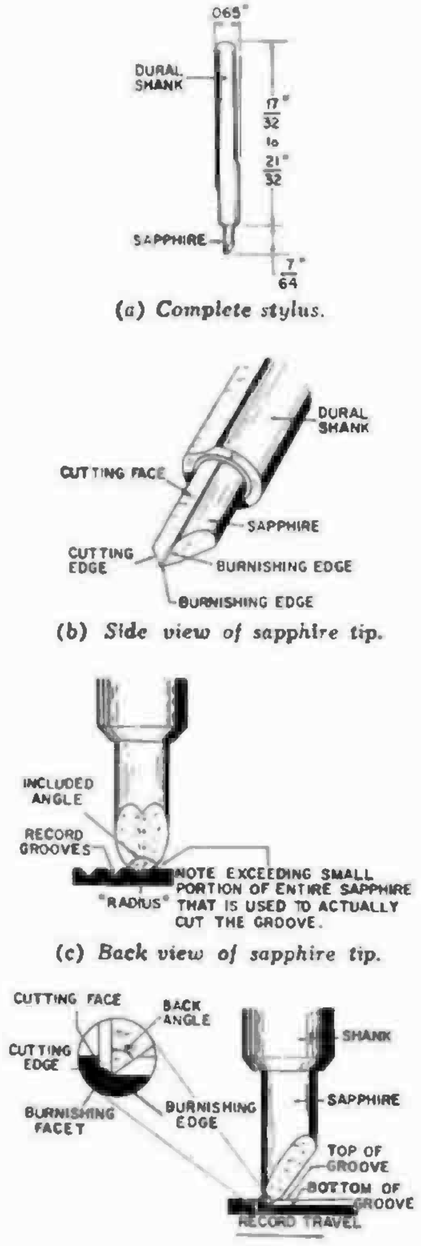


(a) 56-degree ruby, dural shank. (b) 26-degree sapphire, brass shank.

Fig. 15-2. Coarse-pitch reproducing styli for acetate disc.

years it has been found that *corundum* is the most practical of all material for recording styli. Corundum can be obtained in several different colors, among them are: red, blue, green, violet, yellow, and colorless, which is many times used for identification. Both the natural and synthetic sapphire (blue, ruby red, etc.) are corundum (AL_2O_3) of a hexagonal crystalline structure with a hardness of 9 on the Moh scale and 1525 to 2000 on a Knoop scale. Because of its lack of grain, crystalline structure, and cleavage, sapphire may be ground to very accurate dimensions and angles, while still retaining a very fine cutting edge. This latter property is of prime importance in the manufacture of recording styli. Although the diamond is much harder than corundum for use in recording styli, it is impractical because of its grain and internal stresses, as well as its cost. However, as a reproducing stylus, it is ideal because of its long life and ruggedness. As a recording stylus, the sapphire will outlast a diamond and produce superior recordings.

Several views of a recording stylus are shown in Fig. 15-3. At part (a) is shown a complete stylus. The stylus consists of a sapphire held in a dural shank, shown at part (b). The sapphire tip has a flat face ground on one side. The end of the sapphire is ground to a point with a rounded tip. Extending from the tip upward and along its edges are burnishing facets, and it is a most difficult and exacting part of the stylus grinding, as the dimensions are only a few ten-thousandths of an inch. The *burnishing facets* polish the groove as it is cut. The length of the burnishing facet not only affects the signal-to-noise ratio, but also the frequency response. A cold-stylus coarse-pitch recording with a burnishing facet of 0.4 to 0.6 mil will effect a loss at 10,000 Hz, compared to 1000 Hz of 3 to 6 dB. It will be noted at part (c) only a very small portion of the stylus tip is used to engrave the sound track on the record. An enlarged view of the burnishing facet is shown at part (d).



(a) Complete stylus.

(b) Side view of sapphire tip.

(c) Back view of sapphire tip.

(d) Enlarged view of tip, showing the burnishing facet.

Fig. 15-3. Essential parts of a recording stylus. (Courtesy, Electronic Industries and Frank L. Capps Co.)

For high-quality recording of any type, the noise level of the stylus itself must be at least 55 dB, and as a rule if not abused, will run 57 to 60 dB below the reference level of 7 centimeters at a frequency of 1000 Hz. The noise level is measured while playing back an unmodulated groove. It is important that the unmodulated groove used for measurement is normal in every respect, that the chip has cleared the groove properly, that the stylus heat current is optimum, and that the disc is of good quality. (See Question 13.30.)

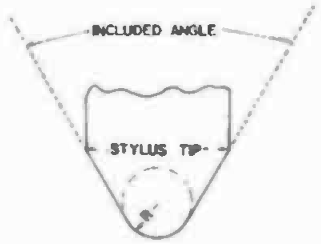


Fig. 15-4. Tip radius of a recording stylus.

15.4 What is the tip radius of a stylus?—The radius formed by the tip of the recording jewel as shown in Fig. 15-4.

15.5 What is the recommended tip radius for recording styli?

For microgroove recording at 16⅔, 33⅓, and 45 rpm, 0.00025 inch or less.

For coarse-pitch 33⅓-rpm recording, 0.0015 inch (16-inch transcriptions).

For coarse-pitch 78-rpm recording, 0.0015 inch.

15.6 What is the included angle of a stylus?—The angle formed by the two sides of the stylus tip. (See Fig. 15-4.)

15.7 Which way does the flat side of a recording stylus face?—Against the direction of motion of the recording blank as shown at part (d) in Fig. 15-3.

15.8 What is a burnishing facet on a recording stylus?—A small flat surface ground on the sides of the stylus tip and used for polishing the groove as it cuts. (See Fig. 15-3.)

15.9 What effect do the burnishing facets have on the frequency response?—If the dimension of the facet is kept to 0.15 mil, it has little effect on the frequency response. However, a facet of 0.5 mil will attenuate 8000 Hz 3 dB and 10,0000 Hz from 3 to 6 dB. Increasing the facet dimensions to 0.60 mil will attenuate 10,000 Hz 6 dB (cold stylus).

15.10 What does the term "mechanical bias" mean when applied to a recording stylus?—For coarse-pitch recording, the stylus is given a slight mechanical twist, relative to the shank, generally about 1 degree. The purpose of the mechanical bias is to throw the chip toward the center of the record as the groove is cut, and thus prevent fouling of the stylus tip. Mechanical bias is not used in stereophonic recording.

15.11 What are the included angles used with coarse-pitch recording?—The finished groove must have an included angle of 88 degrees plus or minus 5 degrees. The top of the groove shall not be less than 4 mils in width for reproducing with a 2.3-mil stylus and not less than 2.0 mils for records to be reproduced with a 1-mil stylus. The bottom radius is 1.5 mils.

15.12 What is the included angle for vertical-recording styli?—The finished groove shall have an included angle of 88 degrees plus or minus 5 degrees, with a top width of not less than 4.0 mils. The bottom radius is 2.0 to 2.3 mils.

15.13 What is the bottom radius of a coarse-pitch lateral groove?—For less than 136 lines per inch, 1.5 mils. For more than 136 lines per inch, 0.25 mil.

15.14 What is the bottom radius for a vertical-cut groove?—From 2.0 to 2.3 mils.

15.15 What is a master contour?—A magnified shadow of a stylus of known size and shape for comparing finished styli. Used in the manufacture of recording and reproducing styli.

15.16 Are diamonds used for recording styli tips?—No. They are used only in reproducer styli.

15.17 What is an "ANM" recording stylus?—A special antinoise modulation stylus (ANM) developed by Capps and Co. to reduce modulation noise while recording. The stylus is ground with two or more burnishing facets compared to the single facet of the conventional recording stylus. A cross-sectional view of a conventional recording stylus with a single burnishing facet in an unmodulated groove is shown at (a) Fig. 15-17.

At (b) in Fig. 15-17, the stylus is shown in a groove with a 20-degree slope and at (c) with a 30-degree slope. As will be observed, the burnishing angle made by the facet is increased depending on the stylus motion. When the modulation is high, the burnishing facets cannot function properly and leave rough places in the groove, (d) of Fig. 15-17, resulting in increased noise. This type noise is particularly noticeable because the rough spots are on the side of the groove which has the greatest effect on the stylus.

An "ANM" stylus has three burnishing facets on each side of the tip as

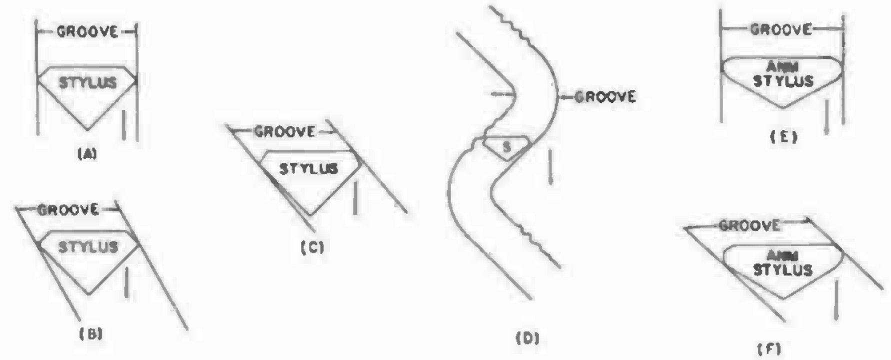


Fig. 15-17. Single and multiburnishing facet recording styli. (a) Single facet—250 degree angle to groove. (b) Single facet—45 degree angle to groove. (c) Single facet—55 degree angle to groove. (d) Facets causing noise patches. (e) ANM stylus with three burnishing facets. (f) ANM stylus with groove slope of 45 degrees. (Courtesy, Audio)

shown at part (e) in Fig. 15-17. It will be observed at part (f) in Fig. 15-17 the additional facets permit the groove to be polished even at high percentages of modulation. The angles of the facets are: leading facet 60 degrees, center facet 25 degrees, and last facet 10 degrees.

ANM styli are used for both standard and microgroove recording. Frequencies up to 20,000 Hz have been successfully recorded at speeds of 78.26 rpm. This stylus cuts a V-shaped groove rather than the rounded-bottom groove. ANM styli are used particularly for microgroove recording using hot-stylus techniques. (See Questions 15.60 to 15.71.)

15.18 What is the standard for reproducer styli included angles?—

For coarse-pitch, direct-acetate playback 33⅓ rpm, 26 degrees.

For coarse-pitch pressings, 33⅓ and 78.26 rpm, 40 to 55 degrees.

For microgroove monophonic 45 rpm, 40 to 55 degrees.

For microgroove stereophonic 33⅓ rpm, 40 to 55 degrees.

15.19 What is the standard tip radius for reproducing styli?—

For coarse-pitch 33⅓ and 78.26 rpm, 2.5 mils plus-minus 0.1 mil.

For coarse-pitch vertical 33⅓ rpm, 2.3 mils plus-minus 0.2 mil.

For microgroove 33⅓ rpm, 0.5 to 0.7 mil. (monophonic and stereo).

For microgroove 16⅔ rpm, 0.3 to 1.0 mil.

15.20 What are the recommended vertical pressures for reproducing styli?—Each manufacturer has a given pressure that is best for their particular type pickup. In the absence of such data, the user may be guided by the following pressures.

For a tip radius of 2.5 to 3.5 mils, 10 to 14 grams.

For a tip radius of 1.0 mil, 1 to 6 grams.

For a tip radius of 0.5 to 0.7 mil, 0.75 to 1.5 grams.

15.21 What is the relationship of a reproducer stylus tip radius to the modulation radius?—For the lowest distortion, the radius of the reproducer stylus at the point of contact with the groove must be less than the radius of the modulation.

15.22 What is the relationship between the included angles of recording and reproducing styli?—The relationship is as shown in Fig. 15-22. It will be noted that, although the included angle of the reproducing stylus

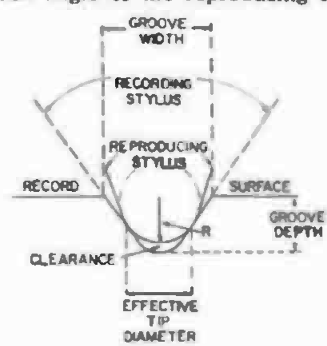
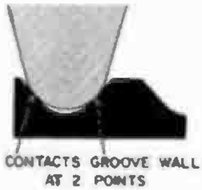
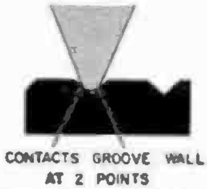


Fig. 15-22. Cross-sectional view of record groove, showing the included angle of recording and reproducing styli, and how they fit the groove.



(a) *Coarse-pitch, 2.5-mil radius.*



(b) *Microgroove, 1-mil radius.*

Fig. 15-23. Cross-sectional view of reproducer stylus in a record groove.

is less than the recording stylus, it contacts the groove side walls at a point where the tip clears the bottom of the groove. This is made possible because the tip radius of the reproducer stylus is greater than the tip radius of the recording stylus. (See Question 13.104 and 13.180.)

15.23 How does a reproducer stylus contact the side walls of a record groove?—Fig. 15-23 shows how a coarse-pitch stylus and a microgroove stylus fit the walls of the record groove. It will be noted the stylus sides contact the groove walls at only one point and that a considerable clearance exists between the stylus tip and the bottom of the groove. This is an ideal fit and results in a minimum of distortion.

15.24 What is the pressure developed at the point of contact between the stylus tip and the record groove side wall during reproduction?—It has been stated that the pressure of the reproducer stylus tip on the side walls of the record groove will run to as high as 50,000 pounds per square inch using the conventional methods of calculating stylus pressure. This would tend to create the impression that the reproduction of a fine groove record (microgroove, long playing) would be impossible. However, this is not the case as evidenced by the present-day use of such records.

To properly evaluate the pressure the stylus applies to the groove wall is quite involved and must take into consideration many factors often overlooked. Among these factors are: the area of the stylus tip contact, tip radius,

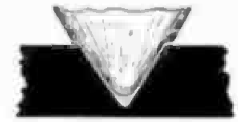


Fig. 15-25. How a worn reproducer stylus fits in a record groove.

included angle of the tip, deformation of the record wall, stylus pressure (vertical), recovery time of the record plastic after deformation by the stylus, the time of contact for a given area, and the heat generated by the friction of the stylus against the groove wall. From the foregoing it may be seen that the calculation is not a simple one. Calculations encompassing the foregoing factors indicate for a stylus of 1-mil tip radius, the pressure may run from 6000 pounds per square inch to 16,000 pounds per square inch.

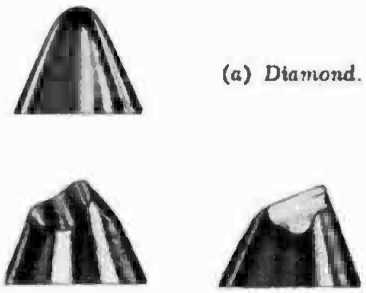
15.25 How does a worn stylus fit a record groove?—Fig. 15-25 shows a worn stylus in a normal record groove. As will be seen, the flat sides of the stylus lay against the groove walls and ride on top of the groove. This condition results in distortion, noisy reproduction, and rapid deterioration of the groove.

15.26 What effect does a worn stylus have on the record groove?—Small portions of the modulations in the groove are removed each time the record is played. The material thus removed acts as an abrasive and adds to the wear of both the stylus and the groove.

15.27 What does the reproduction from a worn stylus sound like?—The reproduction is fuzzy with rather high distortion of the higher frequencies. This is particularly true at the smaller diameters where the groove velocity is the lowest.

15.28 What is the average life of a sapphire reproducing stylus?—Wear tests made by Weiler indicate about 30 hours, depending on the quality of the disc and the sapphire. The wear is chiefly caused by the abrasive action of the dust from the stylus itself. Typical wear patterns of sapphire styli are shown in Figs. 15-28A and B. Vertical pressure was increased purposely to spread the wear.

15.29 How does the wear of a sapphire stylus compare to a diamond stylus?—It has been noted by Weiler that it takes 3½ hours to wear a 0.00075-



(a) *Diamond.*

(b) *Sapphire.* (c) *Sapphire (side view).*

Fig. 15-28A. Coarse-pitch 2.5-mil styli after 1000 hours of play using 1.5-ounce pressure to speed up the wear. (Courtesy, Audio Engineering Society)

inch flat in a sapphire stylus. To wear the same flat in a diamond stylus takes approximately 143 hours of playing.

15.30 What is the difference in pressure and friction for a new and a used stylus?—When a reproducer stylus is new and its shape unaltered, the pressure and friction are the greatest. As a flat is worn in the sides of the stylus tip, the pressure is reduced per unit area. The frictional heat is reduced for the same reason. Thus, the greatest wear to the record groove occurs when the record is played the first few times with a new stylus.

(a) *Diamond.* (b) *Sapphire.*

Fig. 15-28B. 0.7-mil styli after 15 plays, 8-grams pressure to speed up wear. (Courtesy, Audio Engineering Society)

15.31 What is the temperature rise created by the friction between the stylus tip and the record groove?—Temperatures of 2000 degrees Fahrenheit are possible when the stylus is new. As the stylus point wears, the temperature drops and when at its lowest is still in the vicinity of 1000 degrees Fahrenheit.

15.32 What is an osmium alloy stylus?—A stylus made of a hard dense metal alloy with a high degree of resistance to wear. This alloy was used for reproducer styli only. Now obsolete.

15.33 What type reproducing stylus is recommended for the lowest distortion and longest life?—A diamond-tipped stylus. As the diamond has a life many times greater than a sapphire or ruby, less damage is done to the record because of less wear to the stylus tip. Most high-quality reproducers are equipped with a diamond-tipped stylus. Contrary to general belief, a diamond-tipped stylus does not decrease the life of the record groove.

15.34 Define the term "compliance."—It is the ratio of displacement of the stylus to the applied force, expressed in centimeters per dyne. For a high quality pickup, this will range from 9×10^{-6} to about 35×10^{-6}. The term compliance refers to the lateral or vertical motion of the stylus or its compliance with the modulations in the record groove and is the *opposite* of stiffness. For stereophonic pickups the compliance in the vertical direction is made *slightly less* than in the lateral direction. However, if made too light in the vertical direction, the stylus may not be able to support the pickup at the desired stylus pressure. For a stylus pressure of 1 gram, the pressure is equal to 1000 dynes.

15.35 What is the effective mass of a stylus?—The lumped mass at the stylus tip that would result in the same mechanical impedance as caused by the moving elements of the pickup structure.

15.36 What is stylus force?—It is the vertical or downward force the pickup applies to the record groove, expressed in grams. Present-day pickups employ a stylus force or vertical pressure of 0.75 to 6 grams. Older type pickups measured their vertical pressure in ounces, some types taking up to four ounces to properly track a groove.

15.37 How is stylus force measured?—By playing back a constant-frequency record of at least 15 centimeters per second velocity, and measuring the distortion, or observing the waveform on an oscilloscope. The stylus force is adjusted for the lowest distortion consistent with the pressure as measured at the recorded surface.

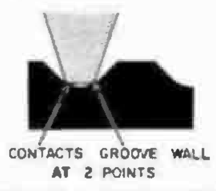
15.38 What effect does the friction of the stylus have on the side walls of the record groove?—A small amount of side thrust is acceptable and unavoidable. If the thrust is too great, it tends to pull the stylus out of the groove, particularly at the smaller diameters.

15.39 ***How are styli coded for identification?***—By a color code similar to that used for identifying resistors.

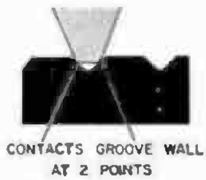
15.40 ***Of what is the dirt in a record groove chiefly composed?***—Silica particles, stylus material, soot, grit, and particles of the record groove mixed with lint and flocculated fibres.

15.41 ***What is a truncated reproducer stylus and its purpose?***—A reproducer stylus shaped as shown in Fig. 15-41 and designed to be used with either coarse-pitch or microgroove recording. It will be noted at part (a) that when used in a standard groove the flat tip rides on the curvature of the groove leaving a small clearance between the bottom of the groove and the stylus tip. At part (b) in Fig. 15-41 the stylus is shown riding in a microgroove record. Here it will be noted the sides of the tip ride near the upper limits of the modulations at the top of the groove. Referring back to part (a), if the groove is coarse pitch and not worn, the stylus tip will ride as shown; however, if the groove is worn the stylus will ride in the bottom of the groove and the signal-to-noise ratio will be considerably reduced. At best, such a stylus is a compromise. The pressure of the stylus is increased to 9 grams to prevent it from skating in the bottom of the groove (now obsolete).

15.42 ***What is a compromise stylus?***—When microgroove recording first made its appearance, a stylus was designed to provide the user with a means of playing back both coarse-pitch and microgroove recordings, and was known as a compromise stylus. *It is now obsolete.* A cross-sectional view of a compromise stylus for a coarse-pitch and microgroove recording is shown in Fig. 15-42.

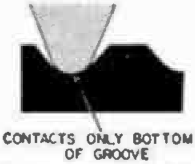


(a) *Stylus in a coarse-pitch groove.*

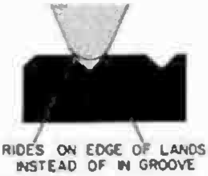


(b) *Stylus in a microgroove.*

Fig. 15-41. Compromise or truncated reproducing stylus (now obsolete).



(a) *In coarse-pitch groove.*



(b) *In microgroove.*

Fig. 15-42. Compromise styli.

15.43 ***How are recording styli tested for noise?***—By recording a series of unmodulated grooves and measuring the noise at a normal playback level. This subject is discussed further in Question 23.78.

15.44 ***What is a shadowgraph?***—A device for projecting the shadow images of styli. The image of the stylus is magnified 25 to 100 times its normal size. Variations in manufacture may be noted by rotating the image against a master contour.

15.45 ***Describe the difference between a long and a short-shank recording stylus.***—Recording and reproducing styli are manufactnred with either long or short shanks. The reason for the difference is the design of the recording head. Some recording heads require a longer shank because of the distance the stylus is projected into the armature tip. It is extremely important that the proper length stylus be used in a cutting head; if not used, the frequency response may be seriously affected. (See Question 15.46.)

15.46 ***Describe the effect of shank shape on the frequency response of a recording stylus.***—Unless the stylus has been specifically designed for a given type cutting head, the frequency response can be seriously altered by the shape and length of the material of the stylus shank. The long shank has a

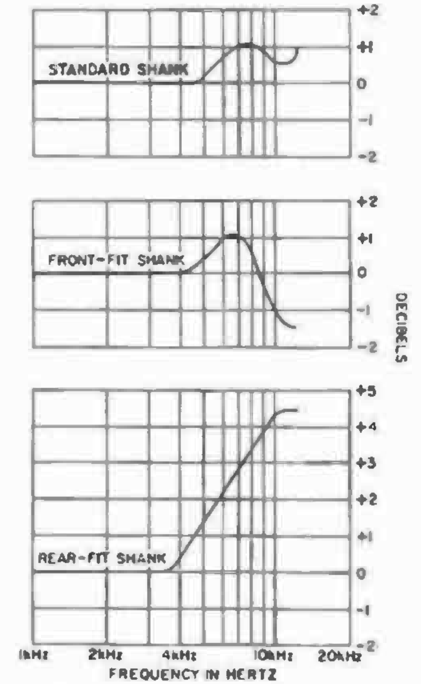


Fig. 15-46A. Frequency response of recording styli using different shank tapers. (Courtesy, Capps & Co., Inc.)

greater amplitude swing but *less* high-frequency response, as compared to the shorter shank. The material of which the shank is made also affects the response. Frequency response curves for three styli of the same manufacture and length, but of different shape are given in Fig. 14-46A. The shapes are shown in Fig. 15-46B. (See Question 15.3.)

15.47 Do recording styli of the same type and manufacture have the same frequency characteristic?—No, there can be considerable variation as may be seen by the plot of eight different styli of the same manufacture and type (Fig. 15-47). This is also true for coarse-pitch

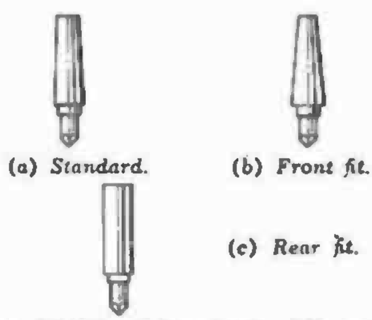


Fig. 15-46B. Stylus shanks. (Courtesy, Capps & Co., Inc.)

reproducing styli. For stereophonic, both the shape and length of the shank have considerable effect, as shown in the plots in Fig. 15-46A. The frequency response of each recording styli should be measured before using and graded.

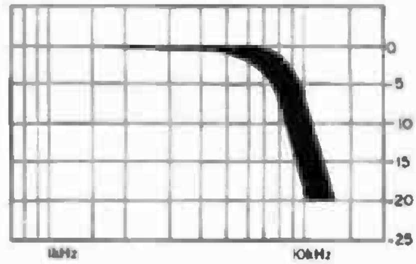


Fig. 15-47. Frequency response for a group of coarse-pitch recording styli of the same manufacture and type.

15.48 Can recording sapphires be resharpened?—Yes. There are several companies which make a specialty of such work.

15.49 Can reproducer styli be resharpened?—Yes, but the cost compared to that of new styli does not warrant it.

15.50 What is a boule?—A carrot-shaped lump of material used in the manufacture of commercial sapphires. (See Fig. 15-50.)

15.51 What is a jewel?—A sapphire or ruby used in recording and reproducer styli.

15.52 What is Stellite?—A commercial trade name for a metallic alloy used for recording styli.

15.53 What type sapphire is used for recording styli?—Commercial sapphire is manufactured from synthetic corundum. For all practical purposes, synthetic sapphires are identical in physical and chemical composition to the natural mineral. The name sapphire is used to identify clear synthetic corundum and differentiate it from the more familiar blue variety of this mineral. Synthetic white sapphire is a single, homogeneous crystal which can be given an exceptionally smooth surface polish by either flame or mechanical processes. There are no interruptions on their mirrorlike surfaces and no potential weak spots such as might develop in multicrystalline structures to offer locations for wear. Typical sapphire and ruby rods used in the manufacture of styli are shown in Fig. 15-53.

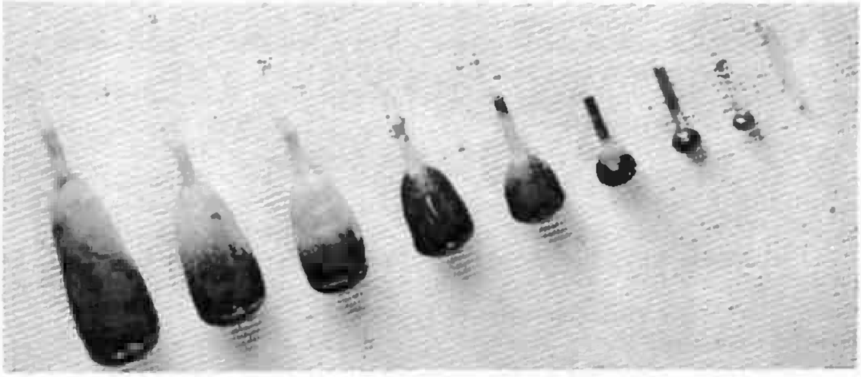


Fig. 15-50. Steps in boule growth in the manufacture of commercial sapphire used for recording and reproducing styli. (Courtesy, Linde Co.)

15.54 Where does sapphire fall on the hardness scale?—Synthetic corundum has a hardness rating on Moh's scale of 9 as compared to 10 for diamond and 5.5 to 7 for hardened steel. Laboratory tests show synthetic white sapphire has a tensile strength of the same order of magnitude as steel. When heat treated between 1700 and 2000C (3090 and 3632F), this material exhibits a plasticity which will permit it to be worked by flame-forming techniques.

15.55 How is commercial sapphire manufactured?—The greater part of synthetic corundum is obtained by feeding finely divided particles of aluminum into a small oxyhydrogen burner. There the material is gradually fused in the flame and a carrot-shaped crystal known as a boule is grown. The finished boule is allowed to cool very slowly in the furnace and then split along a crystallographic plane to relieve stresses built up during the growing process. The average boule weight is 200 carats. Boules in their growing stages are shown in Fig. 15-50.

15.56 What is the life of a recording sapphire stylus?—If treated with care, about six to eight hours. The tip of the stylus should be wiped off with lacquer solvent after each use. This will prevent small microscopic particles from sticking to the tip and fouling the groove while recording.

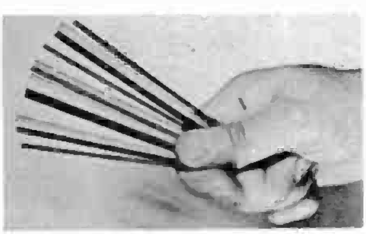
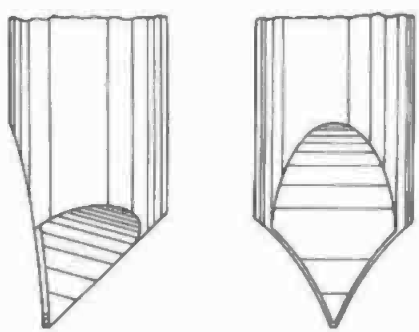


Fig. 15-53. Sapphire and ruby rods used in the manufacture of recording and reproducing styli. (Courtesy, Linde Co.)

15.57 What is the purpose of a scoop-shaped stylus?—It is used to reduce the error of the vertical groove motion es discussed in Question 13.30. The scooped stylus is manufactured by Capps & Co., Inc., and is marketed under the name of Cappscoop. Its construction is shown in Fig. 15-57.

By making the cutting face of the stylus a circular arc, it is possible to reduce considerably the deviation of the cutting head face from the perpendicular to the record surface. In addition, the curved surface of the stylus facilitates the removal of the chip, while improving the signal-to-noise ratio for an average of 3 dB, compared to the conventional flat-faced stylus. Based on considerable research and study, the conclusions reached were that groove walls recorded using a flat-faced stylus



(a) Side view. *(b) Front view.*

Fig. 15-57. Cappscoop recording stylus for stereophonic microgroove recording.

are not actually straight, but are concave, due to the nature of the lacquer disc and the motion of the stylus. This may be verified by a study of the chip, which is not a true triangle.

15.58 What is hot-stylus recording?—A technique of recording using a stylus which is heated to a rather high temperature by a small heating coil wound around the tip of the recording stylus.

15.59 What are the advantages of hot-stylus recording?—The horns caused by cold flow at the upper edges of the record groove are eliminated, permitting a higher level to be recorded. A lower noise level results when cutting at low groove velocities; thus, the signal-to-noise ratio is increased. With hot-stylus techniques, diameter equalization is not always used because of the improved high-frequency response.

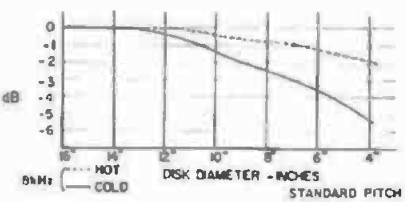


Fig. 15-59. Comparison of the frequency response obtained with a hot and cold stylus.

15.60 To what temperature is a hot stylus heated?—Operating temperatures vary from 350 degrees Fahrenheit upward. In a series of tests conducted by Jackson, under controlled conditions, it was learned that the particular coil in use produced the highest output level at a current of 500 milliamperes. As an example, let the output level for the current of 500 mA at 5 inches equal the reference level. Changing the current through the stylus heater-coil over a range of 50 mA to 650 mA and recording from a diameter of 5 inches to 4 inches revealed the information shown in Fig. 15-60.

These tests were made using a Westrex 3C recording head, Capps & Co. Cappscoop prewired stylus, Scully Instrument Corp. recording lathe, and Audiotape Corp. recording blanks.

15.61 How is modulation noise affected using a hot stylus?—Experience indicates that modulation noise is practically nonexistent.

15.62 What type current is used to heat a hot stylus?—Either alternating or direct current.

15.63 Will alternating current induce hum modulation in the stylus?—No.

15.64 Describe the construction of a typical stylus heating coil.—The average coil will consist of about 7.5 turns of resistance wire, 0.005 inch in diameter, having a resistance of 32 ohms per foot. The coil is closely wound on a drill shank slightly larger than the stylus diameter, to permit the coil to be slipped over the stylus and held in place by the natural spring tension of the coil. As the current through the coil is increased, the signal-to-noise ratio will be increased, up to a point of about 1 ampere. Beyond this point, the chip may be caused to burn and adhere to the stylus, so that its removal will damage the tip. Under no circumstances should the recording be made without the suction pump being on. It is of interest that the use of a hot stylus was experimented with, in 1891, using a flame.

15.65 Can radio-frequency current be used to heat the stylus tip?—Yes, if the stylus is insulated from the cutting head mechanism to prevent heat conduction to the damping materials.

15.66 Is diameter equalization required with hot-stylus recording?—As a rule, diameter equalization is not required with bot-stylus recording since

Current through the coil	*Output level*
650 mA (wire red-hot)	—1.0 dB (5-inch diameter)
500 mA	0.0 dB Zero reference
400 mA	—0.5 dB
300 mA	—1.0 dB
200 mA	—2.0 dB
100 mA	—2.0 dB
50 mA	—0.5 dB (4-inch diameter)

Fig. 15-60. Changing current through stylus heater affects output level.

the high-frequency losses are less than 2 dB at the smaller diameters. However, some recording activities do use a small amount of diameter equalization, even though they are using a hot stylus.

15.67 ***Is less power required to modulate the recording head when using a hot stylus?***—Yes. Slightly less power is required at the higher frequencies; however, at the middle and low frequencies, the power is the same as for a cold stylus.

15.68 ***Does a suction system have a cooling effect on a hot stylus?***—Not to any extent. A normal suction system may be used.

15.69 ***Will a hot stylus permit hard discs to be recorded?***—Yes; discs considered too hard for cold stylus recording may be used if the hot-stylus technique is utilized.

15.70 ***What is the procedure for starting a hot-stylus recording?***—The suction pump is started first. The stylus heater is turned on after the stylus has been lowered onto the disc. If this precaution is not taken, difficulties may be encountered, because the chip becomes quite limp and may foul before the suction system can act.

15.71 ***What is the average increase in the signal-to-noise ratio when using a hot stylus?***—About 18 dB over a cold stylus, resulting in a noise level of approximately 70 dB below the standard NAB signal level. (See Questions 13.85 and 13.86.)

15.72 ***What is a stylus disc?***—A special test record for testing the wear of reproducing styli. The disc is made with cam action grooves which will cause a worn stylus to scrape the side walls changing the color and thus making the wear visible.

15.73 ***Describe the method of mounting styli in a stereophonic recording head.***—One such method is shown in Fig. 15-73A, used in both the Westrex 3C and 3D cutting head. In the 3D head, design of the stylus and the torque tube are such that to replace the stylus, the cutting head does not have to be removed from the recording lathe. The stylus is 1.5 times the diameter of the one used in the 3 C head and has a flattened face 6-mils deep and 39-mils wide, ground the full length of the sapphire. The ground surface serves as the stylus cutting face and as a flat for mating to the precision mount, designed

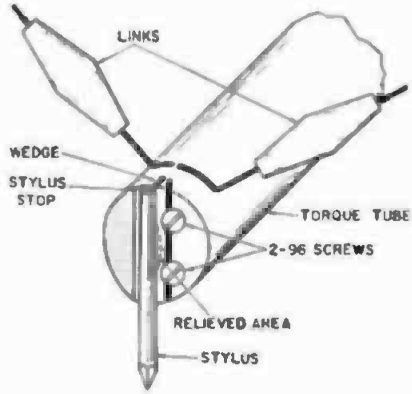


Fig. 15-73A. Torque tube and stylus as used in the Westrex 3C and 3D stereophonic recording head.

to result in a precision alignment of the burnishing facets and cutting face. The sapphire is 0.43 inch in length and inserts in a V way at the end of the torque tube. The V mount is an integral part of the torque tube. Two 2-96 screws located adjacent to a V bind on a wedge and apply pressure to the stylus wall, thereby securing the sapphire in a rigid mount. In Fig. 15-73B, is shown the alignment of the cutting face, which must be within plus-minus 4 degrees of the perpendicular to the central axis of the torque tube. This latter adjustmeut applies to any stereophonic recording stylus. (See Fig. 14-2G.)

15.74 ***Describe the construction of an elliptical reproducer stylus.***—The design of the reproducer stylus has undergone many changes since the invention of the acoustical and electrical phonograph. Problems in the manufacture of styli for monophonic reproduction seem small compared to that now required for the production of styli for stereophonic reproduction. One of the major problems facing the styli design engineer is to obtain a faithful tracking of the sound-track modulations as they

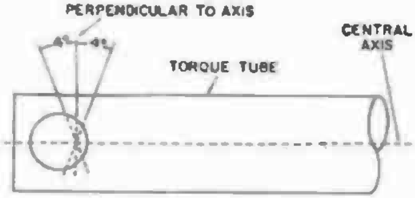


Fig. 15-73B. Torque tube of Westrex stereophonic cutting head showing the alignment of the recording stylus.

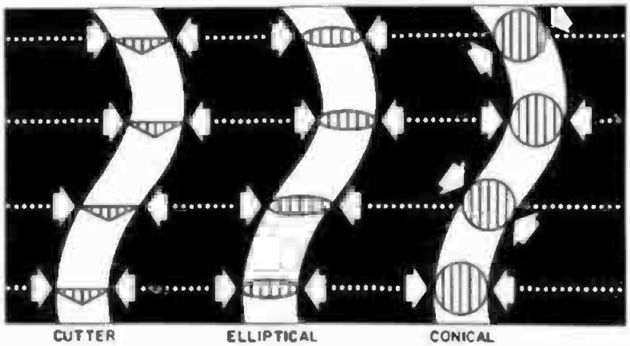


Fig. 15-74A. Tracking angles for an elliptical and a conical reproducing stylus compared to a recording stylus.

were made by the recording stylus. Discussed to some length in Question 13.30 were the problems arising from the shape of the recorded groove in the master, and the difficulty of getting the reproducer stylus to properly track the groove of the pressing, without adding distortion and undue wear. Several types of distortion are possible, namely tracking distortion, tracing distortion, dynamic distortion, and that caused by the loss of contact between the stylus and groove.

Lateral-tracking distortion can be minimized by proper design of the pickup arm. Vertical-tracking distortion is minimized by matching the playback stylus to the vertical-tracking angle of the recorder-stylus angle. Tracking distortion is caused chiefly by the fact that the reproducer stylus is different in shape from that of the recording stylus. Dynamic distortion occurs under reproducing conditions when the stylus deforms the record material and follows a contour that is not quite the same as when the record is static. Distortion caused by the loss of stylus contact and the groove occurs when the reproducer stylus starts jumping off the sound track surface or rattling in the groove.

Referring to Fig. 15-74A, at the left is a drawing, showing the wedge-

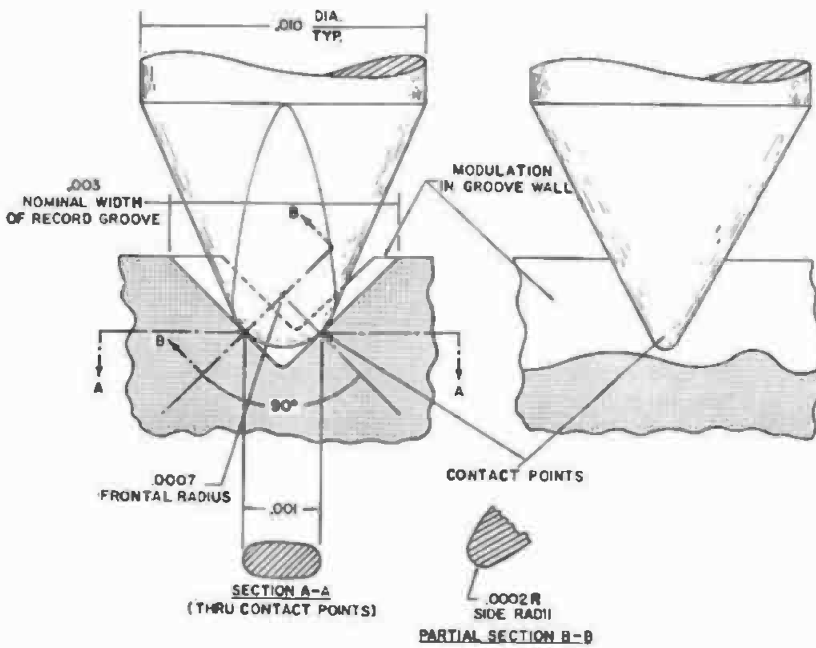


Fig. 15-74B. Bilateral-elliptical pickup stylus dimensions and radii. (Courtesy, Shure Brothers, Inc.)

shaped recording stylus. For perfect reproduction, the reproducing stylus must move in a manner the same as the recording stylus. This cannot be accomplished using a round-end stylus, because the point of tangency between the round end of the stylus and the record changes as a function of modulation. This difference in motion in termed *tracing distortion*, and is directly proportional to frequency and cutting velocity, and inversely proportional to the groove velocity squared. Therefore, tracing distortion is most noticeable as the center of the record is approached for heavily modulated high frequencies. The difference in design between the recording and playback stylus also causes a pinch effect, which is most noticeable when playing back a monophonic record with a stereophonic cartridge.

Fig. 15-74C. Elliptical pickup stylus in a stereo groove.

Referring to the recording stylus groove again, it will be noted that the groove, as cut by the wedge-shaped recording stylus, imparts a vertical motion to the reproducer stylus having a round tip. The perpendicular distance between points of tangency for the round tip is smaller where the sine wave crosses the zero axis at the peaks of the waveform.

Because a stereo cartridge will respond to both vertical and lateral modulation, an electrical output is produced which is primarily second harmonic in nature. Although tracing distortion may be minimized by reducing the stylus point to an extremely small dimension, this can cause noise and distortion because the stylus rides in the bottom of the groove. A minimum acceptable dimension is 0.4 to 0.5 mil. Tracing distortion could also be reduced by making the reproducer stylus an exact copy of the recording stylus. However, because of the cutting edges, this would cause undue wear on the modulations. Therefore, a compromise is necessary between the shape of the recording and reproducing stylus; hence came the *elliptical or biradial* reproducer stylus. The term "biradial-elliptical stylus" is one introduced by Kogen and Samson, to indicate the existence of two different radii, one at the *point of contact* with the sound track modulation, and a second at the *end of the stylus tip*. However, this design is referred to as an elliptical stylus.

The minute dimeusions of a biradial-elliptical reproducer stylus manufactured by Shure Brothers, Inc., appear in Fig. 15-74B. The groove radii portion of the tip are to a scale of 1000:1. A simulated view of a biradial-elliptical stylus in a record groove is pictured in Fig. 15-74C. To achieve the full benefits from a biradial-elliptical stylus, about 1.5 grams vertical pressure should be employed, with the stylus angle set to 15 degrees from the perpendicular.

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Section 16

Pickups

Men like Pierce and Hunt, Pickering, Olson, Stanton, LeBel, Badmaieff, Bauer and many others have all made valuable contributions to the design of the modern day pickup. It has been developed from a device with a vertical force of about one-half pound with an upper response of 3500 Hz, to a unit with vertical pressure as low as ½ gram and a frequency response from 10 Hz to 20,000 Hz and greater. The pickup arm has also received considerable attenuation, with much thought given to tangent error and resonance. In the early days of mechanical reproduction, the arm was designed to serve as a throat which connected the sound box (pickup) to the horn; hence, the name tone arm. Since the arm no longer serves in that capacity, the term "tone arm" is a misnomer, although it is still frequently referred to as such.

This section deals with such aspects as driving force, acceleration, compliance, and tracking, for both monophonic and stereophonic recordings. It supplies information as to types, such as ceramic and crystal, magnetic, variable-reluctance, dynamic, fm, and others. It discusses the design of arms, resonance and mounting, and gives other pertinent data relative to the use of snch devices.

16.1 What is an electric pickup?—An electromechanical device for tracing the vibrations in a sound track of a record in such a manner that the instantaneous vibratory velocity of its moving system is in proportion to the undulations of the record groove. The mechanical vibration of the pickup system causes an electrical current to be generated, which is then amplified and reproduced over a loudspeaker.

The output voltage of an electric pickup, depending on the design, may be proportional to the velocity or the amplitude of the moving system.

16.2 Describe the basic construction of a magnetic pickup.—The basic principle of an early magnetic pickup, used for record reproduction and known as a balanced armature, is shown in Fig. 16-2. Although present-day designs are quite different, the same principles of operation, electrically, still apply. The essential parts are: a permanent magnet A, with its pole pieces B, a soft-iron armature C, pivoted at D and mounted between the pole pieces of the magnet in rubber bearings E. Movement of the armature by the stylus H disturbs the magnetic lines of force (shown by the dotted lines) causing a voltage to be generated in coil F. Two tungsten-loaded rubber damping blocks G are used to center the armature in the center of the magnetic field.

In these early model pickups, the stylus pressure was several ounces (at least 3 to 4), while the modern pickup uses 1 to 2 grams. The stylus was generally a steel needle, similar to those used in an acoustic sound box, as described in Question 16.24. The greatest drawback to this design was the frequent replacement of rubber bearings E, damping blocks G, and the centering of the armature C in the magnetic field. The output voltage was proportional to the stylus velocity, and, as there were no standards for recording and reproduction, little or no equalization was used. If the armature was not accurately centered in the magnetic field, the distortion was quite high and was compensated for by attenuating the high frequencies with a "fuzz filter" (a simple low-pass filter) consisting of a coil and capacitor. This filter also helped somewhat in reducing the surface noise of the shellac pressings used at the time.

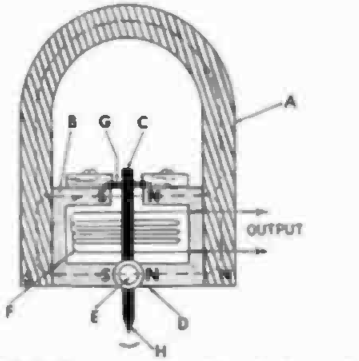


Fig. 16-2 Construction of a balanced-armature type pickup, showing the magnetic lines of force.

16.3 *Explain the basic design of variable-reluctance pickups.*—Since the introduction of the original variable-reluctance pickup, many different versions of its design have appeared. One of the earliest was that by Clark in 1947 and is given in Fig. 16-3A, and in principle, is still used for both monophonic and stereophonic design. The magnetic structure consists of two pole pieces A, with a small permanent magnet B between them. At one end, coil C is mounted with a soft rubber insert D.

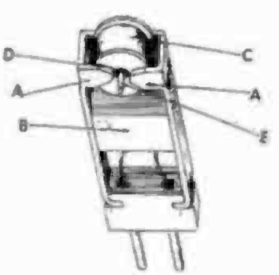


Fig. 16-3A. Variable-reluctance magnetic pickup (after Clark).

The stylus E, which is also the armature, is held in the exact center of the magnetic structure by the rubber insert. When the stylus is actuated, its movement causes a voltage to be generated in the coil. Because of its construction, the frequency response was above the normal audio-frequency band. Output voltage was on the order of 100 millivolts at 1000 Hz, with an output impedance of 500 ohms. The recommended stylus pressure was 15 to 20 grams. The stylus weighed 31 milligrams and was removable. Although the recommended pressure was 15 to 20 grams, the pressure could be as low as 7 milligrams.

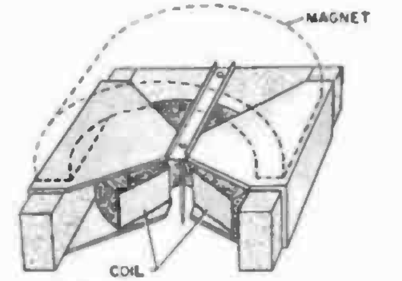


Fig. 16-3B. Internal construction of early model Pickering variable-reluctance pickup.

The frequency response was plus-minus 2 dB, 20 to 20,000 Hz.

A second variable-reluctance pickup to make its appearance was that by Pickering (Fig. 16-3B) and was of similar construction and design. This early model used a stylus pressure of 15 grams and developed an output voltage of 17 millivolts for an impedance of 500 ohms. The magnet was placed as shown by the dotted lines. The frequency range was about the same as for the first structure by Clark. The compliance was 1×10^{-6} cm/dynes. The stylus was not replaceable.

A third design, shown in Fig. 16-3C was that used by General Electric in their variable-reluctance pickup, which made its appearance about 1952. In this design, the magnetic circuit consists of two coils A, yoke B, and pole pieces C. Armature D, which also acts as a stylus bar, is mounted in such a manner that one end is held in the exact center of the pole pieces. The armature rests on a rubber damping block G, which also reduced the vertical compliance. The sty-

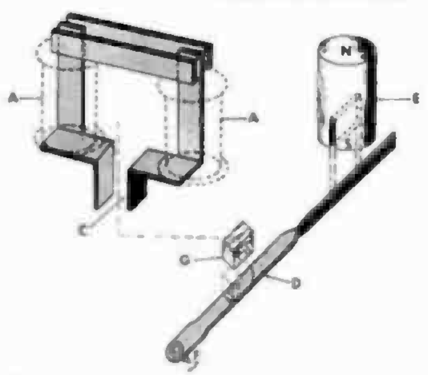


Fig. 16-3C. Basic principle of a variable-reluctance magnetic pickup structure, showing the magnetic circuit, pole pieces, coils, and stylus bar.

lus is mounted on the forward end of the armature at F. The magnetic circuit extends from the south pole of the magnet E, through the armature, up through the pole pieces, and back through the air to the north pole of the magnet to complete the circuit. If the armature is properly centered, the magnetic flux is divided equally between the pole pieces. Lateral motion of the armature between the pole pieces conducts the magnetic flux alternately through the cores of the two coils which are connected to operate push-pull. When the armature is moved in a lateral direction by the stylus action, the magnetic flux is increased or decreased through the coils. Thus, the output voltage varies directly with the change of flux, or in direct proportion to the velocity of the armature. Mechanical design limits the motion to the lateral direction only, thus reducing noise and distortion. The impedance of such a design is rather high and is on the order of 3000 ohms at 1000 Hz. The output voltage is about 10 millivolts. The frequency response was within plus-minus 2 dB, from 30 to 15,000 Hz. Stylus pressure was 6 to 8 grams. It also could be obtained in a dual model, in which the stylus bar carried a 1.0-mil and a 3.0-mil stylus. Either stylus could be selected by rotating the stylus bar.

In about 1956, a decided improvement was made in the design of variable-reluctance pickups, with the appearance of the Pickering Fluxvalve, designed by Stanton. The pickup was manufactured in two models, a single

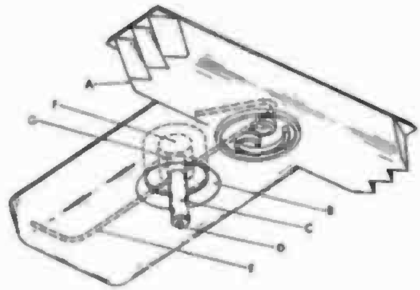


Fig. 16-3E. Removable stylus mounting used in the turnover pickup of Fig. 16-3D.

pickup mechanism, and a dual type in which two replaceable styli inserts were used and could be removed at the front of the pickup body by sliding the inserts forward. One side of the pickup carried a 3-mil stylus for reproducing coarse-pitch recordings, and the other side a 1.0-mil stylus for reproduction of microgroove recordings. A phantom drawing of its construction appears in Fig. 16-3D, with the stylus insert shown in Fig. 16-3E. The insert consists of a plastic mounting A, containing an inverted cup B, surrounding the stylus shank C, containing the stylus tip D. The stylus shank C is held in place by a wire support E, embedded in the plastic insert A. At the upper portion of cup B, a small metal button F surrounds the upper end of the stylus shank, but leaves it quite free to move by the action of the stylus in the record groove. With the stylus insert in place (Fig. 16-3D), the upper end of the stylus G projects parallel over the center pole

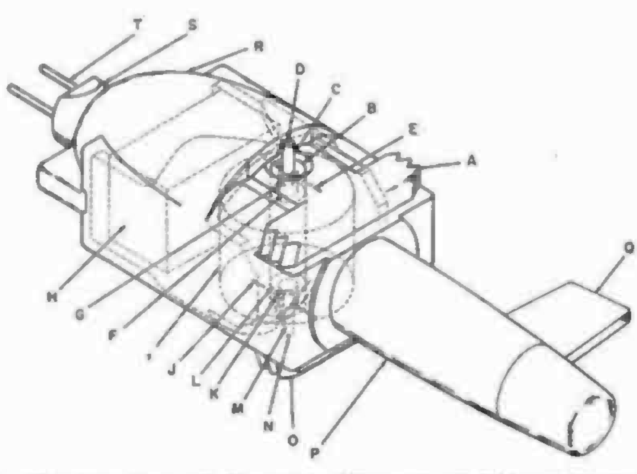


Fig. 16-3D. Phantom view of Pickering Fluxvalve variable-reluctance magnetic pickup.

of coil I in the body. The magnetic flux from the permanent magnet H also in the body flows through the metallic member to the button in the insert and to the stylus shank itself. Directly in back of the upper end of the stylus shank are two metallic members placed side by side and centered on a line passing through the center of the stylus rest position. In this manner, magnetic flux is carried back into the housing and through the coil. With the stylus at rest, the magnetic circuit is so arranged that flux flows in equal amounts to the two metallic members, and being balanced, no voltage is generated in the coil. When the stylus is actuated by the modulation of the record groove, the magnetic flux is disturbed and is varied from side to side as the stylus vibrates, thus generating a voltage. When stylus D is moved, supporting wire E is twisted first in one direction and then the other, but always returns the stylus to the exact center. The vibratory mass is comprised of only the stylus shank and tip; thus, the mass is kept to a very low value. Armature resonance occurs above 30,000 Hz, which removes any peaks in the audible range. The entire mechanism, including the magnet, coil, and magnetic gap, is enclosed in a plastic body. The output voltage is approximately 18 millivolts for a velocity of 7 centimeters per second, with a uniform frequency response between 10 and 20,000 Hz. The principal parts of the lower portion are: stylus support wire J, end of stylus shank M, stylus tip N, turnover handle P, colored flag Q to indicate which stylus is in use, plastic body R, rear support bearing S, and the output terminals T. This pickup may be adjusted for a stylus pressure of 2 to 6 grams, using a stylus of 0.5 mil or 3.0 mils.

16.4 Can a pickup be designed to reproduce both lateral and vertical coarse-pitch recordings?—Yes, pickups have been designed to respond to both lateral and vertical modulations as typified by the early Western Electric Model 9A monophonic moving-coil pickup manufactured some years ago (Fig. 16-4A). The motional structure consisted of two coils A, mounted on an armature B which carried a diamond stylus C. A U-shaped permanent mag-

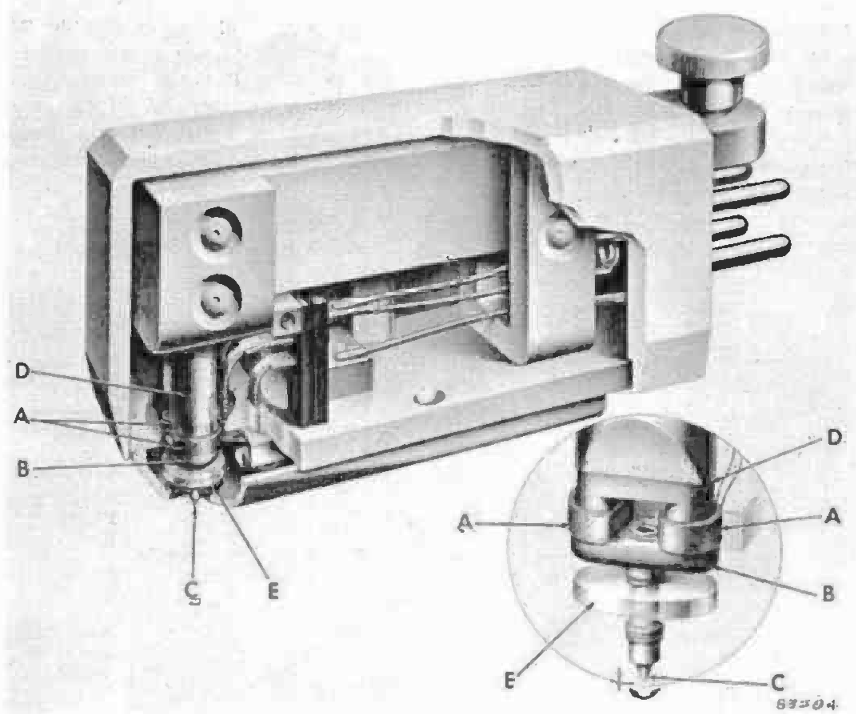


Fig. 16-4. Interior view of an early Western Electric 9A moving-coil vertical-lateral coarse-pitch pickup (now *obsolete*).

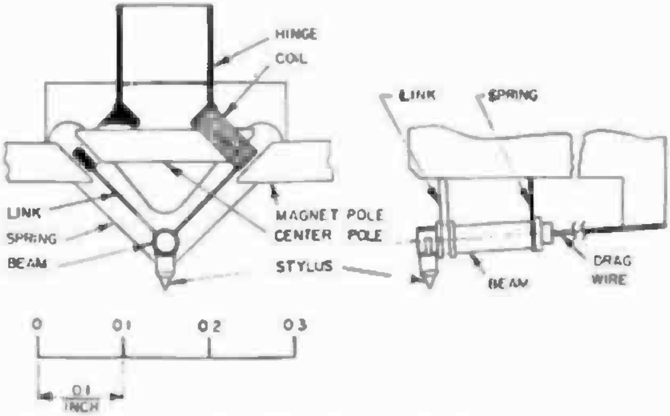


Fig. 16-5A. Simplified cross-sectional drawing showing the interior construction of a moving-coil stereophonic pickup.

net D, with pole pieces extending into the coils, supplied the magnetic field. The coils, armature, and stylus were mounted in a viscoloid disc E, which supports and centers the moving elements. Although this pickup has been *obsolete* for some years, its structure is of interest because of its design that will permit a single mechanism to be used either in the vertical or lateral directions.

The coils move vertically over the magnetic pole pieces for vertically recorded records and transversely for laterally recorded records. In so doing, a voltage is generated which has a waveform similar to the modulations of the record groove. Connections from the two coils are brought to a switch in the equalizer network for changing to either type of reproduction. Thus, discrimination is obtained between the two different type groove modulations.

Vertical noise components which are always present in a lateral recording are suppressed when the reproducer is switched for lateral reproduction. Conversely, when the reproducer is used for vertically cut records, unwanted lateral motion inherent in the sides of a vertical-cut record is suppressed. The output impedance of this head is on the order of 9 ohms and requires a matching transformer at its output to match the input impedance of the preamplifier.

16.5 Describe the basic construction of a moving-coil stereophonic pickup.— One of the first moving-coil pickups was the Western Electric 9A described in Question 16.4, which reproduced both lateral and vertical coarse-pitch records, using a single stylus. Several different types of moving coil pickups have been developed for both stereophonic and monophonic reproduction. Among the first for stereo was the Westrex Model 10A, now obsolete (Fig. 16-5A). The two self-supporting voltage-generating coils are mounted on Mylar hinges with the axes of the coils at right angles to each other and mounted 45 degrees to the horizontal. The lower edge of each coil is connected mechanically through a wire to a beam

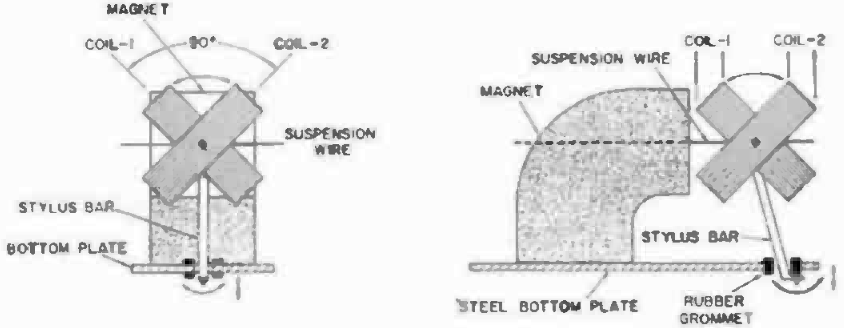


Fig. 16-5B. Constructional view of moving coil pickup for stereophonic reproduction.

which supports the stylus. This beam consists of a small metal tube with an outside diameter of 0.031 inch and approximately 0.15 inch long. The stylus beam or tube is not subjected to twisting or bending. A drag-wire consisting of a flat spring is connected to the rear of the stylus beam and then to the reproducer housing to prevent the beam from rotating. Thus, an equal compliance is obtained at the stylus for any vertical direction. The drag-wire also prevents any longitudinal motion of the stylus.

A semisolid damping material between each link and the reproducer housing (near the coils) provides mechanical damping to the system to remove high-frequency peaks and also helps to reduce crosstalk between the two channels.

The vertical angle of the stylus is set to 15 degrees; however, in early pickups of all manufacture the stylus angle was what the individual manufacturer thought best. The magnetic path of the pickup structure consists of two pole pieces, a center pole, and a permanent magnet. One edge of each voltage-generating coil is placed over the end of the center pole piece, in the gap formed by the center and outer pole pieces.

The coils are phased to produce in-phase output voltages when the stylus is actuated by a laterally recorded groove.

The compliance of the Westrex stereo pickup is about 2.5×10^{-6} cm/dynes. The stylus included angle is 40 to 55 degrees, with a tip radius of 0.5 or 0.7 mil. The output voltage averages about 2 millivolts per coil for a peak velocity of 10 centimeters per second. Separation is on the order of 25 to 30 dB, at 1000 Hz.

A second moving-coil type pickup is shown in Fig. 16-5B. This design employs two moving coils mounted one within the other. The coils are supported on two wires, placed 90 degrees with respect to each other, in the horizontal plane, to permit their movement by the stylus in a rotary manner, both vertically and laterally. As a rule, the coils will consist of only a few turns of wire wound self-supporting, with a stylus bar attached to the center axes of the coils. A permanent magnet is mounted directly behind the coils, as

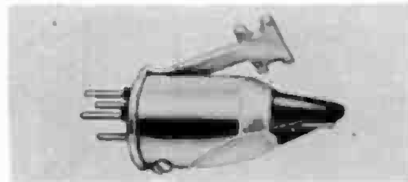


Fig. 16-6A. Stereodyne Model SP-6, 15-degree, push-pull stereophonic pickup by Bang and Olufsen of Denmark.

shown at C. When the coils are actuated by the stylus, they rotate in a circular manner, cutting the magnetic lines of force supplied by the magnet; thus, a voltage is generated in the coils. As the output voltage from the coils is quite low, a step-up transformer is used to increase the voltage output.

16.6 Describe a push-pull stereophonic pickup.—Such a pickup is shown in Fig. 16-6A and is manufactured by Bang and Olufsen of Denmark. The internal magnetic structure is shown in Fig. 16-6B and consists of a small armature in the form of a cross, made of Mumetal, which swings between four pole pins. A stylus bar constructed of aluminum tubing 0.002-inch thick is attached to the Mumetal armature cross at one end. The stylus is secured to the other end of the tube. Four pole pins with four coils are placed at each end of the cross. With a 45-degree motion to the right, a reverse voltage induction takes place. Such action permits the coils to be connected push-pull, thus reducing harmonic distortion induced by the nonlinearity of the magnetic field. In addition, the coils provide an effective hum-bucking circuit. (See Question 8.98.)

Cross talk between the left and right channels is minimized, since such com-

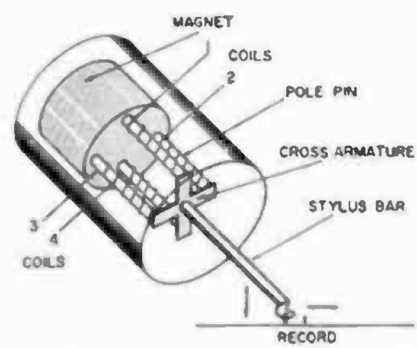


Fig. 16-6B. Simplified drawing of coils and cross armature.

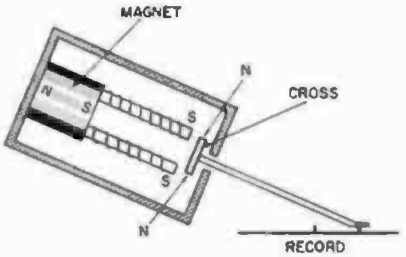


Fig. 16-6C. Magnetic circuit.

ponents are bucked out. The overall cross-talk level is quite low for any frequency, since the voltage induction comes only from changing the spacing of the armatnre cross arms and the pole pieces. Modnlating one channel 45 degrees, the cross arms on the orthogonal channel rotate without changing the spacing, therefore there is no induced voltage in this channel, assuming the positioning of the unit, with respect to the groove, is correct.

A cross-sectional view of the magnetic circuit is shown in Fig. 16-6C and is similar to the magnetic structure of a loudspeaker employing a center magnet. Thus, a closed magnetic circuit is provided which prevents leakage of the magnetic field, and being nonmagnetic, it cannot be attracted to the steel turntable plate, while providing an effective shield for the coils. The stylus bar pivots on a nylon thread, bonded to a plastic support. The armature cross bears on a resilient disc (Fig. 16-6D), which controls compliance and supplies damping for the moving system. The rotational point of the system is at the junction of the armature cross and the nylon thread snpport. The output voltage is 7 millivolts for each channel for a 5 centimeter per second cut. The stylus has an angle of 15 degrees and may be

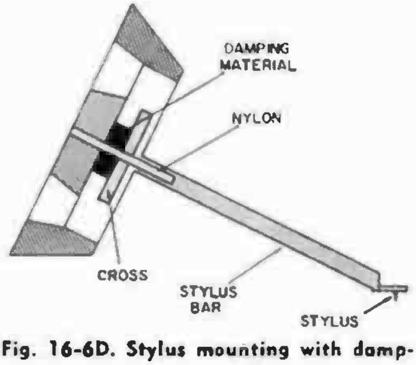


Fig. 16-6D. Stylus mounting with damping control.

operated at a pressure of 1 to 3 grams. Compliance is 15×10^{-6} cm/dynes for both directions of motion. Frequency response is plus-minus 2.5 dB, 20 to 20,000 Hz.

16.7 Describe the construction of a semiconductor stereo pickup.—A semiconductor pickup cartridge developed by J. F. Wood and George Grover of the Euphonics Corp., Guaynabo, Puerto Rico, is shown in Fig. 16-7A and operates somewhat on the principle of the strain gauge. The pickup mechanism employs two small, highly doped silicon semiconductor elements (0.008 inch $\times$ 0.005 inch) whose resistance varies as a function of the stylus deflection. These sensitive elements are mounted on laminated beams of lightweight epoxy with gold-plated surfaces (Fig. 16-7B). A notch in the beam under the assembly acts as a hinge for stress concentration. Referring to Fig. 16-7C, the construction for a stereophonic cartridge is shown. In this structure two beams are used, each driven by an elastic yoke, coupled to the stylus. Aside from the compliance of the yoke and mounting pads, a mechanical advantage of over 40:1 can be attained in the beam and stylus lever. This mechanical transformer provides high compliance and reduces the mass of the elements reflected to the stylus. This stylus, elliptical in shape, is set at an angle of 15 degrees.

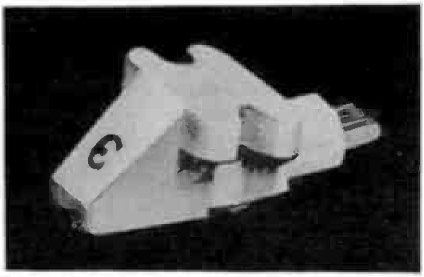


Fig. 16-7A. Miniconic U-15 semiconductor pickup cartridge manufactured by Euphonics Corp., of Puerto Rico.

Since the semiconductor elements are sensitive modulating devices and not generators as in the conventional pickup, very little energy is required for their operation. The compliance at 1000 Hz is approximately 25×10^{-6} cm/dynes, and because of the low mass of the semiconductor elements and driving mechanism, the frequency response is

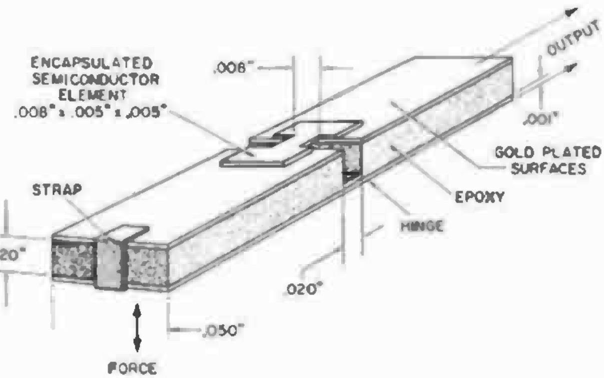


Fig. 16-7B. Beam construction of Euphonics semiconductor stereophonic pickup.

carried out beyond 50,000 Hz. Actually, the device will measure down to dc, but because of the preamplifiers, the low-frequency response is limited to a value below 20 Hz. For its operation, a small power supply (Fig. 16-7D) is required. The supply also contains two single-stage preamplifiers and one inverter stage. A current of 6 mA at 14 volts is supplied to each semiconductor element. As the elements are deflected by the stylus action, the resistance of the semiconductors (about 800 ohms) changes slightly, causing a varying dc voltage across the output. This dc signal is ac coupled to the preamplifiers in the power supply, providing an output voltage of 0.4 volt for each side. The cartridge employs mechanical equalization which, in combination with the RC equalizer at the output of each preamplifier, results in an RIAA reproducing characteristic at the output.

Because of the importance of maintaining symmetry in a stereo cartridge, the beams are oriented with the silicon elements upward in each channel. Such an arrangement gives out-of-phase signals for lateral motion of the stylus. Since the elements have no inherent polarity, reversing the terminals does not change the phase. To properly phase the output signals, the left side is passed through a phase inverter. Using a standard test record with 400 and 4000 Hz, the intermodulation distortion for a stylus pressure of 2 grams at a velocity of 13 centimeters per second, is 2 percent; for a velocity of 15 centimeters per second and a stylus pressure of 1 gram, the distortion rises to about 10 percent. Separation is 25 dB up to 11,000 Hz and better than 15 dB to 20,000 Hz. Square-wave reproduction is quite good, with a slight overshoot on the leading edge at 1000 Hz. The signal-to-noise ratio at the output of the preamplifier is greater than 80 dB below a reference level of 1 milliwatt. Such pickups are not subject to extraneous magnetic or electrostatic fields. The normal precautions taken for grounding

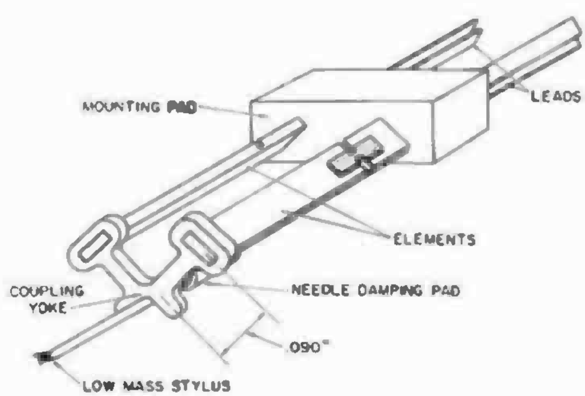


Fig. 16-7C. Cartridge construction of Euphonics semiconductor stereophonic pickup.

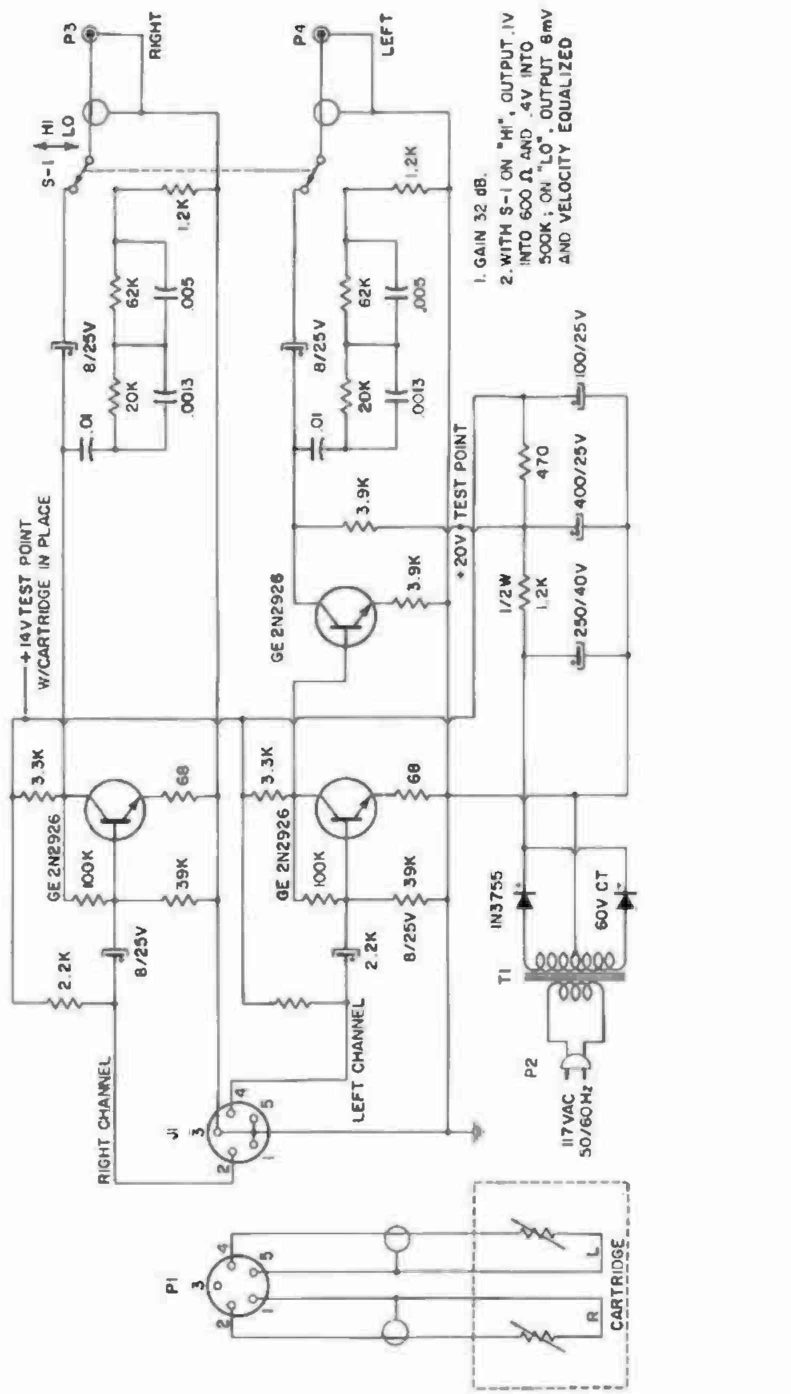


Fig. 16-7D. Schematic diagram for Euphonics Miniconic PS-15 semiconductor pickup power supply.

also apply to this type pickup. (See Question 4.119.)

16.8 Describe the basic principles of a moving-magnet stereophonic pickup. —A moving-magnet type variable-reluctance stereophonic pickup is somewhat similar in design to the moving-coil type described in Question 16.5, except that in the moving-magnet type, the coils are fixed and the magnetic

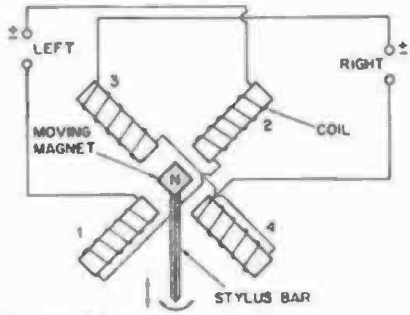


Fig. 16-8A. Basic principle of a moving-magnet type stereophonic variable-reluctance pickup.

lines of force are set in motion, as shown in Fig. 16-8A. A tiny magnet is mounted on the inner end of the stylus bar then suspended in the center of the four coils in such a way that it may be actuated in a rotary manner, similar to the coils of a moving-coil system. Regardless of how the lines of force are cut, either by moving coils or the magnet, a voltage will be generated that is proportional to the stylus velocity. The mass at the inner end of the stylus must be such that little mass is reflected to the stylus. Pickups of this design are noted for their excellent reproducing characteristics. In Fig. 16-8B is pictured a moving-magnet type pickup, Model V15 Type II manufactured by Shure Brothers, Inc.

16.9 Describe the basic structure of a piezoelectric pickup.—Crystal and ceramic pickups are somewhat similar in construction as both devices employ a piezoelectric element for the generation of voltage. The basic construction for a crystal monophonic pickup is given in Fig. 16-9A.

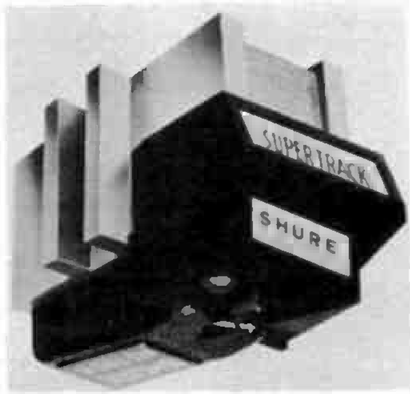


Fig. 16-8B. Shure Brothers Inc., Model V15 Type II variable reluctance pickup, using a moving magnetic structure.

The motional structure consists of two Rochelle salts crystal slabs A and B which are separated by a metal foil C. A foil lead D is attached to each crystal for connection to an external circuit. The opposite ends of the crystals are clamped in a rubber sleeve mounted in a clamp E called a torque jaw. The lead ends of the crystals are clamped between the rubber blocks F. The stylus G is held in a chuck H and clamped by the screw I. The chuck is moved in a lateral direction due to the motion of stylus G in the record groove.

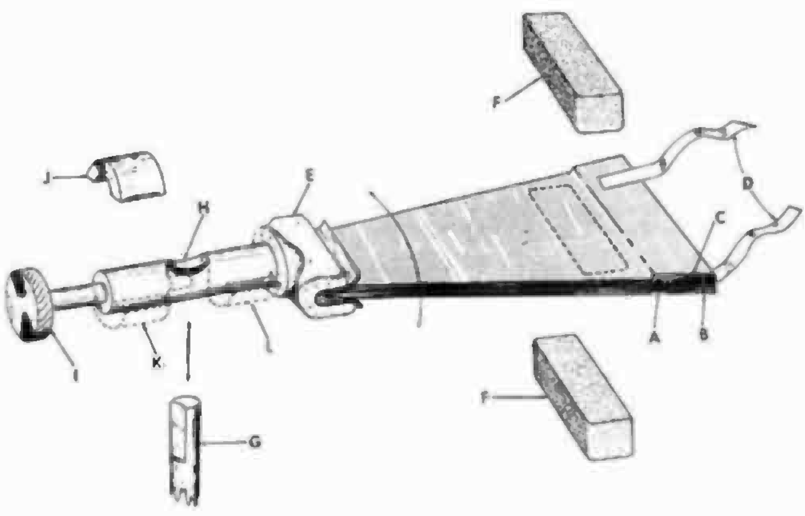


Fig. 16-9A. Interior mechanism of crystal pickup.

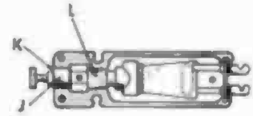


Fig. 16-9B. Bottom view of crystal pickup cartridge.

The twisting motion of the torque jaw causes the crystal slabs to generate a voltage due to the piezoelectric characteristics of the crystals. (See Question 4.13.) The voltage generated by the crystals is proportional to the amplitude of the stylus displacement. The rubber sleeves J, K, and L are used to hold the crystal assembly in its case, as shown in Fig. 16-9B.

The output voltage of the average piezoelectric pickup is considerably higher than for other type pickups. Piezoelectric pickups are treated electrically as a capacitive-reactance device since the impedance rises with a decrease of freqnency, and vice versa. Simple RC networks are used with this type pickup to obtain a frequency response corresponding to the standard RIAA reproducing characteristic. Records recorded using a constant-amplitude characteristic may be reproduced without equalization. (See Questions 14.30 to 14.38.)

Crystal pickups must be carefully mounted on a turntable board, otherwise a disagreeable rumbling noise will result. Before mounting the arm for a crystal pickup, the direction of maximum vibration in the turntable board shonld be determined. The pickup arm is then mounted parallel to the vibratory motion.

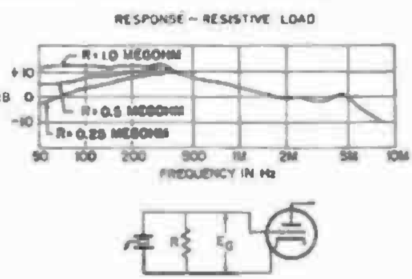


Fig. 16-10A. Low-frequency equalization for a crystal pickup.

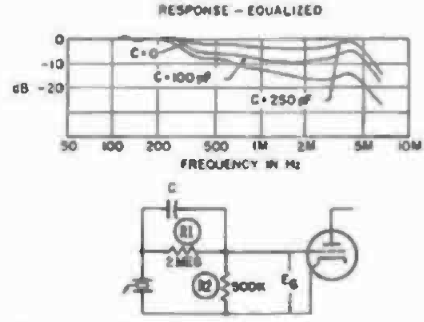


Fig. 16-10B. High-frequency equalization for crystal pickup.

16.10 *Show a simple RC network for use with piezoelectric pickups.*—Six RC equalizer networks for both crystal and ceramic pickups are shown in Figs. 16-10A to D. The networks are connected between the output of the piezoelectric pickup and the input of the preamplifier. The characteristics of these networks are such that they correspond to the standard RIAA reproducing characteristic.

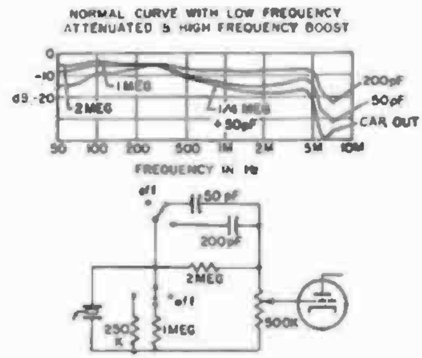


Fig. 16-10C. Complete equalizer circuit for crystal pickup.

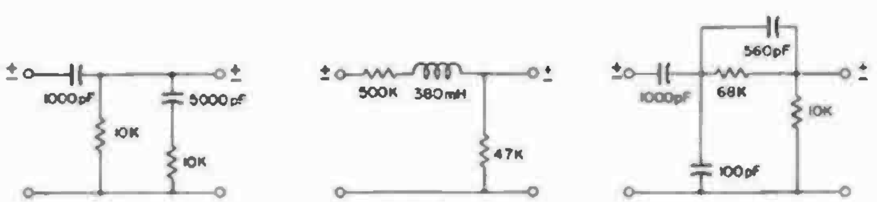


Fig. 16-10D. RC frequency-correction networks for ceramic pickups to reproduce the RIAA characteristic, using a pickup with a compliance of 15 × 10^{-6} or greater. Response will be within plus or minus 2 dB.

ducing characteristics given in Fig. 13-95. (See Question 16.47.)

16.11 What is the internal impedance of the average crystal pickup?—Approximately 100,000 ohms, with a capacitance of 0.001 to 0.0015 μF.

16.12 Is it permissible to couple a piezoelectric pickup by means of a transformer?—Yes, special impedance-matching transformers are available for matching the ceramic and crystal pickup to a lower impedance, such as 600 ohms. However, such transformers are rather hard to design and still maintain good frequency characteristics. It is much more desirable to operate the pickup into a preamplifier containing the equalizer networks as given in Question 16.47.

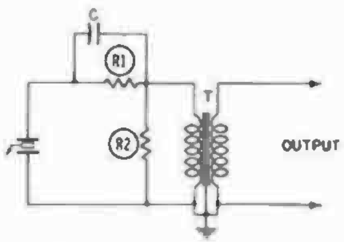


Fig. 16-12. Piezoelectric pickup coupled by means of an impedance-matching transformer to a lower impedance.

16.13 What is a capacitor pickup?—A pickup patterned somewhat after the design of a capacitor microphone. A stylus is attached to a stretched diaphragm mounted on the face of an insulated back plate to which is connected a polarizing voltage. The capacitor head is connected to the input of a resistance-coupled amplifier. A second type capacitor pickup employs the capacity change of the head to modulate a radio-frequency osclllator which is demodulated and then amplified as usual.

16.14 If a lateral pickup has an appreciable amount of vertical compliance, how will it affect the reproduction?—Two output voltages will be generated, one for the lateral and one for the vertical motion. The voltage generated in the vertical direction is added to the lateral voltage in the form of distortion and noise.

16.15 What is a frequency-modulated pickup?—A pickup developed some years ago by Alexis Badmaieff, utilizing an fm circuit in which push-pull action is obtained by varying the resonant frequencies of an oscillator and a discriminator circuit simultaneously, in opposite phase relationship to each other. Modulation of the oscillator and discriminator is achieved through the use of two capacitors shunting the inductance of the two circuits, the common plate of which is grounded and mechanically coupled to the stylus. When the stylus is moved laterally, it, in turn, displaces the common plate of the push-pull capacitor and thus varies the frequency of both the oscillator and the discriminator in opposite phase.

The fm push-pull circuit to which the push-pull capacitor is coupled is shown in Fig. 16-15A. A 6SF7, with its cathode, control grid, and screen, are combined with a tuned inductance L1 and one-half of the push-pull capacitor C1, to form an oscillatory circuit which is electron-coupled to the plate and electrostatically shielded by the sup-

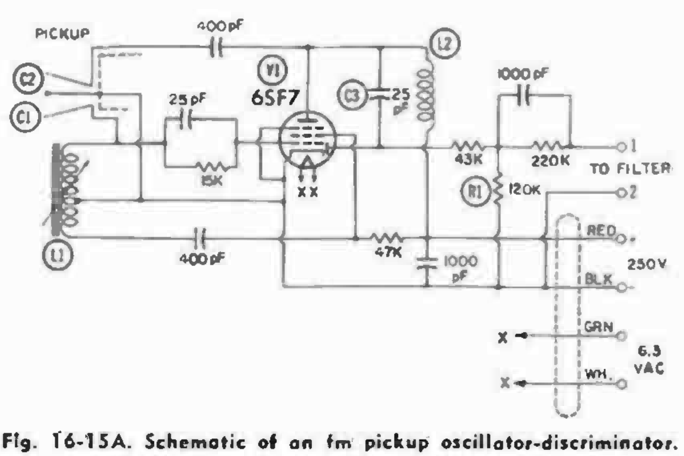


Fig. 16-15A. Schematic of an fm pickup oscillator-discriminator.

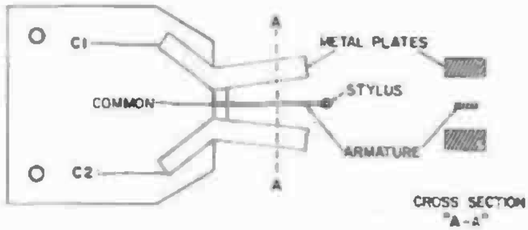


Fig. 16-15B. Simplified form of an fm pickup head consisting of a push-pull capacitor with grounded center plate.

pressor grid. The oscillatory energy from the plate is applied to the discriminator circuit, which consists of another tuned coil L2 and the other half of the push-pull capacitor C2.

The two halves of the push-pull capacitor, C1 and C2, comprise the pickup head of the reproducing unit. The discriminator is slightly off-resonance relative to the mean frequency of the oscillator, the push-pull capacitor being the controlling element of both circuits. The output from the discriminator-tuned circuit is coupled through a capacitor C3 and rectified by the diode section of the 6SF7. After rectification, it is filtered of its rf component and appears as a varying voltage across the resistor R1 which is directly proportional to the stylus displacement.

The push-pull capacitor C1 and C2, which is the head of the pickup, consists of a flat steel wire one-quarter inch long, at the free end of which is mounted a diamond or sapphire tip. The wire is faced on both sides by two metal plates and spaced by a gap of approximately 0.008 inch. The rigidity of the wire is low, being on the order of 1.4×10^{6} dynes per centimeter. The head assembly is shown in a simplified form in Fig. 16-15B. The leads between the push-pull capacitor head and the oscillator-discriminator unit run centrally in the two aluminum tubes comprising the arm for the reproduced unit (Fig. 16-15C). The oscillator-discriminator tube and its circuits are housed in the base of the arm.

The output signal is audio frequency in character, with an output level of minus 45 dBm, after passing through the output filter. The arm resonance is low (about 12 Hz) and is damped mechanically by means of a high viscosity oil in a special ball bearing pivot. The vertical motion of the arm is damped to prevent the head from bouncing. A variable filter or equalizer is shown in Fig. 16-15D which supplies the desired reproducing characteristic. The mechanical resonance of the armature and stylus are above 15,000 Hz, thus enabling the pickup to reproduce frequencies within the audio spectrum without appreciable peaks or dips, over a range of 20 to 15,000 Hz.

The distortion is on the order of 0.5 percent below 5000 Hz, and above this frequency, 1.5 percent. The linearity of the push-pull characteristic is shown in Fig. 16-15E. The oscillator frequency is approximately 40 MHz and is adjusted for a maximum output by using a 1000 Hz constant-frequency record. The output impedance is 250 ohms and may be

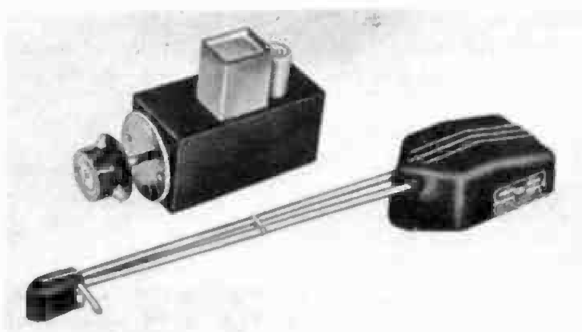


Fig. 16-15C. An fm pickup, arm, and equalizer.

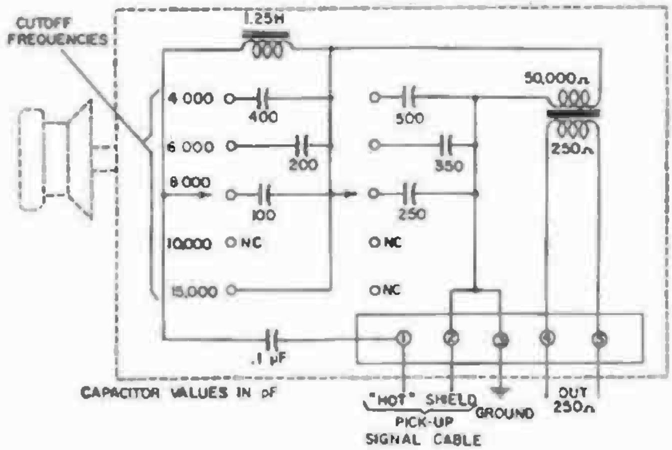


Fig. 16-15D. Schematic diagram of fm pickup equalizer.

used with any normal low-impedance input circuit.

16.16 What is a carbon pickup?—A pickup similar in design to a carbon microphone. A stylus is attached to the carbon button which is connected in series with a source of direct current and an output transformer. Variations of the carbon button resulting from movement of the stylus cause a pulsating current to flow though the primary of the transformer. This current is similar to the modulations in the record groove. The pickup may be designed to use a single or a double button such as are described in Questions 4.8 and 4.9. The disadvantages of this pickup are the same as those of the carbon microphone.

16.17 What is a moving-vane pickup?—A pickup employing a moving member which varies the inductance of an oscillator tank circuit causing a change in its frequency. The changes are detected and amplified in the usual manner.

16.18 What is a photocell pickup?—A pickup employing a photocell which picks up light reflections from the record groove. This pickup was developed many years ago but did not prove very popular because of its poor reproduction. Also, it had to be tracked across the record in such a manner that the photocell was exactly centered above the record groove. The signal-to-noise ratio was low, because of light reflected from the other parts of the record groove which was not part of the modulations.

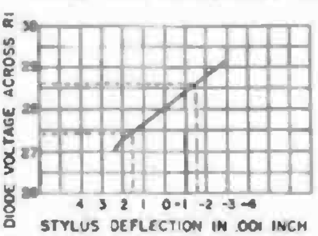


Fig. 16-15E. Linearity of the push-pull pickup head (distortion).

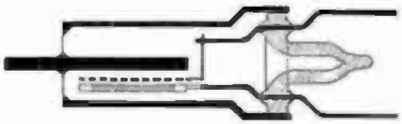


Fig. 16-19A. Cross-sectional view of the RCA 5734 mechano-electronic transducer.

16.19 What is an electronic pickup?—Although all pickups are electronic, the name electronic pickup has been used to designate a unit designed in the form of a small vacuum tube. The device consists of a vacuum tube with a stylus attached to an anode inserted into the interior of the tube through a flexible diaphragm.

The motion of the stylus causes the spacing between the anode and cathode

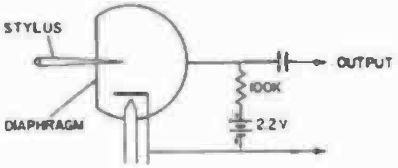


Fig. 16-19B. Circuit diagram for the RCA 5734 mechano-electronic transducer when used as a pickup for reproducing records.

to vary which, in turn, causes the plate current to vary, producing a voltage across the load resistance which varies in proportion to the stylus amplitude in much the same manner as a crystal pickup. The signal voltage may be taken from the plate by means of resistance or transformer coupling. The advantages of this device are its low mechanical impedance and the fact that the output power is taken from the power supply and not the actuating source.

A potential of 22 volts is required at the plate, at a current of 2.5 mA. The effective plate impedance is 5000 ohms, the load resistance is 75,000 ohms, and the amplification factor is 20. A cross-sectional view is shown in Fig. 16-19A and the external circuitry is given in Fig. 16-19B. A typical tube of this type is the RCA 5734 mechano-electronic transducer.

16.20 Describe the construction of a strain-gauge pickup.—The strain gauge is a device used for measuring the strain in structural members of aircraft and similar applications. However, the principle of the strain gauge has been applied to pickups used for record reproduction.

The device consists of a stretched wire through which a continuous direct current flows. A stylus is mounted in the exact center of the wire and makes contact with the record groove. The motion of the stylus causes the wire to be elongated in first one direction and then in the other, causing the electrical resistance of the wire to be changed in accordance with the modulations in the record groove. These minute changes of resistance cause a varying voltage to be generated across the extremes of the wire. These voltages are then stepped up by means of an impedance-matching transformer and amplified in the usual manner. The output voltage is proportional to the stylus amplitude. A simplified view of such a device is shown in Fig. 16-20.

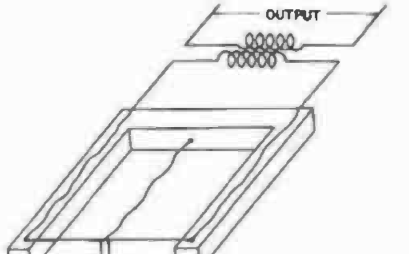


Fig. 16-20. Simplified view of a strain-gauge pickup.

Strain-gauge pickups have been developed with an internal resistance of several thousand ohms and eliminate the need for a matching transformer.

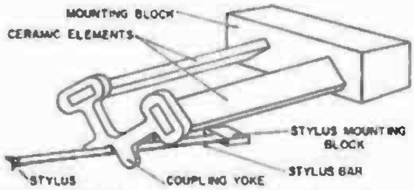


Fig. 16-21A. Simplified drawing showing the construction of a ceramic stereophonic pickup.

16.21 Explain the basic principles of construction for a stereophonic ceramic pickup.—A simplified diagram for a ceramic stereophonic pickup unit is shown in Fig. 16-21A. The moving system consists of two piezoelectric crystal slabs of lead-zirconium titanate, or similar material. This particular material offers good mechanical and electrical properties, with high sensitivity and high capacitance. The ends of the slabs are mounted rigidly in a mounting block, and the front end is connected by a yoke made of injected molded plastic. This coupling is most critical because upon it depends the electrical performance and the mechanical impedance seen at the stylus point by the record groove. The coupling system is defined as that portion of the mechanism that lies between the stylus tip and the ceramic slabs.

The stylus bar is made from heat-treated, thin-walled aluminum alloy tubing, with one end flattened to hold the stylus at the desired angle. The

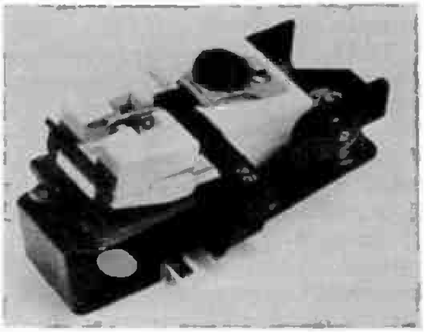


Fig. 16-21B. Ceramic pickup cartridge Model 150-DF manufactured by Electro-Voice Inc.

other end of the stylus bar is held in place by the stylus mounting block. It will be observed that the coupling yoke is connected at a point about midway on the stylus bar. This point is chosen because it affords the most desirable electrical performance and substantially reduces the mechanical impedance of the yoke and ceramic elements as seen by the stylus tip.

Four output terminals are used, two for each channel, to ensure the complete isolation of one side from the other. Damping in the form of a viscous material is used to control the frequency characteristics. The compliance for the cartridge of this type is on the order of 3×10^{-6} cm/dynes, for an effective mass of 14 grams. Since such pickups are of the constant-amplitude type, to reproduce the RIAA reproducing characteristic an RC network is required. The output voltage for such a device is approximately 10 millivolts for a peak velocity of 5 centimeters per second. Ceramic pickups are not affected by magnetic or electrostatic fields. (See Question 16.50.)

A stereophonic turnover-type ceramic cartridge, manufactured by Electro-Voice Inc., is shown in Fig. 16-21B. One side of the stylus bar carries a 0.7-mil stylus for stereophonic reproduction, and a 3.0-mil stylus on the other, for monophonic coarse-pitch reproduction. The compliance in both vertical and lateral directions is 4.5×10^{-6} cm/dynes. Stylus pressure is 4 to 6 grams; however, higher compliance and lower stylus pressure models are available. The desired stylus is selected by a swing of the flanged lever from one side to the other. The output voltage is 0.4 volt, with a frequency response of plus-minus 2 dB, 20 to 20,000 Hz, using a 1-megohm load resistor.

16.22 What is a magnetostriction pickup?—A pickup using a ferromagnetic metal, such as nickel, placed in a strong magnetic field which causes it to expand or contract when in motion. When subjected to compression, the magnetic reluctance changes, varying the magnetic field in which a coil is placed for the purpose of picking up the generated voltage. As a rule, such pickups are designed with magnetic structures resulting in a push-pull output voltage.

16.23 What is a pickup cartridge?—For several years pickup heads have been designed to be easily removed from the pickup arm—therefore the name cartridge. A majority of modern designed pickups use similar mounting holes and terminals to afford a means of easy replacement by the user.

16.24 What is a tone arm?—A hollow arm which supports an acoustical sound box used for reproducing phonograph records acoustically. The sound box is equivalent to the electric pickup, except that it reproduces the modulations of the record groove by acoustical pressure.

The sound box consists of a metal or mica diaphragm with the stylus attached to the center of the diaphragm. The sound box is mounted on a tone arm. The opposite end of the tone arm is connected to a horn. The motion of the stylus in the record groove causes the diaphragm to vibrate. This motion disturbs the air column in the horn, causing sound waves to be generated. A cross-sectional view of a typical acoustical reproducing machine is shown in Fig. 16-24. Although the electrical pickup arm does not carry acoustice sound, the term "tone arm" is still used by some manufacturers and writers.

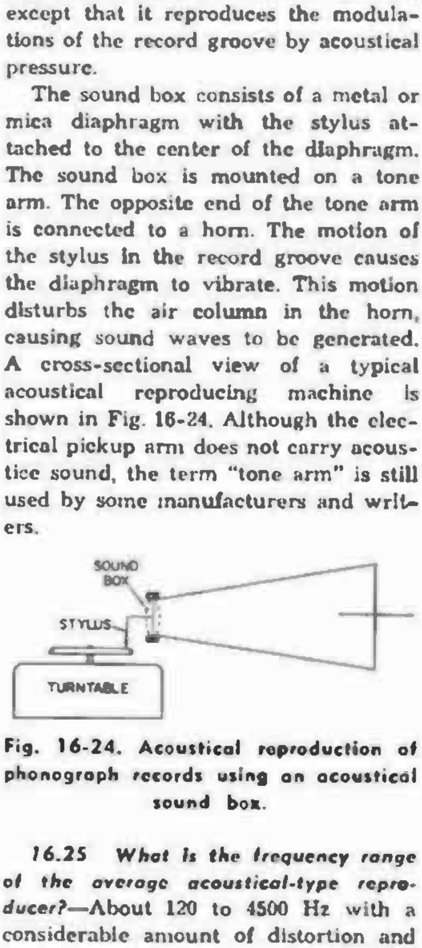


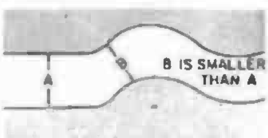
Fig. 16-24. Acoustical reproduction of phonograph records using an acoustical sound box.

16.25 What is the frequency range of the average acoustical-type reproducer?—About 120 to 4500 Hz with a considerable amount of distortion and a number of resonant peaks occurring in the reproduction. Acoustical pickups or soundboxes are used in nonelectrical record reproducers.

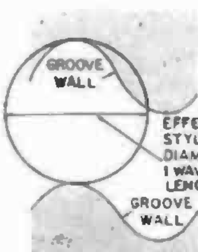
16.26 What is tracking error?—Another name for tracing distortion. This subject is discussed in Questions 13.30 and 13.173.

16.27 What is pinch effect and its causes?—Pinch effect is the result of the record groove pinching the tip of the reproducer stylus and causing it to lift in the groove and sometimes leave the groove entirely. Pinching when present occurs twice per cycle causing the stylus to rise in the groove due to the narrowing of the groove because of sharp changes of directiou in the groove. This is shown at part (a) in Fig. 16-27A. These abrupt changes are caused by the change in direction of the recording stylus as it moves from side to side while engraving the sound track. It is an established fact that when the radius of the reproducer stylus tip is slightly larger than the groove the best tracking conditions exist. This causes the curve of the stylus to ride on the straight sides of the groove wall; however, this must not be carried too far or the stylus will ride on the top edges of the groove. If an attempt is made to trace a sine-wave groove with a point whose effective diameter equals the wavelength of the groove modulation, difficulties will be encountered.

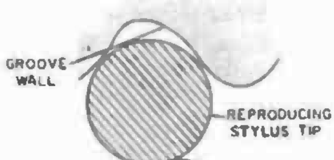
First the pinch effect will cause the stylus to lift out of the groove, as may be seen at part (a) in Fig. 16-27A due to the narrowing of the groove. If the stylus diameter is equal to the modulation wavelength the stylus cannot fit the groove, as shown at part (b) in Fig. 16-27A. What happens when a stylus with a large radius fails to follow the groove when a small radius of curvature is encountered is shown at (c) in Fig. 16-27A.



(a) Sound track narrows because of position of recording sylus.



(b) Stylus tip diameter same as the wavelength of modulation.



(c) Stylus too large a diameter, fails to follow small groove radius of curvature.

Fig. 16-27A. Tracking problems.

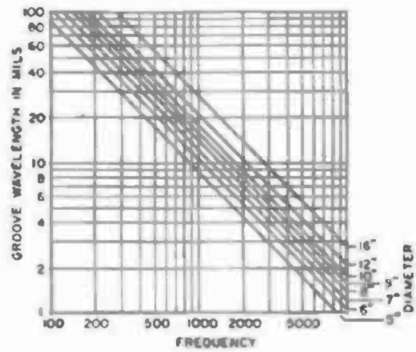


Fig. 16-27B. Groove wavelength versus frequency at a speed of 33⅓ rpm.

The relationship of frequency to recorded wavelength in mils at a speed of 33⅓ rpm is graphically illustrated in Fig. 16-27B. It will be noted for a frequency of 7000 Hz, the wavelength is 3 mils and at a diameter of 5 inches, the wavelength is 1.25 inches long.

Such problems make it quite difficult to reproduce higher frequencies using RIAA recording characteristic if the foregoing problems are not considered.

Pierce and Hunt have shown that the pinch effect produces second harmonic distortion in the vertical direction which will cancel in a lateral recording system if the stylus could lift freely without generating an output voltage. To reproduce the modern microgroove record properly the stylus tip must not exceed 1 mil in radius. Present standards call for 0.5- to 0.7-mil tip radius.

16.28 What are the effects of pickup arm resonance?—The low frequencies will resonate at some frequency, causing a boomy reproduction. This effect is the result of a combination of mass and compliance in the mechanical system and is comparable to an electrical system containing capacitance and inductance where the circuit is resonant at some frequency. A resonant arm will produce a rise at the lower frequencies causing intermodulation distortion components to be generated by combining with higher frequencies.

In a well-designed pickup arm, the resonant frequency is below audibility and is of low amplitude. In some instances motor rumble will cause the arm to be excited and produce an objectionable low-frequency distortion.

16.29 *Describe the effect of pickup arm mass.*—If the mass is too great, the force required to move it exerts excessive pressure on the side walls of the groove, damages the modulations, and decreases the compliance. It also is possible that the arm may be excited by vibrations from the turntable motor, causing resonant peaks in the reproduction.

16.30 *Define compliance in a pickup.*—Compliance in a pickup is the ratio of the stylus to the applied force, expressed in centimeters per dyne. This subject was discussed in Question 15.34.

16.31 *How is pickup impedance rated?*—In the early design of pickups, the impedance was rated in ohms, at either 400 or 1000 Hz. It is the practice with present-day pickups to state only the inductance, the output in millivolts, and the dc resistance. A few manufacturers still rate their product in impedance, and some give only the load resistance. However, knowing the inductance and dc resistance of the windings, the impedance may be easily calculated, using the standard formula for inductive reactance. Because the majority of pickups are designed to be operated into a resistive load (22,000 to 47,000 ohms) and are not required to match a given load impedance, it is not necessary to know the impedance, but only the output voltage. It was also a policy in early designs to design moving-coil pickups for a low impedance, or around 150 ohms. Balanced-armature types (Fig. 16-2) ran from a few hundred to several thousand ohms. Impedance-matching transformers were used when a definite impedance match was required. Crystal and ceramic pickups are considered to be a generator in series with a capacitor and are operated into a high resistance (see Question 16.10) to avoid loss of high frequencies.

16.32 *Define load impedance.*—Load impedance is the impedance the pickup sees at its output terminals when operating. Generally, the load impedance consists of a resistance, and its value for a given frequency characteristic is specified by the manufacturer of the pickup.

Moving-coil pickups (dynamic), generally are of low impedance and operate into a step-up transformer with the secondary terminated or unterminated. Crystal and ceramic pickups are considered to be generators in series with a capacitor workiug into a high load resistance to avoid the loss of high frequencies.

16.33 *How is the output voltage of a pickup specified?*—It is stated for its open-circuit voltage in terms of a standard reference level of 1000 Hz, recorded at a peak velocity of 5 centimeters per second. In some instances, the manufacturer may use a reference level other than 5 centimeters. The output voltage is stated in millivolts.

16.34 *What is playback loss?*—A loss arising from the effect of tracking force pressing the playback stylus into the record groove. The tendency is for the playback stylus to be pressed less into concave portions of the groove (left wall). The greater the pressure on the convex side of the wall (right side), the less stylus displacement; consequently, less voltage output.

16.35 *What is the purpose of an offset pickup arm?*—To reduce the tangent error when reproducing. (See Question 16.52.)

16.36 *What is the average recommended stylus pressure?*—Stylus pressure will vary with the design of the pickup, and the manufacturer. This information is given the user by the manufacturer; however, for modern pickups the stylus pressure will fall between 0.75 and 6 grams. This is also known as the vertical-tracking force.

16.37 *What is viscoloid?*—A material used for damping the action of the mechanical elements in a cutting head, pickup, or other devices where resonant peaks must be reduced.

16.38 *What is a twister element?*—A term applied to crystal elements cut on an axis which results in a piezoelectric effect only as a result of twisting the crystal slab. Such elements are employed in crystal pickups. (See Question 16.9.)

16.39 *What is the maximum temperature at which a crystal pickup may be operated?*—About 133 degrees Fahrenheit. Higher temperatures may cause the crystal to melt.

16.40 What is the cause of distortion in a pickup?—Intermodulation distortion is generated when two or more frequencies are applied to the pickup simultaneously, and is caused by nonlinear tracking of the pickup, and also the mechanics of tracking the groove. The vertical-tracking angle can have considerable effect on the distortion characteristics, as may be seen on the graph in Fig. 16-40. Tracking-angle distortion is caused by the angle of the pickup stylus not being the same as the cutting head stylus. When making pickup-distortion measurements, if possible, a test record having a 15-degree groove angle should be used. Special test records are available for such measurements. The effects of vertical-tracking angle are measured by using a turntable that may be tilted and rotated 90 degrees, with respect to the pickup arm. (See Question 13.30.)

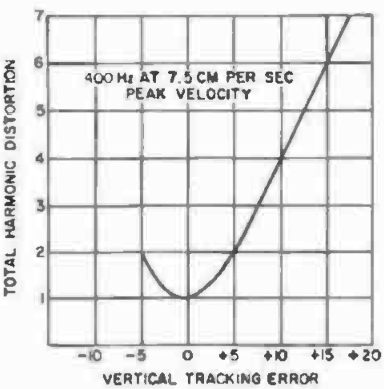


Fig. 16-40. Total harmonic distortion versus vertical tracking error in a stereophonic pickup.

16.41 How is the distortion of a pickup measured?—Pickup distortion may be measured by either the harmonic or intermodulation method. In the first method, a recording is made containing a single frequency of low distortion. This recording is then played back using the pickup to be measured and a distortion factor meter.

In the second method, intermodulation distortion is measured by recording two frequencies of 400 and 7000 Hz simultaneously in a ratio of 4:1. The higher frequency is 12 dB lower in amplitude than the lower frequency. An intermodulation analyzer is connected at the output of the playback circuit when the measurement is made.

In both these methods, the distortion of both the recording and playback systems is included in the measuremeut. Special records containing a single frequency or intermodulation frequencies may be obtained for testing reproducing equipment. Methods of making such distortion measurements are discussed in Section 23.

16.42 Define the term "trackability" as related to a pickup.—Trackability is a term introduced by Kogen and Samson to rate the ability of a pickup to track heavily modulated grooves in a disc record. Considerable research has gone into the study of heavily modulated records, now so prevalent in the industry, and into the development of a pickup with high compliance using a biradial-elliptical stylus. (See Question 15.74.) As this subject is beyond the scope of this work, the reader is referred to the references.

16.43 Why is the intermodulation method of measuring pickup distortion preferred?—Because it is three to four times more sensitive than the single frequency method and is similar to the manner in which the human ear hears. Also, intermodulation measurements are not generally affected by turntable flutter.

16.44 What is a vibration pickup?—A special pickup, generally a crystal, used for making vibration measurements of moving machinery.

16.45 What effect does load resistance have on the frequency response of a pickup?—Considerable effect is noted on the frequency characteristics when measured using no load, and then loading the cartridge. If the cartridge is not operated into its proper load resistance, the frequency response may be seriously affected. The proper load resistance for a given frequency response may be obtained from the manufacturer.

The high-frequency response of the average variable-reluctance pickup may be increased by increasing the value of the load resistance. The load resistance for present-day monophonic pickups range from 22,000 to 27,000 ohms, and 47,000 ohms for the majority of stereophonic cartridges. However, in some instances, a monophonic cartridge may use a higher value of load resistance than 27,000 ohms.

16.46 How is the compliance of a pickup measured?—Compliance in a pickup is a measure of the ease with which the stylus may be deflected from a position of rest, and follow the modulation in a groove. Compliance is defined as the amount of movement resulting when a given force is applied to the stylus tip and is expressed in millionths of a centimeter deflection for a force of 1 dyne or 0.00102 gram.

Compliance is measured by placing the pickup in a fixture and exciting the stylus with an audio frequency, and determining the resonant frequency of the pickup by measuring the voltage at the output of the pickup with a vacuum-tube voltmeter. Knowing the resonant frequency and the mass of the stylus tip, it may be calenlated:

$$\text{Comp} = \frac{1}{4\pi^2 M F^2}$$

where,

M equals the total mass referred to the stylus tip,

F is the resonant frequency in Hz.

At the present time, no standard has been set for the measurement of compliance.

16.47 Give schematic diagrams for magnetic, crystal, and ceramic pickup preamplifiers.—Four preamplifier circuits are shown in Figs. 16-47A to G. Fig. 16-47A shows a low-noise preamplifier, using two common-emitter stages, with a feedback loop containing the equalizer components connected between the collector of the output stage and the emitter of the first stage. The feedback at 10,000 Hz is 30 dB, which results in an input resistance of about 350,000 ohms. Local feedback is supplied by the unbypassed 470-ohm resistor in the emitter of the first transistor. The output level is 1 volt, with a signal-to-noise ratio of 76 dB below 1 volt. The frequency response is that of the RIAA reproducing curve, within plus-minus 1 dB. The gain is 36 dB, with a total harmonic distortion of 0.5 percent for a plus 11 dBm outpnt.

The circuit of Fig. 16-47B is a vacuum-tube preamplifier, using a low-noise 7025 tube. Resistor R1 has a value as recommended for the particular pickup used. The output circuit must not be loaded with less than 220,000 ohms. Figs. 16-47C and D show preamplifiers for use with ceramic pickups.

Referring to Fig. 16-47C, the output is RIAA equalized when used with a ceramic pickup having a capacitance of 5000 to 10,000 pF. The input impedance is approximately 620,000 ohms at 50 Hz. The output voltage is about 1 volt and has a signal-to-noise ratio of 70 dB. The total harmonic distortion at 1000 Hz is 0.6 percent at 1-volt output. The RIAA turnover frequency of 500 to 2122 Hz is obtained by the use R1, R2, and C1. (See Fig. 13-95.) With cartridges of 1000 and 10,000-pF capacitance and using the circuit of Fig. 16-47D, the output is also RIAA equalized. In this circuit, the feedback is taken from the collector base with the equalizer components in the loop, which lowers the

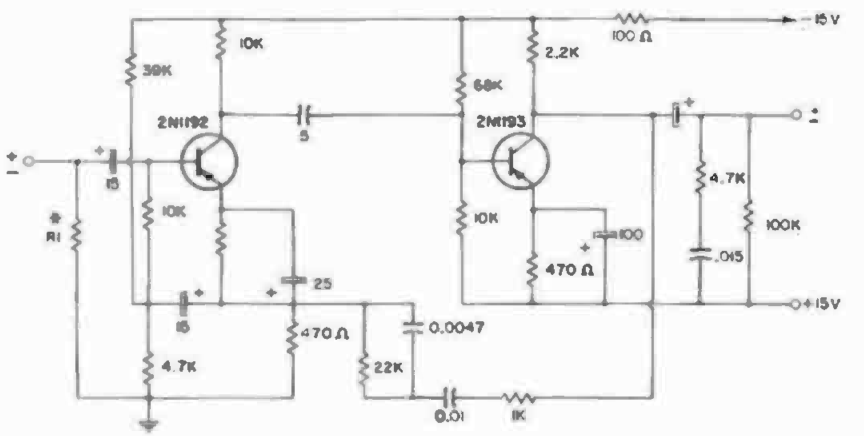


Fig. 16-47A. Transistor preamplifier for magnetic pickup. RIAA equalization. Resistor R1 to be of the value recommended by the manufacturer of the pickup (after Rheinfelder).

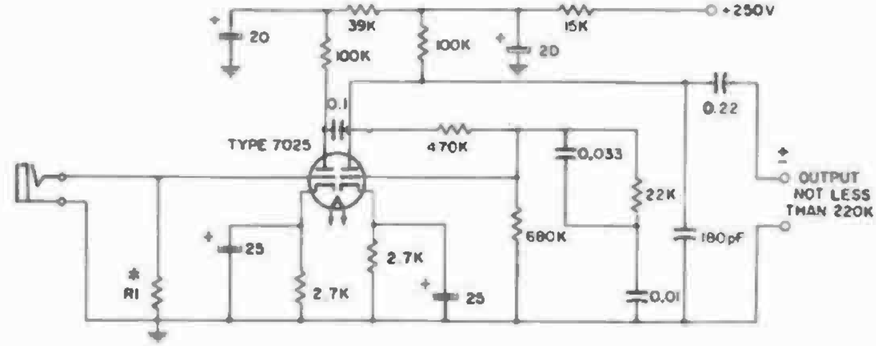


Fig. 16-47B. Vacuum-tube preamplifier for magnetic pickup. Resistor R1 to be a value recommended by pickup manufacturer.

input impedance, thus permitting it to accept a large range of pickup capacitance. The input impedance at 40 Hz is about 30,000 ohms, which decreases with an increase of frequeucy thus causing a velocity response from the cartridge. The frequency response is within plus-minus 1.5 dB from 40 to 12,000 Hz, with a signal-to-noise ratio of 70 dB. Harmonic distortion is less than 0.1 percent at 1.25-volts output. In vacuum-tube type preamplifiers, if practical, dc should be supplied to the heater circuit to reduce the possibility of hum and noise pickup.

A schematic diagram for the Shure Brothers Inc., Model SE-1 stereophonic preamplifier is given in Fig. 6-47E. This preamplifier is designed for studio and broadcast use. Controls are provided for cutting in a 45-Hz high-pass rumble filter and a 7000 Hz low-pass filter, for removing the RIAA equalization, and for reproducing a flat-frequency response. The hum and noise is 64 dB below an output level of plus 4 dBm, over a frequency range of 50 to 10,000 Hz. The separation between channels is 37 dB at 10,000 Hz, using RIAA equalization. Two controls are provided for balancing the two sides. This preamplifier is used in the Westrex transfer channel, discussed in Question 17.228.

Exceptionally good results may be obtained by using field-effect transistors (FET's) in pickup preamplifiers, since results have indicated that a greater signal-to-noise ratio can be obtained by their use. The particular circuit shown in Fig. 16-47F is claimed by its designer to have a signal-to-noise ratio 7 dB lower than the best vacuum-tube amplifier of similar design, with a 5 dB higher overload factor. FET's were discussed in Section 11 and also in Section 12.

16.48 What is an N-A beam test record?—A special test record for mea-

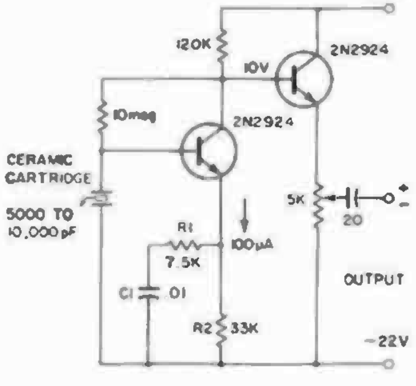


Fig. 16-47C. Transistor preamplifier designed for use with ceramic pickups of 5000 to 10,000 pF capacitance.

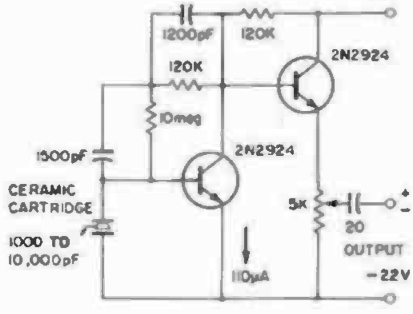


Fig. 16-47D. Transistor preamplifier designed for use with ceramic pickups of 1000 to 10,000 pF capacitance.

suring the intermodulation distortion of a phonograph reproducing system.

16.49 What Is a sweep record?—A special test record on which are recorded frequencies from 50 to 12,000 Hz swept 20 times per second from the lowest to the highest frequency. A light pattern of a typical sweep frequency record is shown in Fig. 16-49A. This record is played back on a reproducing

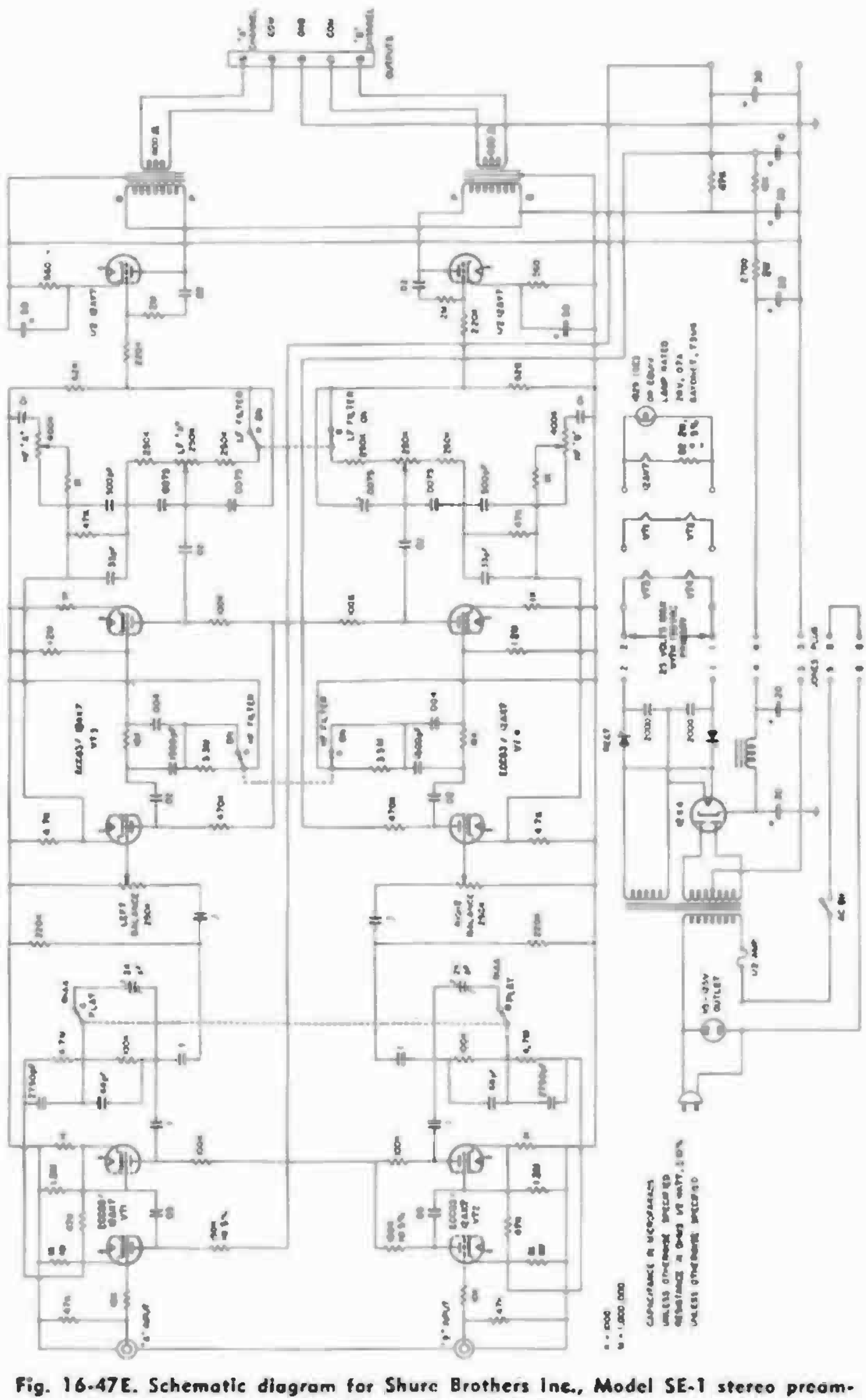


Fig. 16-47E. Schematic diagram for Shure Brothers Inc., Model SE-1 stereo preamplifiers and equalizers.

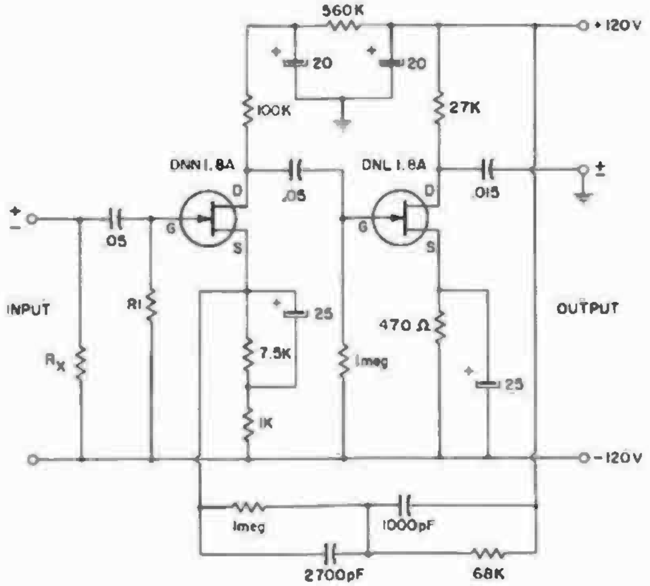


Fig. 16-47F. Pickup preamplifier using field-effect transistors. Resistor R_x and R1 are to be of a value specified by the pickup manufacturer (after Rheinfelder).

system, with an oscilloscope connected across the output. A pattern similar to that shown in Fig. 16-49B is obtained if the system has uniform frequency response. If it does not, the pattern will be distorted. Frequency markers appear at intervals to indicate the frequencies as they appear on the spectrum. This subject is covered in more detail in Question 23.141.

16.50 How does cable capacitance affect the frequency response of a pickup?—Both the cable capacitance and load resistance at the output of a pickup have an effect on the frequency response; the higher the internal impedance of the pickup, the greater will be the loss at the high frequencies. In selecting cable for connection to a pickup, the capacitance per foot should be as low as possible and the cable well shielded. Cable capacitance has the ef-

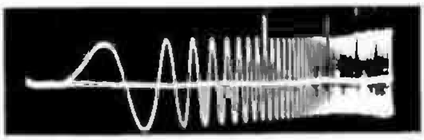


Fig. 16-49B. Pattern obtained for a reproducing system of uniform frequency response, using a sweep-frequency record as shown in Fig. 16-49A.

Fig. 16-49A. A typical light pattern of a sweep record. The pattern shown covers a frequency range of 50 to 12,000 Hz repeated 20 times per second using the standard RIAA reproducing characteristic.

fect of a low-pass filter. Typical values for 4-wire cable, measured between conductors, and the shield and conductors are:

1 wire and shield, 33.4 pF per foot.

2 wires, connected in parallel at one end only, and shield, 52 pF per foot.

2 wires, no shield, 21.1 pF per foot.

16.51 What is the relationship of frequency to wavelength for a given recording diameter?—This relationship is shown graphically in Fig. 16-27B.

16.52 How may pickup-arm tangent error be minimized?—Tangent error may be minimized by use of either an offset or curved arm which progressively reduces the tangent error as the arm moves across the face of the record.

The conventional pickup arm is generally mounted in a swivel bearing at the rear and to one side of the turntable. With such a mounting, the pickup can be tangent to the groove at only one point in the recorded area, the center. In the early design of pickup arms, the arm was straight and it was not uncommon to find tangent errors of 16 to 30 degrees at the inner and outer diameters of a 12-inch record. The use of this type mounting tears out the side wall of the record groove. Tangent errors may be reduced by increasing the length of the arm, but if carried too far becomes impractical. Modern pickup arms employing an offset have on the order of 1-degree tracking error.

A precision arm, manufactured by SME of England, is shown in Fig. 16-52A, with its essential components indicated. At A are two knife-edge bearings having a pivot with friction less than 20 milligrams in both vertical and horizontal directions. The arm B is of stainless steel, *wood-lined* to place the resonant frequency well below the reproducing range. Weight C is used to statically balance the arm longitudinally and laterally. Rider weight D adjusts the tracking force ¼ to 5 grams, in ¼ or ½ increments. Alignment is obtained by the use of a sliding base E through a 1-inch area, with alignment in the vertical plane up to ¾ inch. These latter adjustments permit the arm to be raised to the proper height, length, offset, and overhang. A small weight F provides a bias adjustment to prevent the arm from skating across the face of the record. Lever G provides a slow motion let-down for the pickup. Connections for the pickup cartridge are provided in the head I, with external connections at

Fig. 16-52B. Marantz Model SLT12 straight-line turntable and pickup. The movement of the pickup is tangent to the record groove.

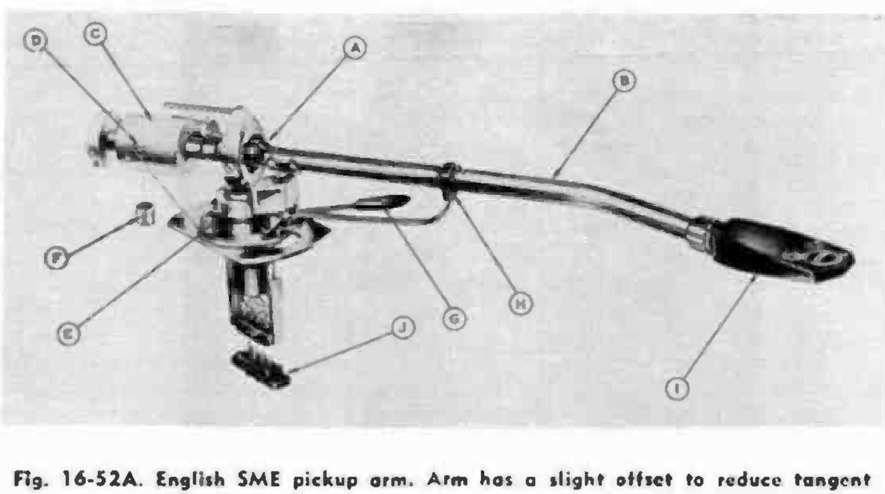


Fig. 16-52A. English SME pickup arm. Arm has a slight offset to reduce tangent error. (Courtesy, Shure Brothers Inc.)

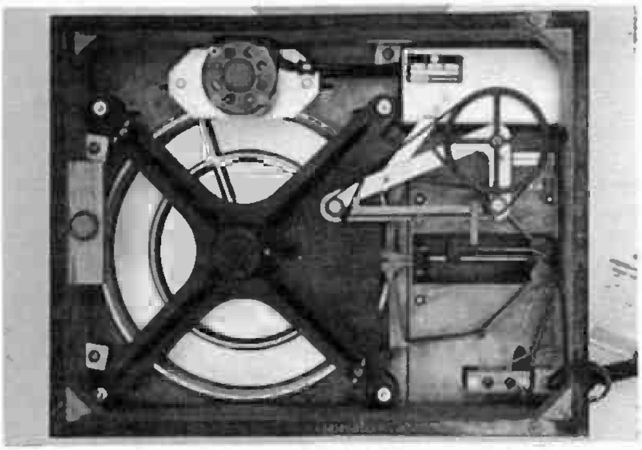


Fig. 16-52C. Underside view of Marantz Model SLT12 straight-line turntable, showing the pickup mechanism.

plug J. Tangent error is reduced to less than 1 percent by the use of an offset in the arm back of the pickup cartridge mount, and the arm placement at the rear of the turntable.

A turntable and pickup developed by Marants Co. Inc., designed to solve the tangent error, is shown in Fig. 16-52B. The pickup unit is guided across the face of the record in a *straight line*, so that the pickup is *always tangent* to the record groove, therefore the tangent error is *zero*. The turntable is belt driven by a hysteresis-synchronous motor. The turntable weighs 12 pounds, and is supported by a tungsten-carbide thrust bearing. Push buttons control the starting, stopping, and the dropping of the pickup on the record. The pickup employs an elliptical diamond stylus, with a compliance of 30×10^{-6} cm/dynes. The rumble is said to be 112 dB below a reference level of 7 centimeters per second. Two speeds are provided, 33⅓ and 45 rpm. An interior view, showing the driving mechanism for guiding the pickup across the record appears in Fig. 16-52C.

16.53 *What is the procedure for mounting a pickup arm?*—Generally the manufacturer of the arm supplies a template and mounting instructions for a particular arm. However, in the absence of such information, the pickup arm is mounted in such a manner that the tangent error is at a minimum. One method of mounting the arm is shown in Fig. 16-53. A template is plotted to indicate the inner and outer areas of modulation, and the arm so placed for a *minimum tangent error*. The procedure is the same for any length arm and diameter platen. It will be found, regardless of where the arm is placed, tangent error cannot be eliminated entirely. Several excellent methods of mounting arms are discussed in the references.

16.54 *What is overhang?*—The distance the stylus projects beyond the center pin of the turntable when the pickup arm is in such a position that a line joining the pickup stylus tip and the lateral-arm pivot passes through the turntable center pin.

16.55 *In what frequency range should the resonant frequency of a pickup arm fall?*—The design should be such that resonance at 30 Hz is avoided, as this is the rumble frequency for a four-pole motor. As a rule, the resonant frequency of a pickup arm on a record changer occurs between 30 and 40 Hz. Reproducing equipment for broadcasting and other professional use is designed to place the resonant frequency of the arm, below 10 Hz. For a number of commercial arms, resonant frequency is around 2 Hz.

16.56 *What is the effect if the magnetic structure in a pickup causes a drag on the turntable?*—The attraction of the magnetic structure of a pickup to a steel turntable can increase the stylus pressure and could cause the turntable to run slightly slow if the attraction is strong enough. The attraction of the pickup can be reduced by placing a pad between the pickup and turntable. In modern design, the attrac-

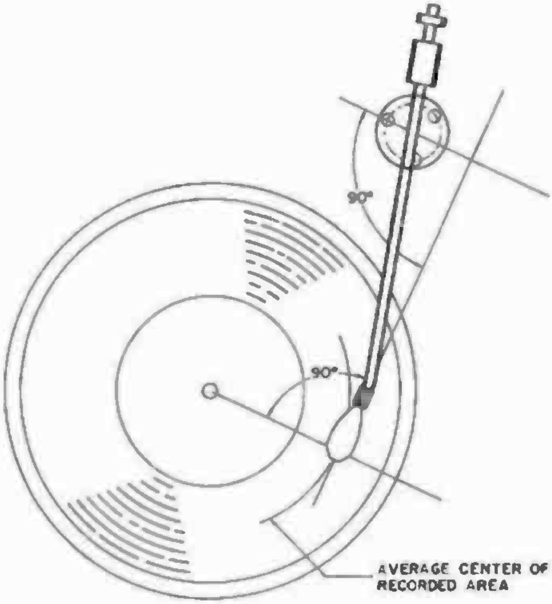


Fig. 16-53. Typical mounting for an offset pickup arm.

tion effect has been reduced to a negligible amount or eliminated completely. Some manufacturers of turntables use aluminum, which is nonmagnetic and load the turntable with a nonmagnetic material to increase its weight.

16.57 What is torsional resonance? —A form of resonance often found above 100 Hz in a pickup arm. This effect may be eliminated by making the arm of the greatest torsional rigidity for a given weight, such as by the use of rectangular tubing.

16.58 What is the procedure for adjusting the height of a pickup arm?— For a manual turntable, using a 15-degree stylus, both the pickup and arm must be parallel with the surface of the record, with the stylus pressure adjusted to its specified value. If this procedure is followed, the stylus will be at the correct angle and is the same for both monophonic and stereophonic cartridges. For automatic record changers, the adjustments are generally made for a height of three records. (See Question 15.60.)

16.59 Describe the effect of cartridge tilt on the electrical separation of a stereophonic pickup. — Separation, sometimes referred to as cross talk, is the ability of a pickup to reproduce the stereo effect. If the stylus is not vertical, but tilts to either the right or left side, the stereo effect is reduced considerably. The measurement is made, using a special test record with two grooves, one groove containing a frequency recorded on the left wall, and the other with the signal recorded on the right wall. First a measurement of the left wall is made and plotted. The second groove is then traced and plotted. The difference between the two measurements is plotted in dB separation.

Separation is generally stated at 1000 and 10,000 Hz and is most important in the range between 200 and 6000 Hz. A

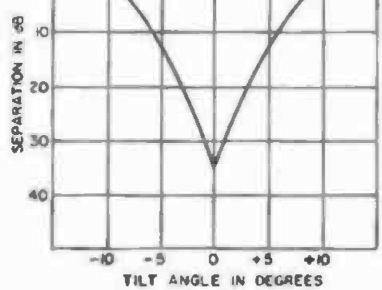


Fig. 16-59A. Separation versus stylus tilt at 1000 Hz (after Kogen and Samson).

separation of 10 to 15 dB in the higher ranges is considered adequate.

Fig. 16-59A indicates how the separation is affected by the tilt of the cartridge. Precautions should be taken to make sure the signal in each groove of the test record is exactly the same level, as it has been found that some test records do not have the same level in each groove. If this occurs, the measurement

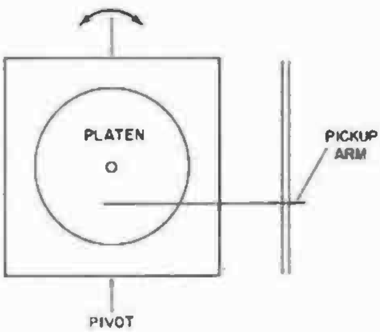


Fig. 16-59B. Turntable mounted on pivots for tilting at different angles when making vertical and tilt-angle measurements. The table is shown in the position for making vertical tracking angle tests. For pickup tilt measurements, the table is rotated 90 degrees (after Kogen and Samson).

will be in error. When the stylus is exactly vertical, the greatest separation is obtained. The turntable used for these measurements is supported on two pivots on a line running through the spindle center (Fig. 16-59B). The arm is mounted on a straight line support to permit it being placed in the groove with as near zero tangent error as possible.

16.60 Show how a 15-degree angle is applied to a reproducer stylus.—The 15-degree angle is applied to the stylus

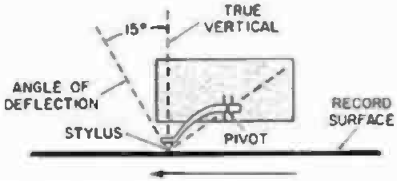


Fig. 16-60. Cross-sectional view of a pickup and stylus set to an angle of 15 degrees.

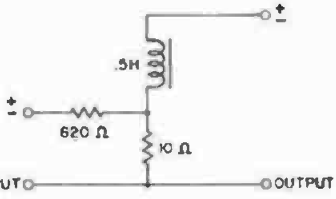


Fig. 16-61. Dummy circuit for magnetic pickup. This circuit may be substituted for a magnetic pickup when making measurements. The inductance must be encased in a Mumetal shield.

in a direction against the motion of the record, as shown in Fig. 16-60.

16.61 What is the electrical equivalent of an average magnetic pickup?—The electrical equivalent of a typical magnetic pickup is shown in Fig. 16-61. For measurement work involving crystal or ceramic pickups, a capacitor having a capacitauce equal to the stated capacitance of the pickup is substituted for the pickup element at the input to the amplifier.

16.62 Is it permissible to play back stereophonic recordings, using a monophonic pickup?—Stereophonic recordings should not be played back using a monophonic pickup as the stiffness in the vertical direction will damage the sound track. However, it is quite satisfactory to play back monophonic recordings using a stereophonic pickup. However, the turntable rumble may be increased, since most turntables have their greatest vibration in the vertical direction.

16.63 Describe the listening effect of playing back a monophonic recording using stereophonic reproduction.—If the pickup is connected in the stereophonic position, the effect is that of a quasi-stereophonic signal, as the signal is heard on both the left and right channels.

Combining the left and right channels to supply a sum signal to both channels simultaneously, the lateral signals are added in phase, while vertical turntable rumble components are out of phase, causing cancellation of the rumble frequencies. This is the most desirable condition for reproducing monophonic recordings using a stereophonic pickup. (See Question 16.62.)

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Section 17

Magnetic Recording

Although magnetic recording was invented in 1900 by Valdemar Poulsen, it is probable that the greatest advancement was contributed by German researchers who in 1928 developed recorder-reproducers which used a magnetically coated paper-base tape. By the end of World War II they had perfected not only a highly satisfactory cellulose base tape, but equipment capable of excellent sound reproduction for that time.

Factors contributing to the pre-eminence of magnetic recording are: portability and compactness of equipment; its freedom from dependence on laboratory processing; immediate playback; cost of the recording stock plus the ability to reuse tape; ease of storage; wide dynamic range; low distortion; and excellent sound reproducing qualities.

This section discusses transport systems, single and multiple sound-track heads, alignment, shielding, crosstalk, high-frequency bias current, types of magnetic tape and film, equalization, resolvers, looping systems, degaussing, and many different types of studio recording and associated equipment. Video-tape recording is discussed only briefly as it is not within the scope of this work.

The Standards quoted in this section are those in present usage. However, several are under review and may be changed in the near future.

17.1 What is magnetic recording?—Recording on a wire having magnetic qualities or on a paper- or plastic-base tape coated with a magnetic emulsion.

17.2 Define magnetism in simple terms.—It is a property found in certain materials which causes them to attract other materials. (See Question 17.21.)

17.3 What does the term "astatic" mean?—Having little or no magnetic properties.

17.4 What is reluctance?—Magnetic resistance. The opposition offered by air or certain substances to the flow of magnetic lines of force.

17.5 What is residual magnetism?—The amount of magnetism retained by a magnetic substance.

17.6 What is a ferromagnetic material?—A substance having a magnetic permeability greater than that of a vacuum and which varies with the applied magnetizing force. Examples of such materials are iron, nickel, cobalt, and numerous alloys.

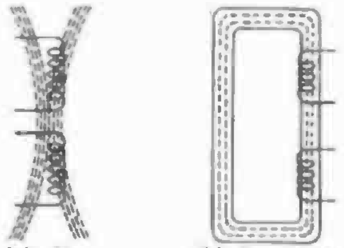
17.7 What is remanence?—The magnetic induction remaining in a magnetic substance after the applied magnetic force has been removed. Magnetic recording tapes having high remanence have a correspondingly high output.

17.8 Define retentivity.—The ability of a certain substance or substances to retain a magnetic charge. As an example, soft iron may be magnetized easily, but it quickly loses this property.

17.9 Define saturation in a magnetic substance.—The point of saturation has been reached when the number of flux lines in a magnetic core material reaches a point where an increase in current causes no additional magnetism, or decreases the magnetization.

17.10 What is coercive force?—The magnetizing force required to reduce the residual magnetism in a previously magnetized object to zero. The applied force must be of opposite polarity.

17.11 What is magnetic coupling?—A method of coupling devices by the use of coils and an iron core, or through their fields being adjacent to each other. Two examples of magnetic coupling are shown in Fig. 17-11.



(a) *Air.* (b) *Iron-core.*

Fig. 17-11. Magnetic coupling.

17.12 What are eddy currents?—Small currents flowing in the interior of a conductor caused by the movement of a magnetic field near the conductor. The action in the conductor, if it could be seen, would appear as swirling water. This term is generally associated with magnetic materials, although eddy currents may be induced in other materials not having magnetic properties.

17.13 What is an instantaneous polarity?—The magnetic or electrical polarity existing at a given instant in an electrical or magnetic circuit.

17.14 What are the names of the magnetic materials used for magnetic shielding and cores?—Conpernik and Hypersil, both manufactured by Westinghouse; Hypernik, manufactured by Allegheny Steel; Permalloy, manufactured by Western Electric; and Mu-metal, manufactured by Telegraph Construction and Maintenance Co., Limited. The foregoing are the most commonly used, although there are many others developed for express purposes.

To the audio engineer, some of the most useful of magnetic shielding materials are Netic and Co-netic alloys, manufactured by Perfection Mica Co. These materials can be obtained in sheets or rolls, from about 2 mils to 60 mils in thickness, and may be formed into any shape by bending and cutting, *without* affecting shielding capabilities. *No annealing after working is required.* The shielding capabilities for fields of 60 and 120 Hz is very high. For a 4-mil sheet, the attenuation for Netic is 6 to 12 dB, and for a sheet of Co-netic, 20 to 24 dB. These materials will also act quite well as an *electrostatic shield* in most instances and have the unique ability to be equally efficient at both high and low intensities. Interleaving of the two materials will result in an extremely high attenuation ratio. (See Question 8.50.)

17.15 What is Alnico V and VI?—The trade name of a magnetic alloy used in the manufacture of permanent magnets which is manufactured by several different steel companies. Such magnets are used in loudspeakers, light modulators, meters, and similar devices requiring a highly intense permanent magnet. (See Fig. 17-15.)

17.16 What is a core?—A core generally consists of a group of laminations, powdered iron, or a solid piece of magnetic material. Cores are used with coils such as an audio or power transformer to obtain magnetic coupling. When used in ac circuits, the core is laminated; that is to say, the core is made up of a number of thin magnetic iron sheets to reduce the effect of eddy currents. (See Fig. 17-16.)

17.17 Define the term "permeability?"—When certain types of magnetic materials are introduced into a magnetizing field, the number of flux lines in the material is greatly increased, and exceeds that out of the field many times.

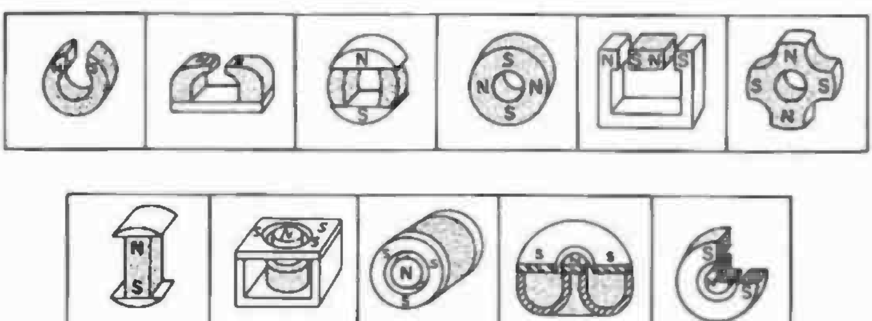


Fig. 17-15. Examples of permanent magnetic structures used in various devices requiring a permanent magnetic field. (Courtesy, "Magnetics," Indiana Steel Products Co.)

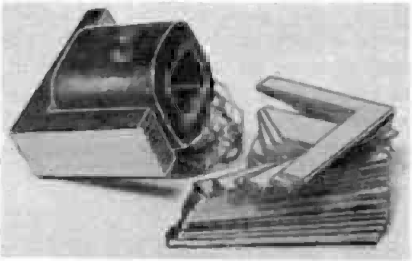


Fig. 17-16. A laminated core and the coil.

The permeability of a material is defined as the ratio of the number of lines of force which pass through a given area when it is occupied by a magnetic substance, to the number of lines of force passing through the same area when it is occupied by air or vacuum. The materials mentioned in Question 17.14 would all be classed as having high permeability. The symbol mu or (μ) is used to denote permeability.

17.18 What precautions should be taken when handling high-permeability materials?—Certain groups of materials cannot be bent or cut without reannealing. Others cannot be dropped or heated. The characteristics of the material should be determined by referring to the manufacturers data sheet before cutting or bending.

17.19 What is hysteresis loss?—Power lost in a magnetic core because of the internal friction caused by the molecules of the material. Hysteresis is caused by eddy currents in the core material and manifests itself in the form of heat.

17.20 Define the terms "gauss" and "oersted."—Gauss is the cgs electromagnetic unit of *magnetic induction*. One gauss represents one line of flux per square centimeter. The oersted is a unit of *magnetic intensity* (H) in the centimeter gram second (cgs) electromagnetic system of units. The value of magnetic intensity in oersteds at any point in a vacuum is equal to the force in dynes exerted on a unit magnetic pole placed at that point.

Hans Cristian Oersted, a Danish physicist, discovered in 1826 that a compass is deflected when placed near a wire in which there is a current flow, and also that a magnet exerts force on a wire carrying current. The gauss is named in honor of Karl F. Gauss, a German mathematician. In 1930 by international agreement, the term oersted replaced the term gauss.

17.21 What takes place when a bar of iron is magnetized?—A simple explanation of magnetic theory based on present knowledge is given in Fig. 17-21.

17.22 What is the rule pertaining to like and unlike poles of a magnet?—Like poles repel and unlike poles attract. (See Fig. 17-22.)

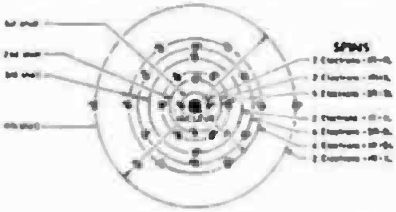
17.23 What is surface induction?—A quantitative measurement of the signal strength stored in a magnetic tape or film. The magnetic flux is measured on the surface of the magnetic emulsion where the signal is normally picked up by the reproducing head.

17.24 Describe the basic principles of magnetic recording.—Although there is no definite knowledge of just when magnetic recording was actually invented, it is known that a Danish inventor, Valdemar Poulsen, a telephone engineer invented a magnetic wire recorder and reproducer before 1900 and received a US Patent 661,619 Nov. 13, 1900, called the *Telegraphone*. However, the use of this device was hampered because of the lack of amplifiers and the available type recording media. In about 1927, a German inventor, Pfleumer, was experimenting with a powder coating, nonmagnetic-oxide, that deposited metallic particles on a tape. In 1930, a tape with a plastic backing was manufactured in Germany and after World War II was developed into its present state.

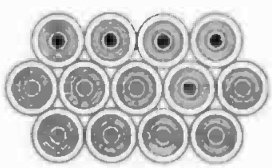
If magnetic material is placed in a magnetic field, the molecules of the material will be oriented with the direction of the magnetic field. Several methods may be used to produce this field, but to the audio engineer, the one of most interest is the field that is produced by a current flowing through a coil of wire with a core as used for magnetic recording.

Magnetic-recording media generally consist of finely divided particles of iron-oxide deposited on a plastic tape backing. (See Question 17.116.) During the recording process, the tape is pulled at a constant linear speed over a magnetic recording head containing a minute gap (Fig. 17-24A). While the tape is passing over the head gap, audio frequencies are applied to the head coil. Any particle of the magnetic medium crossing the gap is magnetized and remains in a permanent state of magnetization, which is proportional to the flux flowing through the head at the instant the particle passes over the gap. Thus, the actual recording takes place at the *trailing edge* of the recording head gap.

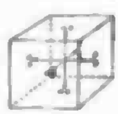
If the signal being recorded is of a sinusoidal nature, the intensity of the magnetization on the tape will vary simultaneously (Fig. 17-24B). A wavelength of recorded signal will occur for each complete alternation of the input signal.



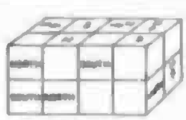
Physical concept of the inner structure of a ferromagnetic atom showing the electron arrangement necessary for the creation of magnetism.
The uncompensated, or off-balance, planetary spin of the electrons in the third incomplete quantum shell, together with specific dimensional characteristics creates a magnetic moment, or force.



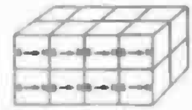
Magnetic moments in neighboring atoms are held parallel by quantum mechanical forces which can be likened to the forces holding the sun, moon, stars and earth in their relative positions.



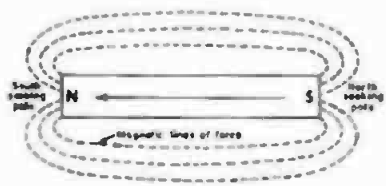
The atoms possessing these magnetic characteristics are grouped into regions called domains . . . A Domain is the smallest known permanent magnet. 6000 domains would occupy an area comparable in size to the head of a common pin.
A domain is composed of approximately one quadrillion (1,000,000,000,000,000) atoms . . . If an atom were the size of a ½ inch ball, then a domain would contain enough of these balls to surround the earth with a band 30 miles wide.



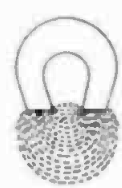
In unmagnetized ferromagnetic materials the domains are randomly oriented and neutralize each other.
HOWEVER, THE MAGNETIC FORCES ARE PRESENT!



Application of an external magnetic field causes magnetism in the domains to be aligned so that their magnetic moments are added to each other and to that of the applied field.
With soft magnetic materials such as iron, small external fields will cause great alignment, but because of the small restraining force only a little of the magnetism will be retained when the external field is removed . . . With hard magnetic materials such as Alnico a greater external field must be applied to cause orientation of the domains, but most of the orientation will be retained when the field is removed, thus creating a larger permanent magnet, which will have one North and one South pole.

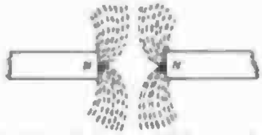


A freely suspended bar magnet will always tend to align itself with the North and South magnetic poles of the earth—for example—the magnetic compass.
This occurs because unlike poles of a magnet are always attracted to each other by invisible lines of force whereas like poles repel each other.

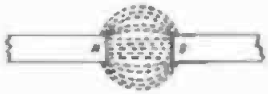


The horse-shoe shape is most commonly used in magnetic separators because its lines of force are more adaptable to the tasks which must be performed in the separation of ferrous from non-ferrous materials . . . A piece of iron placed within the effective range of a magnet will, in turn become magnetized. It will have its own North and South poles which will be attracted to the parent, or larger magnet in proportion to its mass.

Fig. 17-21. A simple explanation of magnetism. (Courtesy, Indiana Steel Products Co.)



(a) Repulsion of like poles.



(b) Attraction of unlike poles.

Fig. 17-22. Repulsion and attraction of two magnetic poles.

Wavelength is directly proportional to tape speed and inversely proportional to the frequency of the applied signal. Therefore,

$$\lambda = V/F$$

where,

λ is the recorded wavelength,
V is the linear speed of the tape in inches per second,
F is the frequency of the applied signal in hertz.

During playback the magnetized surface of the tape is passed over the gap of the reproduce head, which is similar in construction to the recording head. The portion of the magnetized tape in contact with the head gap is bridged by the magnetic core of the head, and the tape in its passage over the head gap causes magnetic lines of force to be induced into the core of the head, thus generating a voltage. The magnitude of this flux is a function of the average state of magnetization of the portion of the tape spanning the head gap at any given instants. As the tape passes over the gap, the amount of flux through the head core will vary with the changing state of magnetization on the tape, and cause a voltage to be generated in the head winding.

It is to be observed that the voltage generated is proportional not to the magnitude of the flux, but to the *rate of change*. Therefore, the playback voltage generated is dependent upon the frequency, and for constant-current recording (recording constant-current for all frequencies) the output voltage varies in direct proportion to frequency. The output voltage may be equated:

$$E = N\ (d\phi\, dt)$$

where,

E is the induced voltage,
N is the number of turns of wire in the head winding,
$(d\phi\, dt)$ is the rate of change of the flux.

Fig. 17-24C illustrates the reproduction of a relatively long wavelength on a tape passing over the reproducer head gap, and Fig. 17-24D shows a short wavelength equal to the reproducer head-gap height. In the latter circumstance, the average magnetization in the gap is zero and does not change with tape travel; therefore, the output of the head will be zero. The high-frequency response can be increased by either reducing the height of the reproducer head gap, or by increasing the linear speed of the tape. However, if this is carried too far, the voltage output from the head decreases to an unusable amount, and a compromise must be made. If the reproducing gap of 0.00025 inch is used and the tape transported at a linear speed of 60 ips, it is possible to reproduce a frequency of 100,000 Hz or 1600 sine wave cycles per inch of tape.

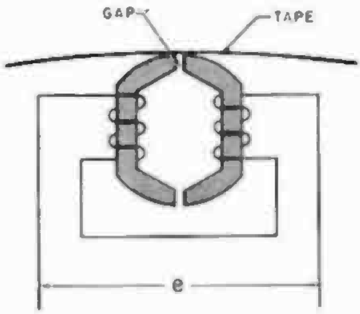


Fig. 17-24A. Tape showing how the magnetic field is induced in the tape.

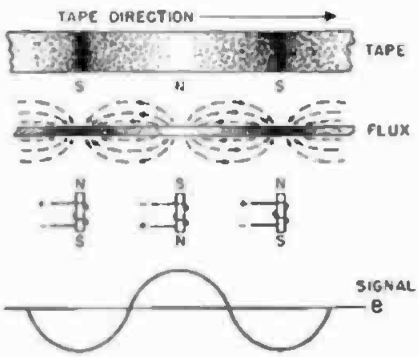


Fig. 17-24B. Basic construction of a recording or reproducing head.

This relationship is shown in Figs. 17-24E and F.

The dynamic range (signal-to-noise ratio) is the ratio of the maximum signal that can be recorded for a given harmonic distortion to the minimum signal which can be recorded. The minimum signal level is determined by the inherent noise level for the system. Stability relative to the amplitude is deter-

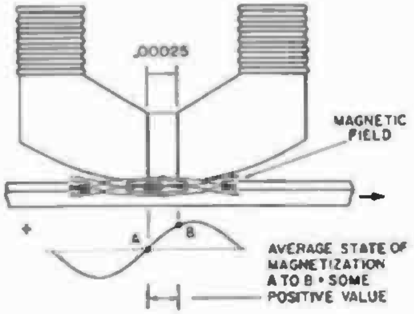


Fig. 17-24C. Showing the gap effect of a reproducing head when playing back a low frequency.

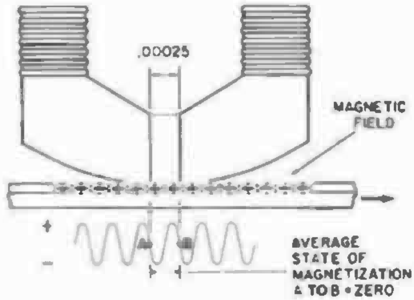


Fig. 17-24D. Gap effect when playing back a high frequency equal in wavelength to the head gap.

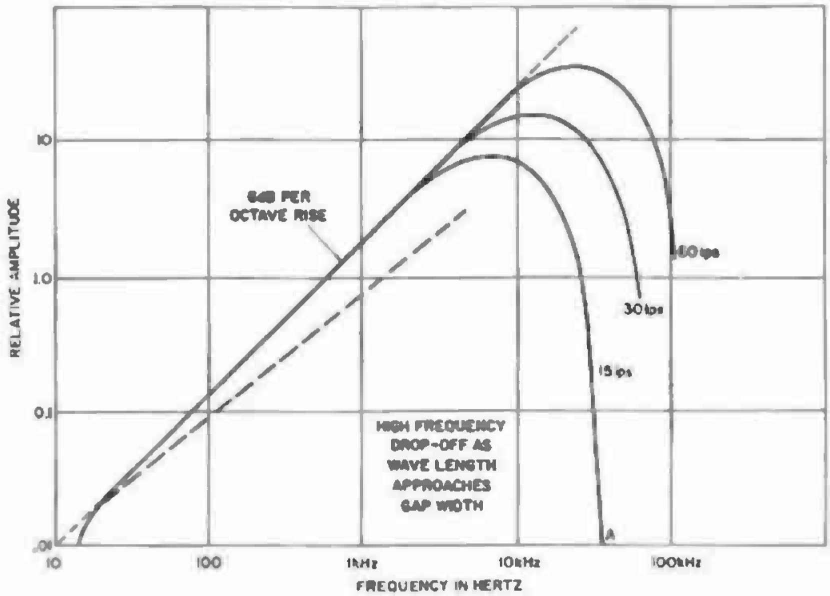


Fig. 17-24E. Linear speed versus frequency. Note the 6 dB per octave rise.

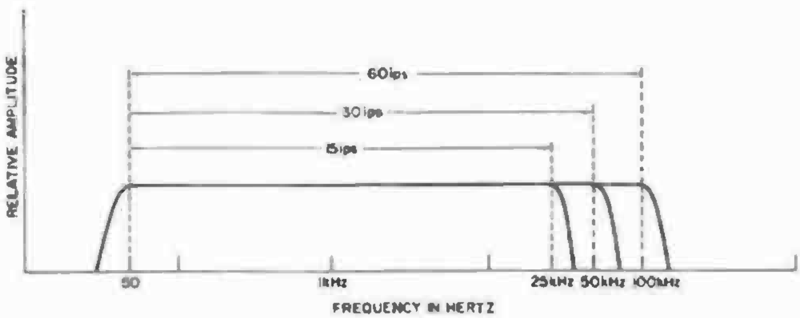


Fig. 17-24F. Linear recording speed versus the frequency response for a constant gap width.

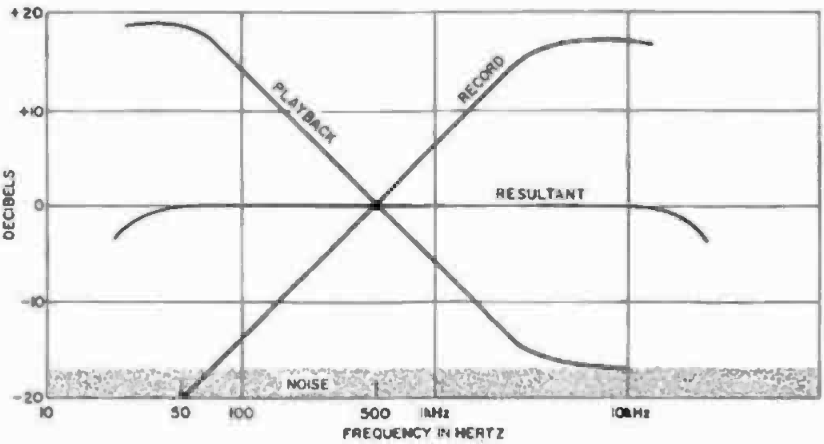


Fig. 17-24G. Basic scheme of equalizing magnetic tape.

mined by the surface of the recording media. If the surface is rough, large lapses or reduction of signal levels will result, causing dropouts. This is discussed in Question 17.159.

Because the magnetization curve of magnetic oxide is not linear, some means of reducing the distortion during recording is necessary. This is accomplished by the application of a high-frequency bias current to the recording head along with the audio signal, as explained in Questions 17.39, 17.40, and 17.44. It is highly desirable that the gap in the reproducer head be as small as possible, so that it will intercept less than one wavelength of the recorded signal on the tape at the highest frequency to be reproduced. Modern recorders use gaps of 0.25 mil to 38 millionths of an inch (0.000038 iuch). As previously mentioned, as the gap is reduced in height the voltage falls off rapidly. The voltage output may be equated:

$$E = B_m V \pi \omega / \lambda$$

where,

E is the induced voltage,
B_m is the maximum flux density of the recording media,
V is the linear velocity of the tape,
π is the gap height,
λ is the wavelength of the recorded signal.

To achieve an overall uniform frequency response, equalization is employed in both the recording and reproducing circuits. Low-frequency equalization is used in the *playback* circuits and high-frequency equalization is used in the *recording* circuits (Fig. 17-24G). The recording equalization compensates for the recording-head core losses and self-demagnetization of the short wavelengths approaching the head-gap dimensions.

It should be remembered that as the frequency is decreased, the output voltage from the reproducer decreases at a rate of 6 dB per octave, nntil it approaches the inherent noise level of the playback amplifier. At this level, it is impossible to recover the signal by further equalization. This situation may be termed the low-frequency limit of the system. Bandwidth is also important and is the determining factor in low-frequency reproduction, because the 6 dB per octave response starts with the highest frequency and continues to fall off at a constant rate, regardless of the tape speed, until the noise level is reached. Thus, the lowest frequency is dictated by the signal-to-noise ratio of the system. The effective *bandwidth* for magnetic recording is approximately 10 octaves.

The essential components of a magnetic recorder are: a transport system that will provide a constant linear speed in moving the tape over the recording and reproducing heads; an equalized amplifier system for both recording and reproduction that will result in a uniform frequency response; a high-frequency bias system of low distortion; and an erase head that will return the tape to its original state of demagnetization. The function of both recording and reproducing heads may be combined into one unit. The erasure head may be eliminated if bulk degaussing is available. (See Question 17.66.) Another very important effect noted in magnetic recording is that of the fringe effect at the reproduce head. This subject is discussed in Question 17.199.

17.25 What are the advantages offered by magnetic recording over photographic (optical) film recording? — For original sound tracks or rerecording, magnetic recording offers several advantages. They are: greater signal-to-noise ratio, wider frequency range, lower distortion, immediate playback, no processing required, the tape may be used many times over, and the recorded material may be transforred to other recording media without an appreciable transfer loss. The transfer losses per generation are quite small.

As a rule, except for stereophonic release prints, the final magnetic sound track is transferred to an optical sound track on the same base as the picture. Different methods of reproducing magnetic and optical sound track are discussed in Section 19. The transfer of magnetic sound tracks for release prints is discussed in Question 17.202.

17.26 What type winding is used with ¼-inch tape?—It is wound with the magnetic oxide side in (see Question 18.139). For recorded tapes to be stored for any length of time, the start of the program material should be on the inside end (hub).

17.27 What are the standard tape widths used for magnetic recording?—Normal ¼-inch tape is actually 0.246, plus-minus 0.002 inch. The thickness shall not exceed 0.0022 inch. Mylar tapes use a base thickness of 0.5 to 1.5 mils. Professional equipment may be designed to use ¼-, ½-, ¾- and 1-inch tapes. The 1-inch tape is generally used with three or four heads; however, up to eight tracks are also used with the 1-inch tape for special applications. (See Figs. 17-147D and E.)

17.28 What is magnetic film?—A standard motion picture film base (triacetate) is coated with a magnetic oxide. The base carries standard sprocket hole perforations for 16-mm, 17.5-mm or 35-mm film. The perforations in magnetic film are the same as for positive photographic film. Perforations in magnetic film are affected by both temperature and humidity. However, this condition is reversible and returns to normal when the film is placed in a temperature range of 60 to 80 degrees Fahrenheit, with a relative humidity of 40 to 60 percent. The perforation and pitch of magnetic film must not be compared to that of unprocessed photographic motion picture film, because the magnetic film employs a positive perforation pitch. Perforation pitch can only be measured using a calibrated perforation pitch gauge, or an optical comparitor. Both 16-mm and 17.5-mm magnetic film have sprocket holes on one edge only.

For the recording of motion pictures, it is essential that absolute synchronization be maintained between the picture camera and the sound recorder. The use of sprocket holes in the magnetic film base assures synchronization at all times, as both the camera and sound recorder are driven by a synchronous motor system.

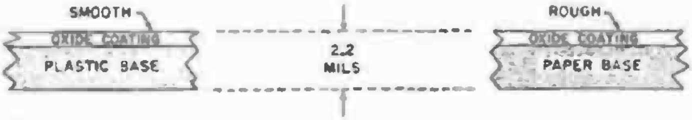
17.29 Why was paper discarded in favor of plastic as a base for magnetic tape?—Originally, all tape used a paper-base backing. Because the noise level of a tape is dependent on the smoothness of the base surface and the smoothness of the oxide surface, with the paper base being somewhat rough an uneven surface was reflected in the tape coating; also, the paper base was subject to moisture absorption and tore readily. Modern tapes employ a base of Mylar or cellulose triacetate from 0.5 to 1.5 mils in thickness. Plastic-based tapes have a noise level from 20 to 30 dB lower than a paper-based tape. The roughness of a tape surface causes recording noise and dropouts. The dropout, in decibels, may be calculated:

$$\mathrm{dB} = 54\frac{d}{\lambda}$$

where,

- d is the departure of the tape from the head in inches,
- λ is the recorded wavelength on the tape in inches.

The affect of dropouts is most serious at the higher frequencies, particularly where the nodules of foreign particles in the tape coating approach the recorded wave length. Most manufacturers of magnetic recording tape and film inject a lubricant into the magnetic oxide coating as it is being milled, to reduce head wear, noise, and dropouts, the result being a much higher quality recording.



(a) *Plastic base, smooth.* (b) *Paper base, rough.*

Fig. 17-29. Comparison of surface smoothness of a plastic-base tape to that of a paper-base tape.

Blow-ups of paper- and plastic-base tapes are shown in Fig. 17-29. It will be noted the surface of the paper base is reflected in the coating surface while the plastic-base tape is quite smooth. Unevenness in the base produces modulation noise. Noise may also be induced into the tape by the machines used for manufacturing the tape, particularly if any amount of mechanical vibration in the 40- to 60-Hz range is present.

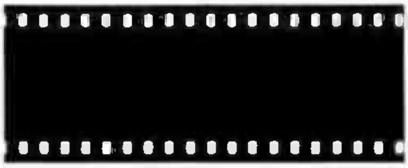


Fig. 17-30A. 35-mm full-coat magnetic film.

17.30 ***Describe the different types of magnetic film.***—Several different types of magnetic film are available; however, the ones most frequently used are shown in Figs. 17-30A through D. That appearing in Fig. 17-30A is 35-mm full-coat, tape used for music and multiple-track recording. It may also be obtained with a clear edge for identifying edge numbers on a picture. Fig. 17-30B shows 35-mm stripe tape, using a 300-mil sound track area, placed 200 mils from the edge and a 100-mil balance stripe at the opposite edge. The purpose of the balance stripe and the small stripe in the center is to prevent the film from warping. This film is used for editorial purposes and has a clear center for viewing a picture placed underneath. Fig. 17-30C shows a 17.5-mm full-coat film; it is the same as that in Fig. 17-30A except for the width. The 16-mm full coat film in Fig. 17-30D is available in either single or double perforations. Single perforation is used for *edge* recording, and double perforations for *center-track* recording, which is now considered obsolete. The sprocket tape of Fig. 17-30E is a standard ¼-inch tape, with a 2-mil base and 16-mm perforations. This tape was used before the advent of ¼-inch sync-pulse recording, discussed in Question 17.179.

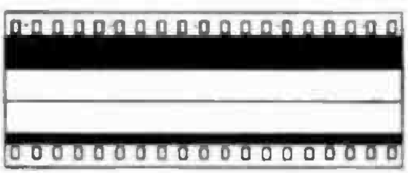


Fig. 17-30B. 35-mm stripe magnetic film.

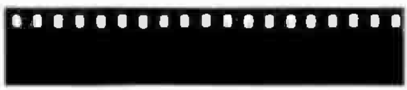


Fig. 17-30C. 17.5-mm full-coat magnetic film.

Both the 35-mm and 16-mm film may be obtained with a high output magnetic oxide, which increases the signal-to-noise ratio, as discussed in Question 17.164. The 35-mm magnetic film may also be obtained with a clear edge and full coating only between the inside edges of the sprocket holes. This is sometimes used to aid in editorial work.

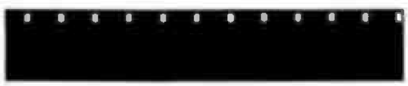


Fig. 17-30D. 16-mm full-coat magnetic film.

If the film with edge numbers is not available, it may be numbered by running it through an edge-numbering machine. Edge numbers are placed at one-foot intervals, for identification of a particular sound or action. When completely synchronized, the edge numbers

Fig. 17-30E. ¼-inch sprocket tape magnetic film.

of the sound track and picture are keyed together on a cue sheet for future identification. Magnetic film may be purchased with either an "A" or "B" wind, as discussed in Question 18.139.

Two grades of *striped film*, (1) new base and magnetic oxide, and (2) reclaimed base with new coating, are available. It is policy of most large studios to manufacture their own striped film by reclaiming print stock and clearing off the picture and striping. If the base is not damaged, it is quite satisfactory.

17.31 What are the standard speeds for ¼-inch tape recorders?—The NAB Standard (April, 1965) specifies the preferred speed to be 7.5 ips plus-minus 0.2 percent. Supplementary speeds are 15 ips and 3¾ ips. In addition to the above mentioned speeds are 1⅞ ips and 15/16 ips, although they are not mentioned in the Standard. Machines using these latter speeds are generally portable and are classed Special Purpose Limited Performance Systems. Also, 30 ips is frequently used for special recording.

17.32 What are the standard speeds for magnetic film? — Standard speeds are: for 16 mm, 36 fpm; 17.5 mm, 45 fpm; 35 mm, 90 fpm. In some instances the 35-mm magnetic film may be run 45 fpm.

17.33 What is the base thickness for magnetic film?—The base has a thickness of 4.5 to 5.0 mils, with a magnetic-oxide coating of 0.3 to 0.4 mil in thickness.

17.34 Give the minimum specifications for magnetic recorders.—

a. Frequency response—for semiprofessional and professional, as given in Figs. 17-162A and B. For limited use, as in Fig. 17-162C. (Also see Figs. 17-172 and 17-173.)
b. Signal-to-noise ratio—for limited use, given in Fig. 17-159B.
c. Signal-to-noise ratio—professional equipment, as in Fig. 17-159B. It should be remembered that values given for signal-to-noise ratios are the minimum values. Most professional machines fall between 58 to 65 dB, and some types 80 dB. Signal-to-noise measurements are generally made using a weighted network, as discussed in Question 5.98.
d. Flutter—for professional machines the total unweighted flutter is not to exceed 0.15 percent at 15 ips, and 0.2 percent at 7½ ips. Using a weighted network, these values become 0.05 and 0.07, respectively. However, most professional machines do not exceed 0.10 percent. This type measurement includes the flutter contributed in both the record and reproduce modes. For limited service, an unweighted curve is generally used, and the flutter is not to exceed 0.3 percent.
e. Distortion—to be less than 3-percent THD at 400 Hz, including that contributed by the tape, recorded at a level that will be reproduced 6 dB above the standard NAB recording level. (See Questions 17.139 and 17.140.)

Flutter measurements may be made in two different ways. First, record a 3000-Hz signal, then play the tape back and measure the flutter; second, use a flutter tape and measure the flutter. Since standard flutter tapes are recorded at a frequency of 3000 Hz, with not more than 0.03 percent total rms flutter, these tapes should not be rewound for storage.

17.35 What are the minimum specifications for magnetic film recording systems?—

a. Signal-to-noise ratio—60 to 65 dB below a total harmonic distortion (THD) of 3 percent at 400 Hz.
b. Frequency response — for 16-mm film (36 fpm), see Question 17.174.
c. Frequency response—for 17.5-mm (45 fpm) and 35-mm (90 fpm), see Question 17.175.
d. Harmonic distortion—although the specifications (a) state 3-percent THD, it is the policy of most motion picture sound departments to operate as closely to 1-percent THD as possible, and less in some instances.
e. Flutter—0.05 to 0.15 percent total.
f. Running time—absolute synchronism.

17.36 What is perpendicular magnetization of the recording medium?—A method of magnetizing the recording medium perpendicular to its direction of travel as shown in Fig. 17-36. The recording medium is magnetized by varying the intensity of the applied magnetic field. The maximum frequency that may be recorded is limited by the thickness of the pole pieces.

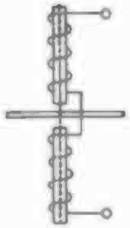


Fig. 17-36. Perpendicular magnetization.

17.37 What is longitudinal magnetization of the recording medium?—The longitudinal method of recording (Fig. 17-37A) on magnetic tape is the one most commonly employed in commercial recording of tape and magnetic film, except that the actual design of the head is a *ring* as shown in Fig. 17-37B. The maximum frequency that can be recorded is limited by the size of the head gap, and the linear speed of the tape.

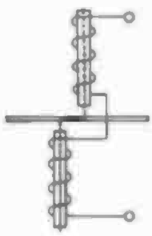


Fig. 17-37A. Longitudinal method of magnetization.

17.38 What is transverse magnetization of the recording medium?—Magnetization of the recording medium perpendicular to its direction of motion and parallel to the greatest cross-sectional dimensions in tape from edge to edge. (See Fig. 17-38.)

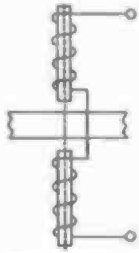


Fig. 17-38. Transverse magnetization.

17.39 Describe the X-field (cross-field) method of recording.—The X-field or cross-field method of recording on magnetic tape is a development of Marvin Camras of the Illinois Institute of Technology Research (formerly the Armour Research Foundation of Illinois Institute of Technology). The purpose of the cross-field head design is to improve the recording resolution by sharpening the critical zone, especially

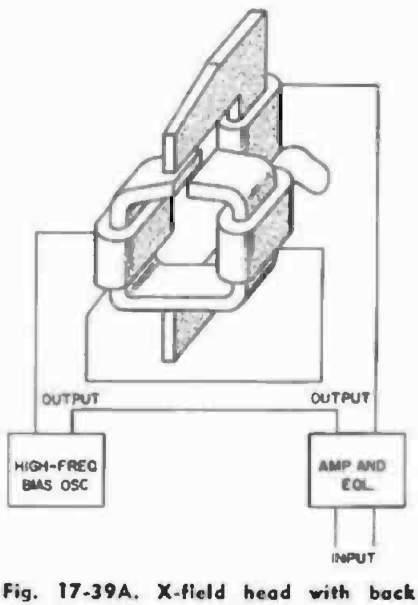


Fig. 17-39A. X-field head with back pole-piece. (Courtesy, Illinois Institute of Technology Research Institute.)

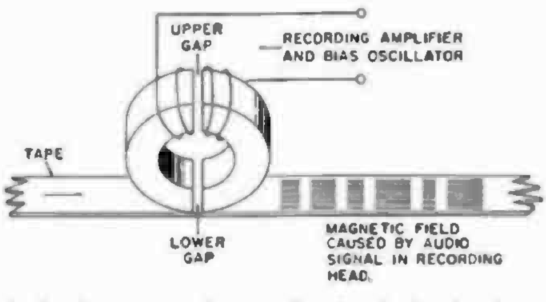


Fig. 17-37B. A ring-type magnetic recording head, showing how the head induces magnetic pulses in a magnetic tape or film.

at the bottom of the surface of the tape layer adjacent to the head gap. By adding a vertical field to the semicircular field of the conventional recording head, the resultant is more intense at one edge of the gap, with a sharper gradient at the other gap.

Fig. 17-39A shows the design of a X-field head. The core, with its two coils, resembles a conventional head with a very small gap over which the tape rides. A second core with a single coil and a back pole piece overhanging the gap, separated by enough distance to allow the tape to be conveniently threaded over the head, has been added. The three coils are energized with high-frequency bias current and audio signal, therefore, the field intensity varies with time, preserving the space-field pattern. Special heads using this principle have been developed with a 1.25-micron gap, for use with relatively deep magnetic layers, where high field densities are required.

A cross-sectional view of an X-field head, designed for quarter-track stereophonic use is shown in Fig. 17-139B. A two-gap erase section precedes the X-field gap and the recording playback gap. The X-field gap is energized by the erase winding, so that only two coils are required per channel. Elimination of the separate erase head compensates for the additional X-field structure, making the overall design cost comparable to heads of conventional design. This design is particularly adaptable to linear tape

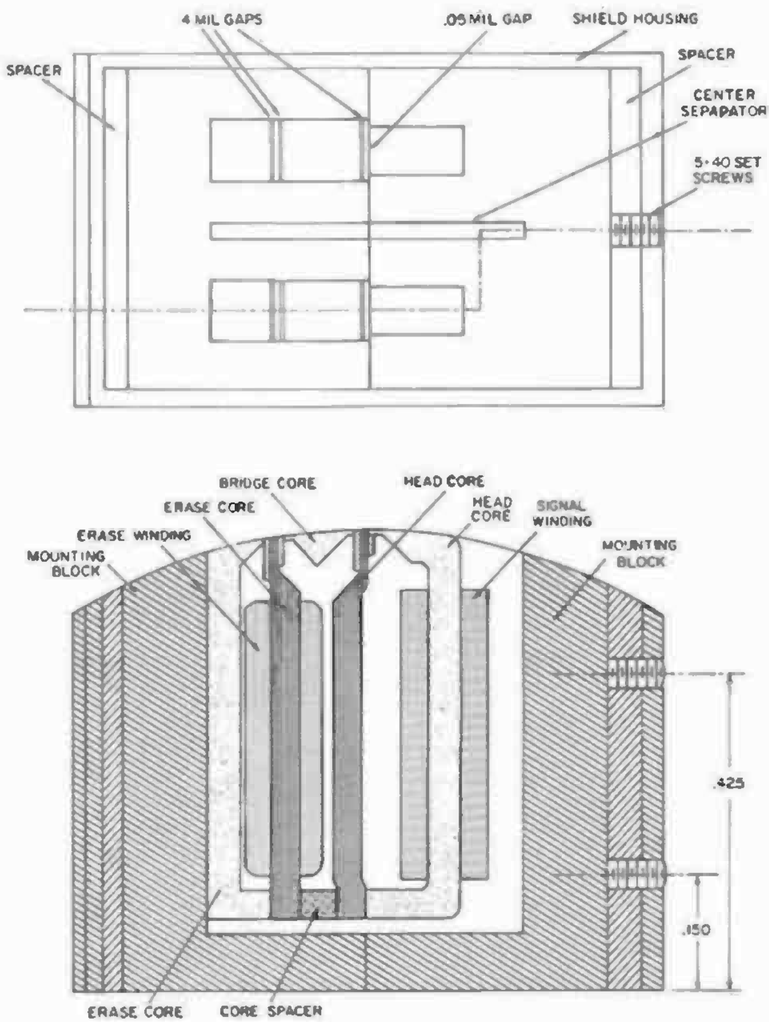


Fig. 17-39B. X-field head designed for stereophonic recording and reproduction. (Courtesy, Illinois Institute of Technology Research Institute)

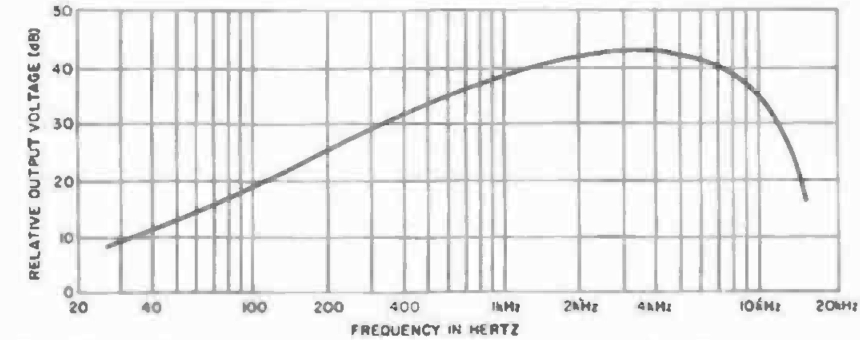


Fig. 17-39C. Constant-current response of X-field head at 1⅞ ips. Recording current 0.7 mA, played back through an amplifier having flat-frequency characteristics (no equalization).

speeds of 1⅞ ips and lower. A plot of the head characteristics using a linear speed of 1⅞ ips unequalized, with a constant recording current of 0.7 milliampere is given in Fig. 17-39C.

A somewhat different type X-field recording head is used in the Roberts Model 770A recorder-reproducer described in Question 17.222. Here the X-field system employs three heads, namely the erase, record-reproduce, and a separate head for the recording bias current. This latter head provides only the recording bias (Fig. 17-39D) and swings out of the way during playback (Fig. 17-39E). In operation, the bias head is never in actual contact with the tape but is spaced 0.110 inch from the tape surface; thus, the recording bias is supplied only with the audio signal in the recording head.

The bias-current head gap has a length of 250 mils and a height of 120 mils. The record-reproduce head is a two-head in-line cluster, constructed of NC-88 steel and employing a gap height of 38 millionths of an inch (0.000038), with a length of 0.43 inch. This extremely minute gap height is necessary to record and reproduce the very short wavelength of 13,000 Hz at linear tape speeds of 1⅞ ips.

In the conventional recording system, the passage of the tape over the recording and reproduce head wears down the pole pieces, changing the gap dimensions which, in turn, affect the bias current and thus the recording of high frequencies. Since the Roberts X-field method of recording bias is applied from a separate head not in actual contact with the tape surface, the effect of head wear on the recording bias current is eliminated.

17.40 Describe the focus-gap method of recording.—The focus-gap head is a development of D. P. Gregg and Keith O. Johnson, of Fairchild Recording Equipment Corp., designed to increase the dynamic range of magnetic recording. The design is based on the *anhysteretic process,* in that the bias mag-

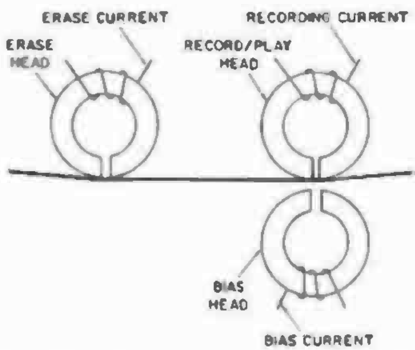


Fig. 17-39D. During recording, the bias head is brought near the tape but does not make actual contact.

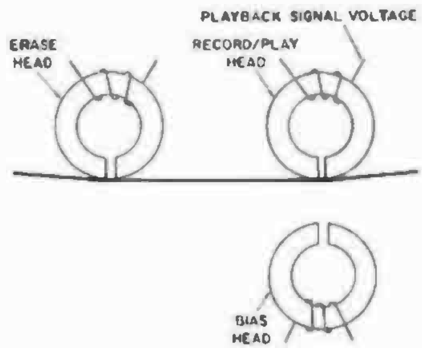


Fig. 17-39E. During playback, the bias head is cut off and swings away from the tape.

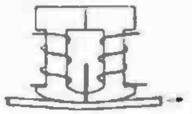


Fig. 17-40A. Focused-gap magnetic recording head.

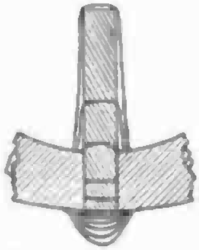


Fig. 17-40C. Close-up of focused-gap head.

netic field is made to decay at much faster rate than the audio-frequency magnetic field. It is claimed for this head that the low-frequency output and high-frequency saturation levels are increased by 6 dB. The signal-to-noise ratio for conventional recording is limited by the bias-induced noise and the amplifier system. In this new technique, the bias and even-ordered harmonic components from the bias oscillator are prevented from recording on the tape surface. The bias-oscillator frequency is increased to around 4-megahertz, and even higher. This prevents recording of the bias frequency because the magnetic-oxide particles are approximately the same physical length as the bias wavelength on the recording media. In addition, the even-ordered harmonics of the bias frequency are suppressed by a bias oscillator buffer amplifier.

The construction of a focused-gap recording head is shown in Fig. 17-40A; construction of a conventional head is shown in Fig. 17-40B. Enlarged views of a conventional and focused-gap heads are given in Figs. 17-40C and D.

A focus-gap head has a reduced reluctance path around the core to reduce losses at the bias frequency. A conductive material, which behaves as a shorted turn in a transformer winding, is placed in the head gap. With the bias frequency operating in the megahertz range, this gap generates strong secondary currents. The resulting magnetic field opposes the core flux, since the shortest path of least opposition occurs outside the conductive path. The bias flux lines around the gap region are distorted and resemble a beam profile which is similar to that generated by the conventional head gap. As the tape travels over the gap, the bias field decays at a faster rate than the audio signal.

17.41 Describe the basic principles of a transport system for a ¼-inch tape recorder and a magnetic film recorder.— Referring to Fig. 17-41A, a transport system for a ¼-inch magnetic tape recorder, the tape A leaves the supply reel and passes over a tension arm B and flutter filter C, which has a fly-wheel mounted at the rear. Leaving the flutter filter the tape passes over the erase, record, and reproduce heads D, E, and F. The heads generally are mounted on a mechanical arrangement that can be moved downward when the tape is in motion to apply pressure between the heads and the tape to assure proper head contact. At G is a steel capstan mounted on the motor shaft. Above the capstan is a neoprene puck H termed a pinch-wheel, that applies pressure to the tape which contacts the capstan and pulls the tape through the transport system, over take-up tension arm I, and up to the take-up reel.

Semiprofessional machines usually use a single motor which pulls the tape, and by a group of friction idler wheels, operates both the supply and take-up reels. In machines designed for professional work, three motors are generally used — two permanent *split-capacitor torque motors* for the supply and take-up reels, and a *multispeed hysteresis-synchronous* motor for driving the tape

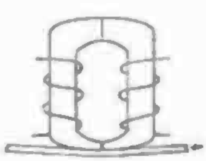


Fig. 17-40B. Conventional recording head.

Fig. 17-40D. Close-up of conventional gap head.

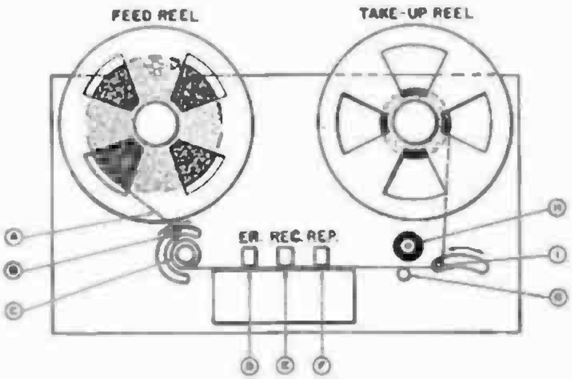


Fig. 17-41A. Simplified transport system for ¼-inch magnetic tape recorder.

capstan. (See Question 3.70.) The torque motors apply constant tension to the tape as it leaves the feed-reel (slight backward pull), and to the take-up reel.

In the record or playback position, the torque motor, although rotating *counterclockwise*, applies a clockwise torque to hold back the tape and apply an even tension as it unwinds. The take-up motor is rotating counterclockwise with sufficient torque to supply the correct tension to the tape for a smooth wind. In the rewind position, the supply motor has a high torque in a clockwise direction, and the take-up motor has a low torque in a counterclockwise direction. When thrown to the fast forward position, the supply reel rotates counterclockwise, but with low torque in the reverse direction, while the take-up reel moves counterclockwise with a high torque. Electromechanical brakes mounted on the torque motors hold and stop the supply and take-up reels to prevent the reels from spilling the tape.

Two methods are used for operating the capstan motor. In the first method, the motor runs continuously; in the second method, the motor rotates only when the start button has been depressed. In either design, the pinch wheel is moved downward, forcing contact between the tape and the capstan. The purpose of lever I is to apply tension to the tape as it winds. Other devices, such as automatic stop if the tape breaks, and record iuterlocking and editing controls are bnt a few of the added features which will vary with different manufacture.

A second type capstan drive system termed a dual-capstan or differential capstan drive, is illustrated in Fig. 17-14B. Certain designs of this system rely on the reeling functions to establish the tape tension over the head areas, similar to a close-loop system to be described later. However, most transport systems of dual-capstan design turn each capstan at a slightly different speed and establish the tape tension with differential action. This system has the advantages of less external tension required at the head area, and better isolation from the feed and take-up reel is afforded. This system is also referred to as a dual-capstan closed-loop system.

Another type transport system eliminates the pinch-roller completely by employing a capstan with a high surface friction and a large tape-wrap angle. When used with a dual-capstan drive (differential), better isolation from reel-

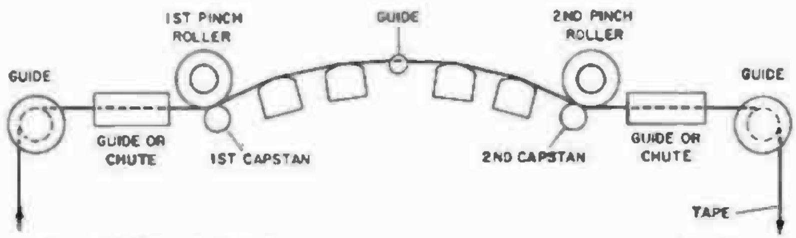


Fig. 17-41B. A ¼-inch tape transport system using two pinch rollers.

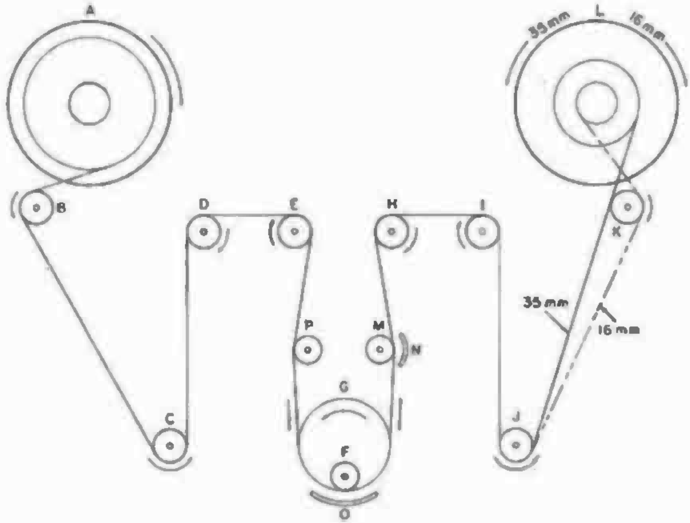


Fig. 17-41C. Davis tight-loop transport system used with magnetic film recorders and reproducers.

ing perturbations is obtained. Variations of this system appear in models of different manufacture. Removing the pinch wheel (pinch roller) removes one more source of irregularity in tape speed, and tape guiding is eliminated. Pinch rollers with surface deformities, slick and sticky surface spots, or noisy bearings all contribute to tape irregularity (flutter).

The transport system shown in Fig. 17-14C is a tight-loop system, developed by Charles C. Davis of the Westrex Corp., and is used in both magnetic and photographic (optical) recorders. Starting at the film feed-reel A, the film passes over idler B, to a second idler C, which lines up the film with the driven sprocket D. Leaving this sprocket, the film passes over idler E, which is a part of the tight-loop filter system. The film then passes over roller P, which may be replaced with an erase head, if desired. Leaving the roller, the film passes around the lower surface of impedance drum G, which has a heavy fly-wheel on the opposite end of its shaft. Magnetic recording head F is mounted in such a manner that it overhangs the impedance drum and makes contact with the magnetic oxide on the magnetic film. Leaving the drum the film then passes over a monitor head M, and then to idler H, which is also a part of the tight-loop system. From this idler, the film contacts driven sprocket I, then to idlers J and K, and to take-up reel L. A Mumetal shield O prevents stray magnetic fields from being picked up by the recording head. Monitor head M is also protected in a similar manner by shield N.

Idlers E and H are mounted on a cantilever at the rear, connected with a system of springs and an air dash-pot. As the machine is brought up to speed, idlers E and H oscillate in a vertical plane until the system settles down to a steady linear speed. The inner end of impedance drum G carries a heavy fly wheel to iron out irregularities in the speed of the film, thus reducing the flutter to a minimum. It is not too uncommon to have a total flutter (rms) of 0.05 or less with this type of transport system.

In some types of photographic recorders a free-loop system is used, which is not spring loaded. This system is discussed in Section 18.

17.42 Show the mechanical design of a ¼-inch tape recorder.—The front panel view of a Model 1024 Magnecord stereophonic quarter-track tape recorder, manufactured by Midwestern Instrument Inc., is shown in Fig. 17-42A, with the principal components indicated. The tape leaving the supply reel A, passes over tape-break compliance arm B and stabilizer roller C to the head assembly D, where it encounters erase, record, and reproduce heads. At

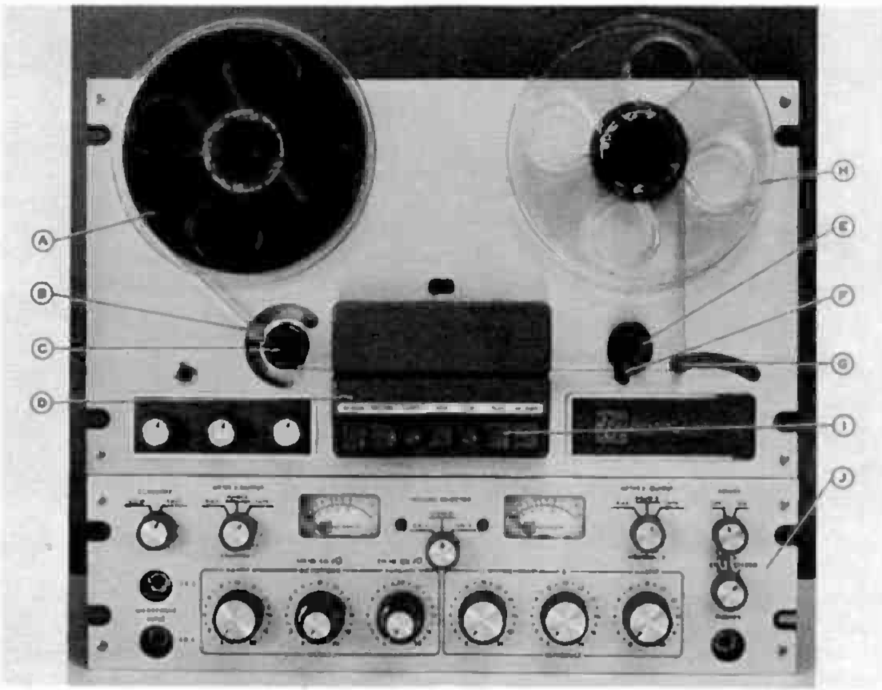


Fig. 17-42A. Front panel view of Magnecord Model 1024 quarter-track stereophonic ¼-inch magnetic tape recorder.

E is a pinch wheel; and at F a capstan. The tape is pinched between these latter two items and pulled over the head assembly. Take-up compliance arm G also provides a guide for the tape on its travel up to the take-up reel H. Items I are push-buttons for selecting the various modes of operation. Panel J carries the VU meters and various controls.

An interior view with the front panel removed is given in Fig. 17-24B, showing the mechanism associated with the front panel controls and the transport system. Starting at A is the supply-reel

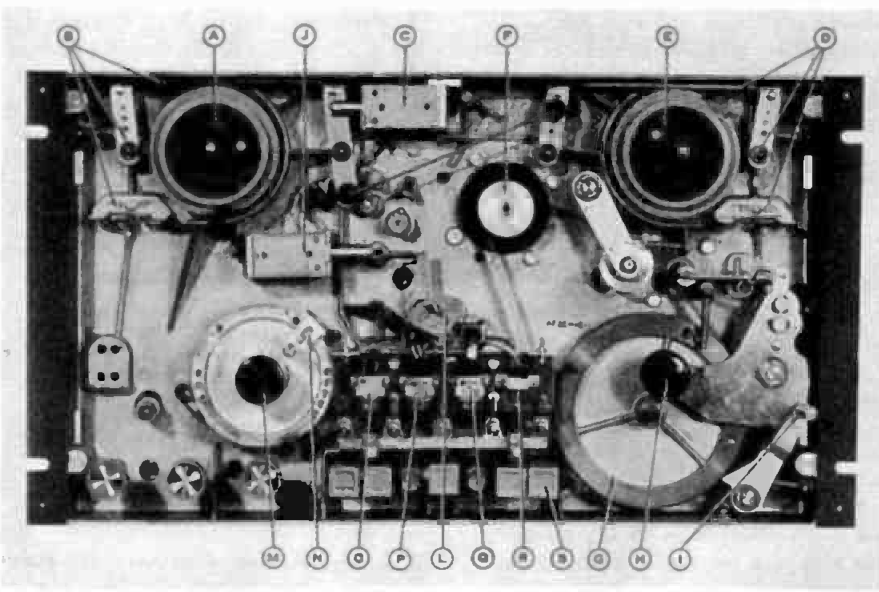


Fig. 17-42B. Front interior view of Magnecord Model 1024 recorder.

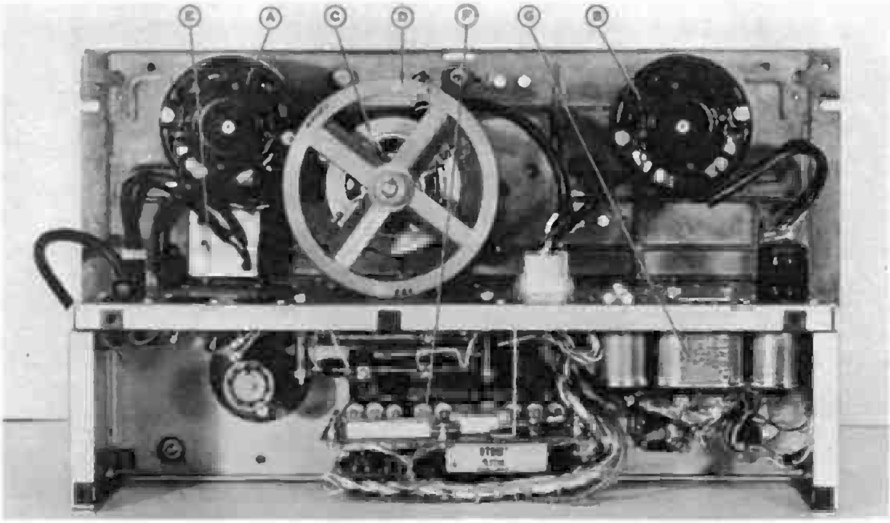


Fig. 17-42C. Rear interior view of Magnecord Model 1042 tape recorder.

spindle and turntable with its braking mechanism B, which also is linked with the braking mechanism D of take-up spindle E. A pulley F mounted on a two-speed capstan motor helt drives the capstan flywheel G. The pinch-wheel mechanism H is solenoid-controlled in both the record and playback modes. The take-np compliance arm is shown at I. Solenoid J raises and lowers the head assemhly on the tape. When the transport system is at rest, the heads are pulled up and away from the tape. The fourth head K provides a means of playing back half-track prerecorded stereophonic tapes and is activated hy changing the position of level L. Item M is the stabilizer roller, N is the tape-break compliance arm, O, P, Q, and R are the heads, and S is the control push-button assembly.

The rear of this same section is pictured in Fig. 17-42C. At A and B are two split-capacitor torque motors for the supply and take-up reels, and at C is the capstan drive motor consisting of a two-speed hysteresis-synchronous motor with a large fly-wheel D. Item E is a solenoid for operating the pinch-wheel, and at F is a group of resistors for adjusting torque motors A and B. The motor-starting capacitor is shown at G.

17.43 What effects do the capstan and pinch wheel have on the reproduction of a tape recorder?—Both the capstan and pinch wheel have a pronounced effect on the reproduction, because the linear speed of the tape is dependent on both these items. The capstan and pinch wheel should be cleaned frequently, as both pickup dirt and small pieces of magnetic coating which can cause slight irregularities in the linear speed. Several cleaners are available for this purpose. Any slippage between these two components can cause the machines to run slow and, in some instances, induce a pronounced flutter.

17.44 Describe the use of bias current and its purpose.—The normal characteristic of magnetic tape or film used for recording sound is quite nonlinear. It was discovered by W. I. Carlson and G. W. Carpenter of the General Electric Co., in 1927, that if a high-frequency current of several times that of the highest frequency to be recorded is applied to the tape along with the audio

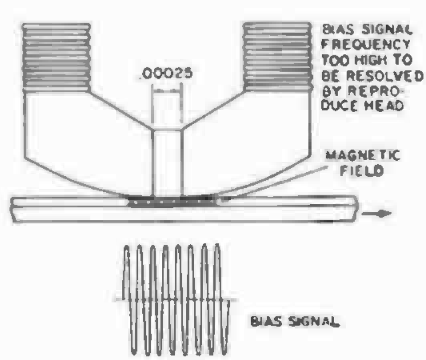


Fig. 17-44A. Reproduce head gap showing an audio signal and bias current combined on the tape.

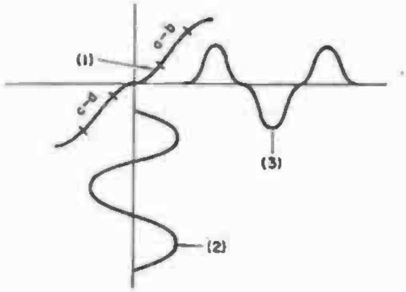


Fig. 17-44B. Characteristic of magnetic film or tape without bias current.

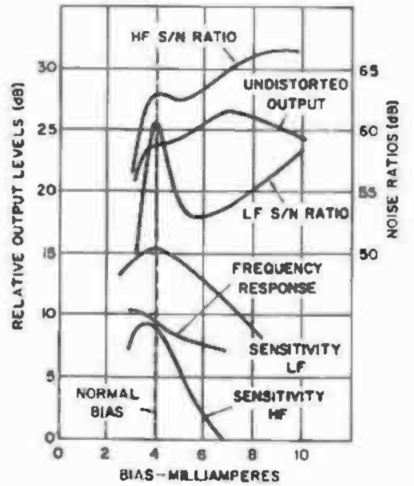


Fig. 17-44D. Characteristics dependent on the bias adjustment.

signal, the result is a signal of low distortion with a considerable increase in the signal-to-noise ratio. Although the signal is referred to as bias current, it is not really a bias current but only a high-frequency current added to the audio signal to place the audio signal on the linear portion of the B-H curve. There is *no modulation of either frequency by the other*. When the optimum ratio of the bias current to the audio signal is reached, the output from the tape or film is maximum, distortion lowest, with the greatest signal-to-noise ratio.

The combining of the bias and the audio signal is accomplished by linear mixing, without sum and difference frequencies being generated. The application of a bias current is not an amplitude-modulation process, as the bias frequency does not enter into the recording or playback process. Since the wavelengths of the bias frequency are small, they are not resolved by the playback head. An illustration of how the reproduce head sees the bias frequency is given in Fig. 17-44A. It will be observed that a wavelength of the bias frequency is considerably shorter than the head gap.

The theory of high-frequency bias recording is quite complex; however, a rather simple explanation can be illustrated by the cnrves in Fig. 17-44B. Curve (1) represents the magnetic-oxide response. Curve (2) is a pure sine wave applied to the input. The output curve (3) as will be noted, is distorted at the zero crossover point, because of the characteristic of the magnetic oxide. Although curve (1) is not linear, when taken over its entire length it does have portions that are linear, (a to b) and (c to d). If the signal can be recorded on the linear portion of these curves, the distortion is lowered. This is accomplished by the application of a high-frequency bias current.

The bias current is a minute amount of high-frequency alternatiug current, independent of the audio signal, and is applied to the recorder head. The bias current moves the audio signal to tbe central area of the oxide characteristic into the linear portion of the curve

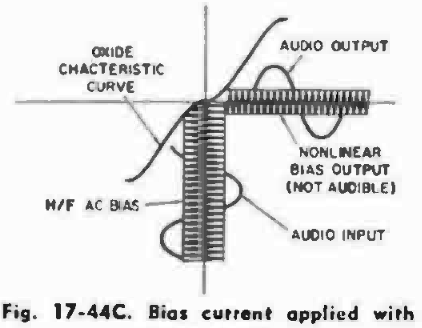


Fig. 17-44C. Bias current applied with audio signal.

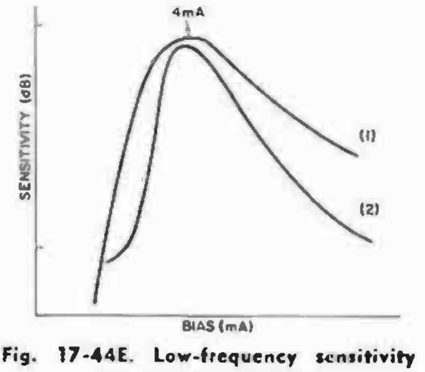


Fig. 17-44E. Low-frequency sensitivity versus bias current.

(Fig. 17-44C). The amount of the bias current is quite critical and will vary with different type recording heads and tape manufacture. Bias current affects the sensitivity of the tape at both the low and high frequencies, as well as the frequency response, distortion, output, and signal-to-noise ratios. This relationship is shown in the curves of Fig. 17-44D. The proper values of bias current may be determined by the method described in Question 17.52.

Curve (1) of Fig. 17-44E is a plot of low-frequency sensitivity, centered around a bias current of 4 milliamperes for a given brand of tape, while curve (2) is for a second brand of tape. It will be noted that curve (1) is much broader, therefore, the bias setting is less critical. Curve (2) has a sharper peak and is more critical to the bias setting, and will require a stable bias supply to prevent slipping off the peak current. With the correct bias setting, the output waveform is almost linear, therefore the distortion will be quite low. Equalization is required in both the recording and reproducing amplifiers, to obtain a uniform record-reproduce characteristic. The above method of recording on magnetic tape or film is termed the *direct method of recording*.

17.45 What are the frequencies employed for high-frequency bias current?—The frequency of the bias oscillator should be at least five times that of the highest frequency to be recorded to prevent beats between the bias frequency and the highest audio frequency. Professional recording equipment employs frequencies between 50 and 250 kHz. This frequency must be of low harmonic distortion.

17.46 What are the characteristics of unequalized magnetic tape and film?—The coating used on magnetic tape and film is the same. However, because of the method used to record—and the linear speed—to obtain a uniform frequency response during playback preequalization is required during recording and postequalization when reproducing. Typical tape characteristics are shown in Fig. 17-46 for unequalized playback using a constant-current recording characteristic, for linear velocities of 5.1 to 24 ips.

17.47 How is the high-frequency bias oscillator coupled to the recording head?—As shown in Fig. 17-84. The oscillator may be push-pull or single-ended with a coupling transformer in its plate circuit for coupling to the recording head circuit. A series capacitor is connected in the output to resonate the head circuit for maximum bias current. The recording head is connected to a potentiometer to control the amount of bias current fed to the head. A parallel-resonant circuit is connected across the recording amplifier to prevent the bias current from entering the output circuit

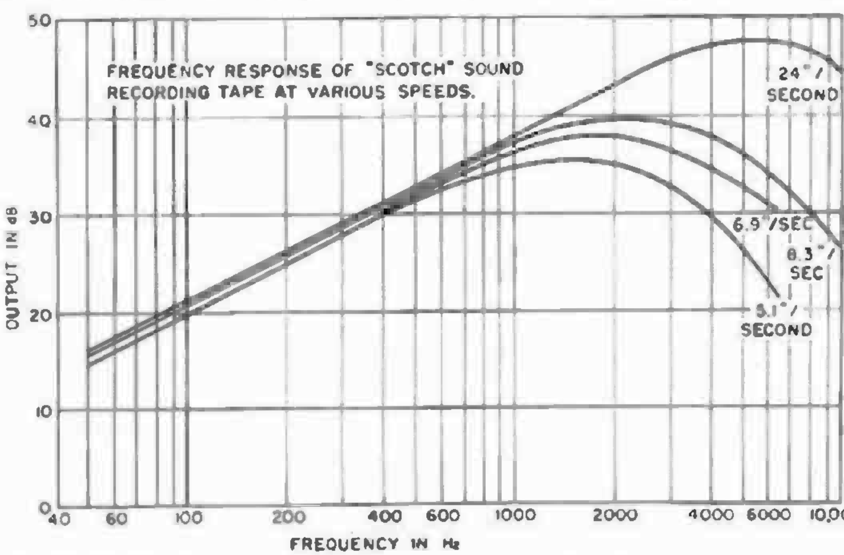


Fig. 17-46. The frequency characteristic of an unequalized magnetic tape, using a constant-current input to the head. (Courtesy, Minnesota Mining and Manufacturing Company)

of the recording amplifier. The circuit shown may be used for either ¼-inch tape or magnetic film recorders.

17.48 What effect does the bias current have on the output level?—The relationship between the bias current and the relative output level for a typical magnetic film recorder is shown in Fig. 17-48. The curves shown were

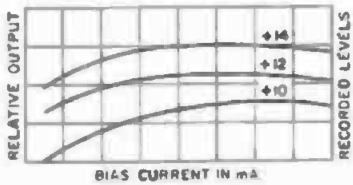


Fig. 17-48. Bias current versus relative output for a frequency of 1000 Hz.

plotted by recording 1000 Hz at a given recording level for various values of recording current over a range of 20 mA. The tape was then played back and the variation in the output level measured.

17.49 What effect does bias current have on the frequency response?—The bias current has considerable influence

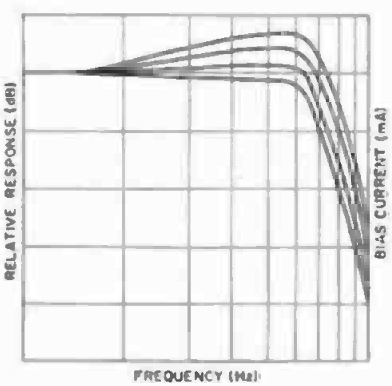
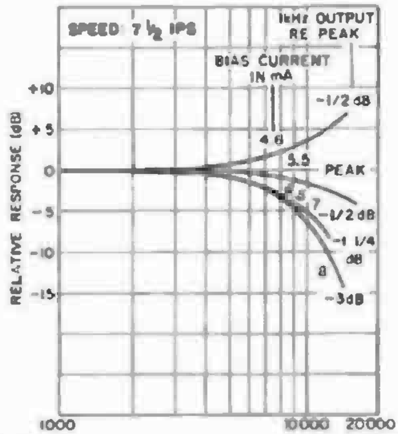


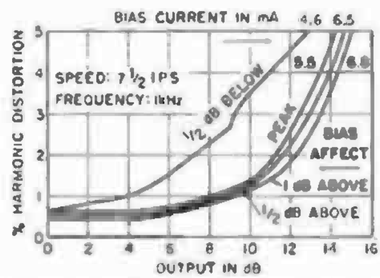
Fig. 17-49A. Bias current versus high-frequency loss for a 16-mm magnetic film recorder.

on the frequency response and distortion, and must be taken into consideration when selecting the bias-current value for a given distortion and signal-to-noise ratio. High values of bias current will partially erase the high frequencies as they are recorded; thus, the recorder will show a loss of high frequencies when played back. It is always a compromise between the best frequency response and bias current for the lowest distortion and a maximum signal-to-noise ratio.

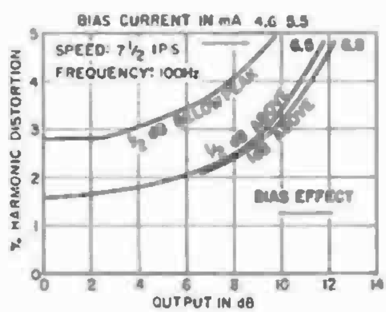
Typical bias current values versus frequency response for a given set of conditions are shown in Fig. 17-49A.



(a) Frequency response at 7.5 inches per second for a 0.5 dB change in bias current.



(b) Relationship between recorder output and harmonic distortion as the recorder input signal level is changed, at 1000 Hz.



(c) Relationship between recorder output and harmonic distortion as the recorder input signal level is changed, at 100 Hz.

Fig. 17-49B. The effect of increasing and decreasing the high-frequency bias current above the value required for peak output at 1000 Hz.

At a speed of 36-feet per minute or less, bias erasure will be more noticeable because of the lower linear speed and the shorter wave lengths. A definite change in the overall frequency characteristics will be noted when the bias current is increased above a given amount. After a given value of bias current has been selected, a frequency measurement should be made.

LeBel, Radocy, and Kramer conducted a series of tests to determine the effect on the frequency response, of changing the bias current for a one-half dB change in output signal level, above and below that obtained at the optimum bias current setting. These tests were made using a one-quarter inch tape recorder running at a speed of 7.5 inches per second using an input signal of 1000 Hz. The results of these tests are shown in Fig. 17-49B. It will be noted at part (a) in Fig. 17-49B, changing the bias current to a value which will reduce the output signal one-half dB below that at the optimum bias value, and one-half dB above the optimum bias value, reduces the high-frequency response at 12,000 Hz 7 dB, and at 15,000 Hz 11 dB, when compared to the optimum bias value. At 15 inches per second these effects virtually disappear. This clearly indicates the importance of adjusting and maintaining the bias current at its correct value.

The effect of the change in bias current, on the harmonic distortion is shown at parts (b) and (c) in Fig. 17-49B, using three-percent harmonic distortion at 1000 Hz as a reference.

The optimum bias current may be determined by one of two methods. By applying a signal of 1000 Hz to the input of the recorder and adjusting the bias current for a maximum signal output level disregarding the harmonic distortion, or by adjusting the bias current for a given value of harmonic distortion at 400 Hz. (See Question 17.52.)

When adjusting a recorder for optimum bias current, the input signal to the recorder is set to a fixed level sufficiently low, that recorder amplifier is not overloaded, as this would introduce nonlinear distortion. Oscillator should be one of low harmonic distortion.

17.50 What effect does harmonic distortion in a bias oscillator have on the recorded signal?—High values of intermodulation are produced by beating of the oscillator harmonics and the harmonics of the program material. Also, a nonsymmetrical waveform in the bias-oscillator current will cause magnetization of the recording head.

17.51 What percentage of harmonic distortion can be tolerated in a bias-current oscillator?—The harmonic distortion should not be over 0.5 percent. To achieve this low distortion, the oscillator circuit should be of push-pull design. Such oscillators may be designed to have as low as 0.10 percent harmonic distortion. Excessive harmonic distortion in a bias oscillator, if used for erasing, will decrease the signal-to-noise ratio of the tape.

17.52 What is the relationship between the bias current and harmonic distortion?—The harmonic distortion will be at a minimum for a given bias value. The proper bias current is found by making a family of bias curves as shown in Fig. 17-52A. This is done by recording a 400-Hz signal for various values of bias current at several different recording levels. The recorded 400 Hz is played back and the harmonic distortion measured for each value of bias current.

Fig. 17-52B shows how the bias value may affect the waveform of the 400-Hz signal if incorrect. It will be noted that

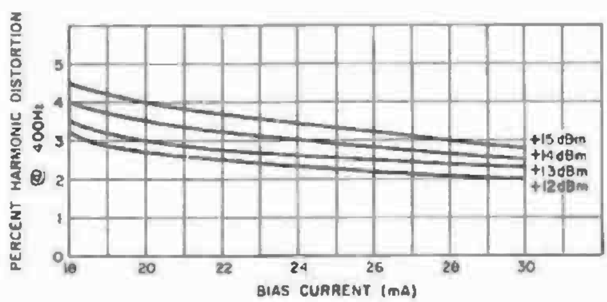
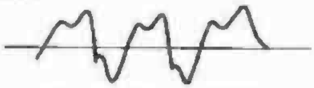
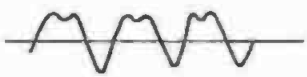


Fig. 17-52A. Bias current versus harmonic distortion for a tape velocity of 36 feet per minute.

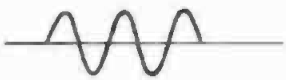
for no bias and half bias the distortion is quite high. When the correct bias current is applied, the distortion is at a minimum. The final value of bias current should be that which results in the lowest distortion, greatest signal-to-noise ratio, and the best frequency response.



(a) *No bias current.*



(b) *One-half bias current.*



(c) *Correct bias current.*

Fig. 17-52B. Waveforms for incorrect and correct bias current.

17.53 At what level are frequency-response measurements made on a magnetic recorder? — Frequency response measurements are always made 10 to 20 dB below 100-percent modulation of the recorder. This applies to both magnetic tape and film. One hundred-percent modulation is defined as the operating level for a given recorder and value of distortion. As an example: the 100-percent level is established for a given THD, or in instances where the VU meter is set to a fixed level (as so many ¼-inch machines are), the level of the input signal is adjusted to 100-percent at a reference frequency of 400 Hz. The level is then lowered 10 to 20 dB, and the frequency run made. If this precaution is not observed, serious overloading of the reproducing amplifier may occur because of the low-frequency post-equalization used in the playback circuits, and the high-frequency equalization used in the recording circuits. This is particularly true in the case of a recorder running at 3¾ and 1⅞ ips, as the high frequency pre-equalization may be 15 to 25 dB above the reference level of 400 Hz, as shown in Fig. 17-172A. It is assumed that before the measurement is made, the azimuth has been adjusted, the bias set to the proper value, and the heads cleaned and degaussed. If difficulty is encountered with the bias voltage appearing in the output of the playback circuit, it may be eliminated by the use of a low-pass filter, described in Question 7.107, or by connecting a 0.50- to 1.0-μF capacitor across the playback output circuit. When making distortion or signal-to-noise ratio measurements, the filter may be left in the circuit, but in the case of the capacitor, it may be necessary to remove it. The use of the capacitor or filter is limited to playback output circuits of 600 ohms or less. (See Question 23.73.)

When making playback measurements using a standard test tape, a reference signal is generally given which precludes the overloading of the playback amplifier. However, it is well to remember the playback equalization increases the frequency response at 50 Hz, 15 to 20 dB. Therefore, if a reference level is used that is not below the maximum equalization of the playback circuits, the playback amplifier may be overloaded.

Certain manufacturers of ¼-inch alignment tape record the tapes in such a manner that they may be played back at the 100-percent or normal-operating level of the recorder without overloading the system. Before using an alignment tape, the manufacturer's data sheet should be consulted. (See Question 17.105.)

17.54 What is the average current used for erasure?—The bias current used for erasure will depend on the head, and varies with different design. However, the erasure current is always quite high. As an example, the bias current in the Ampex Model AG-300 described in Question 17.219 requires approximately 60 mA, at a frequency of 100,000 Hz.

17.55 How is the bias current supplied to a three- or four-channel stereophonic magnetic recorder? — By means of circuits similar to those shown in Figs. 17-55A and B. The bias current to each head is supplied from a separate winding on the bias-oscillator output coupling transformer. A bias-current trap or low-pass filter is connected between the output of each recording amplifier and the point where the bias oscillator connects to recording head.

17.56 Is it possible for the erase current to induce noise in the tape when erasing?—Yes. If the erase or bias oscillator has noise or high-harmonic distortion components, noise will be induced

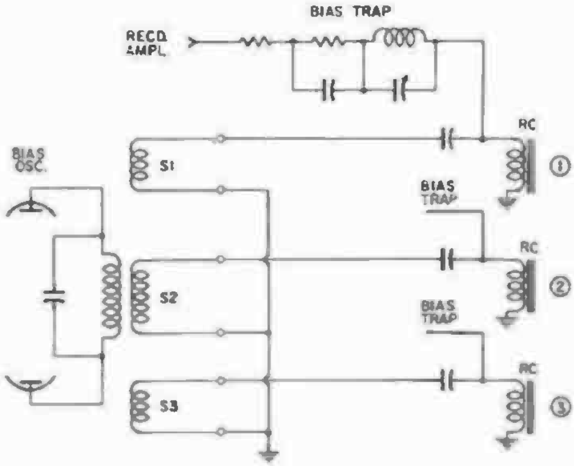


Fig. 17-55A. A bias oscillator circuit for a three-channel stereophonic magnetic film recorder.

in the tape as it is being erased. Noise will also be induced if a strong magnetic field is permitted to encompass the erasure head. This is also true for the playback and recording heads.

17.57 How is the high-frequency bias current prevented from entering the recording circuits?—In magnetic film recording by the use of a low-pass or band-suppression filter as shown in Fig. 17-57. The filter is connected between the output of the recording amplifier and the recording head. The frequency characteristics of the filter are such that it has no effect on the operational characteristics of the recording head or amplifier. The band-suppression filter may consist of two tuned circuits, one resonated to the fundamental frequency of the bias current (F_1) and the other to the second harmonic (F_2) of the bias-current frequency. The design of such filters is discussed in Section 7. For ¼-inch tape recorders, a trap circuit similar to that of Fig. 17-55A may be used.

17.58 How is the voltage of a bias-current oscillator measured?—With a high-frequency vacuum-tube voltmeter.

17.59 How is the bias current measured in a recording head?—By connecting a 10-ohm noninductive resistor in series with the recording head and reading the voltage drop across the resistor with a vacuum-tube voltmeter. The current is then calculated by Ohm's law:

$$I = \frac{E}{R}.$$

17.60 How is the audio-frequency current in a recording head measured?—By cutting off the bias-current oscilla-

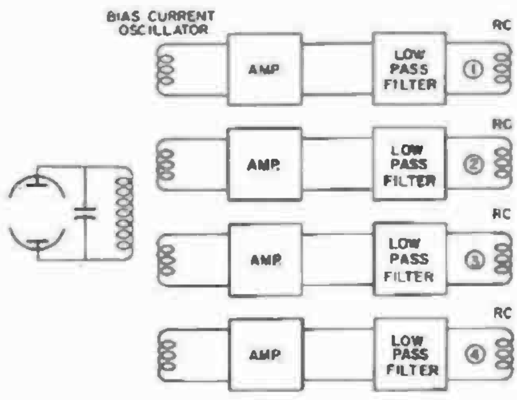


Fig. 17-55B. A basic oscillator circuit for a four-track stereophonic magnetic film recorder.

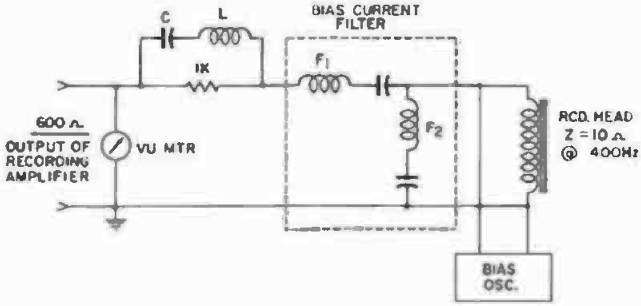


Fig. 17-57. Recording circuit with bias filter to prevent the bias current from affecting the VU meter and other sections of the recording circuits.

tor and then measuring current at 1000 Hz as described in Question 17.59.

17.61 How is the erasure current measured?—By connecting a 10-ohm resistor in series with the head and measuring the voltage drop across the resistor with a vacuum-tube voltmeter. (See Question 17.59.)

17.62 How much signal current is required in the average recording head?—For a recording head of 10-ohms impedance, about 2.5 mA. Again, the current will vary with the design of the head.

17.63 What is a B-H curve?—A characteristic of a magnetic material. The subject of B-H curves for magnetic tape and film is discussed in Question 17.142.

17.64 How are previously recorded signals erased from a magnetic tape?—By the use of a high-frequency current passing through an erase head over which the tape passes before arriving at the recording head. As a rule, the same high-frequency bias current used for recording is also used for erasing. The tape in passing over the erase head is demagnetized or neutralized before being recorded on again.

17.65 Is it necessary to degausse magnetic tape or film with high-frequency current?—No, it may be erased by means of a bulk eraser or degausser using the regular house current. A typical bulk eraser is pictured in Figs. 17-65A and B.

Erasures made using a bulk eraser generally result in a 4 to 6 dB greater signal-to-noise ratio than those erased on a recorder using the high-frequency bias oscillator alone. Magnetic-recording equipment manufactured for motion picture use, as a rule, does not include an erasure head because of the danger of accidental erasure. Also, it permits the high-frequency oscillator to be of smaller design. The interior view of the

Fig. 17-65A. A magnetic tape or film bulk eraser manufactured by the Hollen Corp.

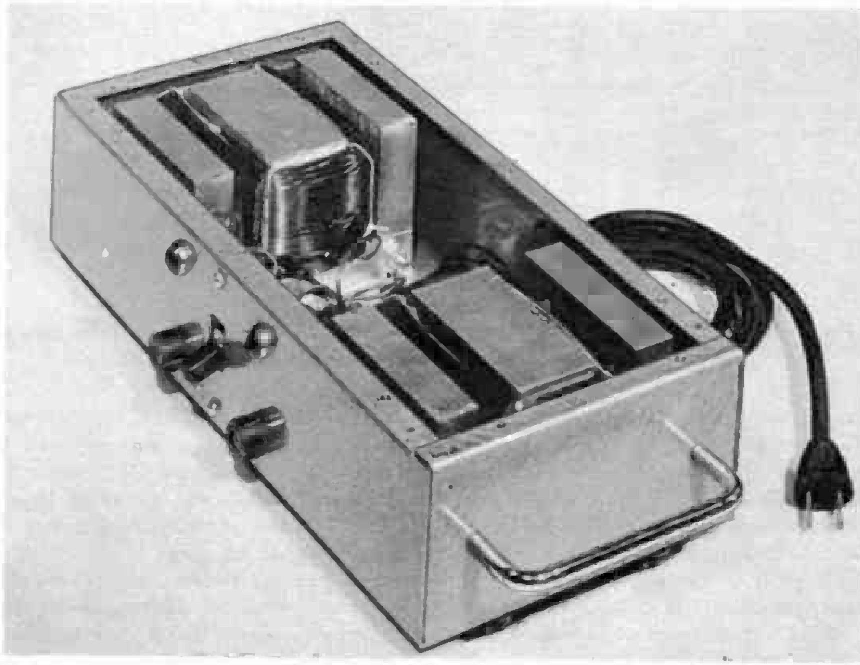


Fig. 17-65B. Interior view of bulk eraser showing erasing coils laminations.

eraser shown in Fig. 17-65A is given in Fig. 17-65B. The schematic diagram of a single-coil bulk eraser appears in Fig. 17-65C.

17.66 Describe the techniques used with bulk erasing.—The purpose of a bulk eraser or degausser is to remove all traces of previously recorded signals from the magnetic emulsion and leave it in a completely demagnetized state. This is quite important from the standpoint of minimizing both noise and distortion.

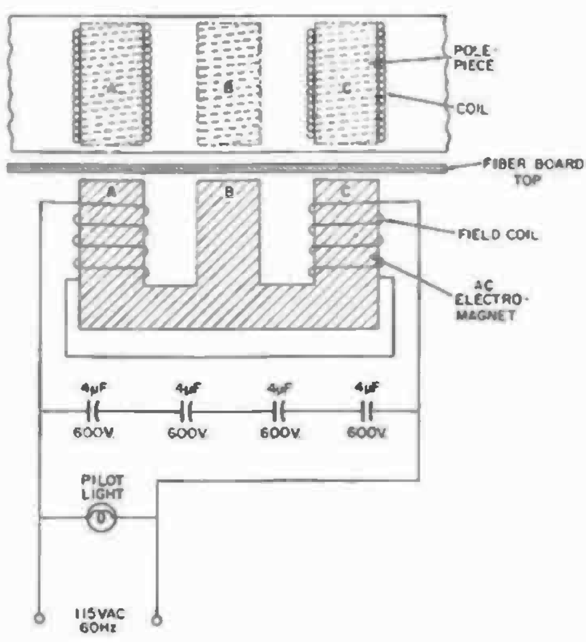


Fig. 17-65C. Circuit diagram of a single-coil magnetic film or tape bulk eraser (degausser).

Experience indicates that the direction of the erasing field with respect to the tape is of considerable importance in securing an effective erasure.

To obtain a satisfactory erasure, the tape must be rotated in the field of the degausser and not slid into the field, as only certain segments would thus be exposed to the field of the degausser.

Erasure is accomplished by saturating the tape in a strong magnetic field which orients the previously recorded signals. Upon removal of the magnetic field, the tape magnetization may change but will not assume any orderly pattern associated with previously recorded signals. To fulfill this latter requirement, the tape must be removed from the saturating field while the field is cycling and gradually diminishing in intensity. If the reduction in amplitude does not exceed approximately 10 percent during one cycle, the tape will be completely demagnetized.

As a rule, a spindle is provided on the top of the eraser for rotating the tape reel over the top of the erasing field. The roll of tape should not be rotated at a speed of over two inches per second. Higher speeds will result in recording the ac field of the degausser in the tape and defeat the purpose. Also, a jerky motion when rotating the tape will result in noise bursts which will be reproduced as a swishing sound behind the signal, thereby increasing the background noise. Such noises have a once-around characteristic and are often mistaken for trouble in the recorder.

The degaussed roll of tape should be slowly removed from the field of the degausser to a point well beyond the influence of the field, then slowly returned for a rotation in the opposite direction in the field of the degausser.

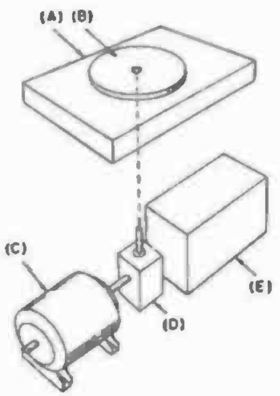


Fig. 17-66A. Elementary diagram of an automatic degausser for magnetic tape and film.

Fig. 17-66B. Studio type automatic magnetic tape and film degausser, manufactured by Radio Corporation of America.

Fig. 17-66C. Hand bulk eraser manufactured by the Amplifier Corporation of America.

The tape must be rotated in both directions of the degausser field. The degausser should never be energized or shut off with the tape anywhere in the degausser field, as the heavy surge of current through the degausser coils will induce a pattern in the tape which will be very difficult to remove. The best method is to employ a variable auto transformer, as described in Question 8.8, in the power line to the degausser and slowly bring the line voltage from zero to maximum and down again to zero while the roll of magnetic tape is being rotated in the field of the degausser.

A sketch for an automatic degausser is given in Fig. 17-66A. It consists of a bulk eraser A mounted on a stand with a nonmagnetic turntable B mounted above the degausser coils and driven by a motor C through a gear reduction unit D. A second motor (not shown) drives a Variac E which slowly increases the line voltage supplied to the bulk eraser from zero to maximum. When the maximum voltage is reached, the motor reverses and reduces the voltage to zero again. At this point in the cycle the two motors are automatically shut off. The turntable rotating the magnetic tape makes several complete revolutions at the rate of approximately two revolutions per minute while the Variac is being cycled. The above degausser unit is typical of those used in the motion picture industry for degaussing magnetic film and tape. Such a device assures an even erasure with a maximum signal-to-noise ratio and requires no attention.

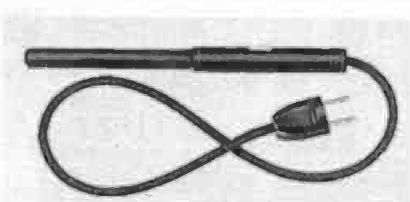


Fig. 17-66D. Pencil eraser manufactured by Cinema Engineering Co.

An automatic degausser, manufactured by Radio Corporation of America and designed for degaussing magnetic tape, film and video tape is pictured in Fig. 17-66B. Normally the turntable of this device is to the right front. The tape or film to be degaussed is placed on the table and moved into the coil, which automatically starts the erasure cycle. Both audio and video signals may be erased down to the noise level of the medium in about 18 seconds. The coil opening is of sufficient size to take 6 rolls of ¼-inch tape, 3 rolls of ½-inch tape, 2 rolls of 16-mm magnetic film, 1 roll of 35-mm film, or 1 roll of 2-inch video tape. The degaussing coil requires 220 Vac at 12 amperes. Power-factor correction capacitors are included in the cabinet.

A small hand degausser, manufactured by the Amplifier Corp. of America, is shown in Fig. 17-66C. This unit is placed over the roll of tape to be degaussed and rotated not more than 2 ips; the tape is then turned over and degaussed a second time. A pencil de-

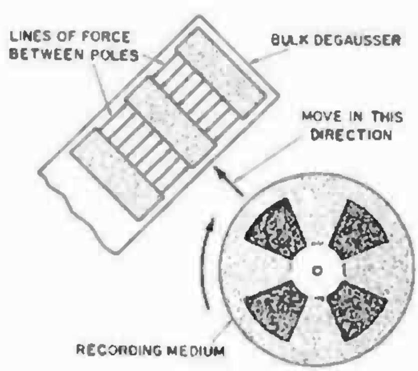


Fig. 17-66E. The roll of tape or film is slid into the field of the degausser as shown. If the lines of force in the degausser are not known, remove the cover and determine their direction. The tape must be rotated at a steady rate of not more than 1 to 2 inches per second.

gausser, manufactured by Cinema Engineering Co., for degaussing small portions of sound track and splices is shown in Fig. 17-66D.

In using the bulk eraser of Fig. 17-65A, if the device does not have a spindle for rotating the tape the cover should be removed and the direction of lines-of-force marked on the top of the degausser to permit the media to be rotated at right angles to the magnetic field (Fig. 17-66E). Experience indicates that regardless of the type of degausser used, the magnetic media must be degaussed first in one direction, turned over, and degaussed again in the other direction.

17.67 What is a constant-current recording characteristic? — A method used for measuring the recording characteristics of magnetic recording tape. The recording current is held constant for each frequency of interest by reading the voltage drop across a 10-ohm noninductive resistor connected in series with the recording head. The current is then calculated by Ohm's law:

$$I = \frac{E}{R}.$$

The recorded frequencies are played back through a flat-amplifier system and the amplitude of each frequency measured and plotted as shown in Fig. 17-46.

A typical measuring circuit is shown in Fig. 17-67.

17.68 How is a magnetic recording head constructed?—The outline for a typical magnetic recording head is shown in Fig. 17-68A. The core A consists of a laminated iron ring wound with several hundred turns of very small wire B. The ends of the core are separated by nonmagnetic shims C_1 and C_2 forming minute gaps D.

The magnetic tape or film E is pulled over the upper poles of the core at the

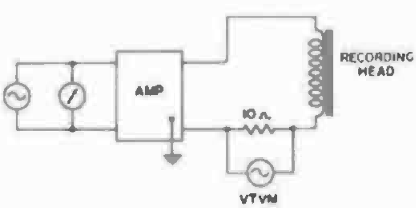


Fig. 17-67. Circuit for measuring the recording characteristic of magnetic tape.

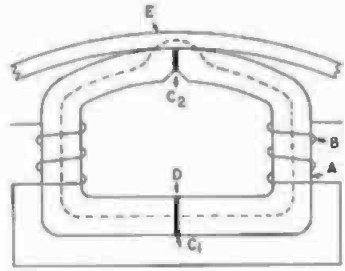


Fig. 17-68A. Construction of a magnetic recording head.

gap. Audio-frequency current flowing through the coils causes a varying flux to be generated in the gap between the pole pieces of the core which is similar in characteristic to the applied audio-frequency currents.

This changing flux causes a field to be introduced into the tape, aligning the molecules of the magnetic emulsion into patterns similar to the impressed audio-frequency currents.

The head must be thoroughly shielded from the effects of stray magnetic fields. An interior view of the construction of a typical magnetic head is pictured in Fig. 17-68B. In some instances, the record and reproduce heads are identical and may be interchanged. (See Question 17.147.)

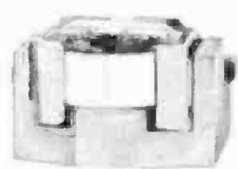


Fig. 17-68B. Interior view of a typical magnetic recording head.

17.69 How is a magnetic reproducing head constructed?—A magnetic reproducing head is constructed in a manner similar to a recording head. In fact, many semiprofessional recorders use one head for both recording and reproduction. When recording, a magnetic flux, varying in intensity, is generated in the recording-head gap and induced into the magnetic tape. When reproducing, the tape passes over the gap in the head and the magnetic flux in the tape, and in passing causes a voltage to be generated in the head windings. These minute voltages are then amplified and reproduced in the usual manner.

The output voltage of a reproducer head may be approximated by the equation:

$$E = 10^{-8} fwn$$

where,

E is in microvolts,
f is the frequency in Hz,
w is the width of the sound track in mils,
n is the number of turns on the head coil.

The foregoing equation may only be used for frequencies recorded on the linear portion of the recording characteristic.

17.70 What is a record-reproduce head?—A magnetic head designed for both recording and reproduction.

17.71 Why do some magnetic recorders employ three magnetic heads?—To perform three separate functions: erasure, recording, and playback. The playback head is used to monitor the signal as it is recorded or for normal playback.

17.72 When two magnetic heads are used, what are they?—They may be an erase and a recording head, or they may be recording and playback. Magnetic recorders used for motion picture recording generally employ only the recording and playback heads. Erasure is performed on a bulk eraser. (See Question 17.66.)

17.73 What is a ferrite core?—A ceramiclike substance molded under high pressure and composed of iron, nickel, zinc, manganese, and copper. The compound is molded into the desired shape and fired similar to any other ceramic.

Ferrite cores have extremely low eddy-current losses. Also, they are quite hard compared to the conventional metal core used for magnetic heads, which accounts for the reason they have not been generally accepted. Because of the hardness, it is difficult to secure sharply defined edges in the gap when the height is of small proportions. Ferrite cores are capable of recording extremely high frequencies, but they are not as efficient at the lower frequencies when compared to conventional metals. Where wear is a factor, ferrite may be used to good advantage.

17.74 What type metals are used in magnetic head construction?—Mumetal,

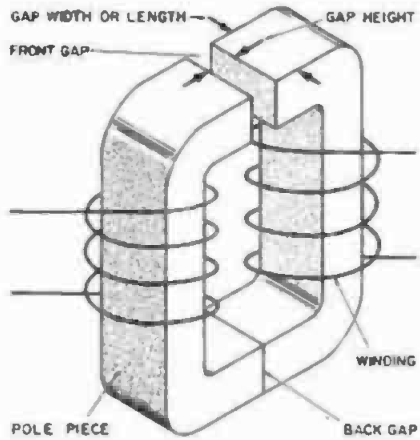


Fig. 17-76. Magnetic recording and reproducing head nomenclature.

Hy-mu 80, NC-88, and Permalloy. To achieve full benefit from these metals, the head laminations are generally less than 0.004 inch in thickness and annealed after working. The laminations are cemented together, using a very thin binder.

17.75 What is the gap height in a recording head?—Generally, on the order of 0.00025 inch or less. The maximum height for wide-range recording cannot exceed 0.0005 inch.

17.76 What is the nomenclature for magnetic recording and reproducing heads?—The nomenclature is that as given in Fig. 17-76. In some instances the gap length is referred to as the width.

17.77 What is the relationship of gap height to frequency response?—When the height of the gap in a reproducer head equals the recorded wavelength, no output signal will be generated because the edges of the gap are on points of equal magnetic potential. The same holds true for integral multiples of the wavelength. The relationship of frequency to gap height may be expressed:

$$F = \frac{V}{2G}$$

where,

G is the gap height in inches,
V is the linear velocity of the tape in inches per second.

In practice, the maximum frequency that may be reproduced for a given gap height is somewhat less than that obtained mathematically, because the gap does not have sharp edges.

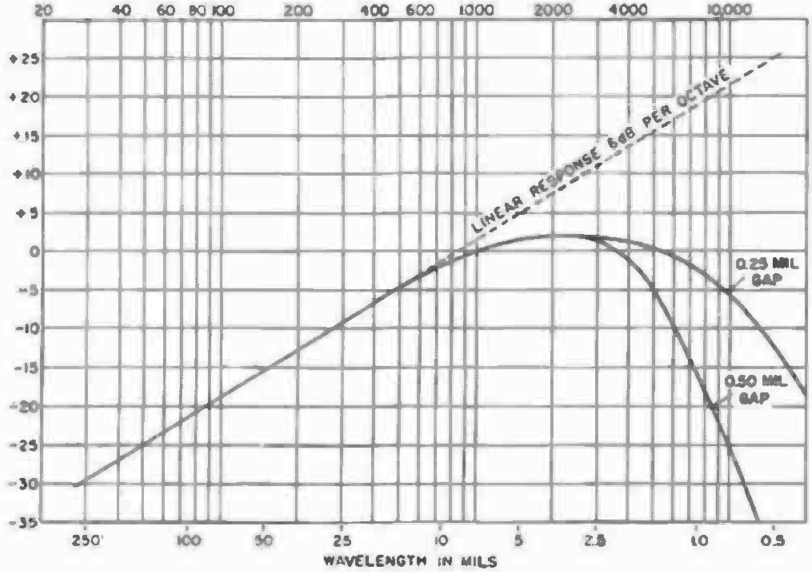


Fig. 17-77. Gap height versus the voltage response for a playback head of negligible losses. Tape recorded constant-flux density at 7.5 inches per second.

Fig. 17-77 illustrates how the frequency response at a given speed is affected by the gap height.

17.78 What is the advantage of a second gap in a magnetic head?—It divides the head into two separate halves making it insensitive to surrounding magnetic fields. This type structure becomes hum-bucking.

17.79 What determines the resolving power of a magnetic recording and reproducing system?—The shortest wavelength or highest frequency that may be recorded or reproduced.

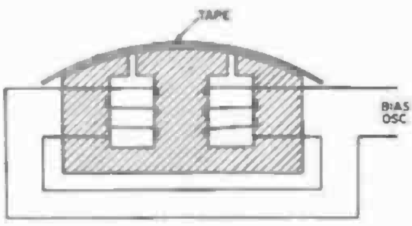


Fig. 17-80A. A two-gap window, or closed-core erasure head.

17.80 How is an erasure head constructed?—Very similarly to a recording head, except that one or more gaps may be used. The gap length is generally longer than the width of the recorded sound track to assure a complete erasure. Also, the gap height is several times the height of the recording gap.

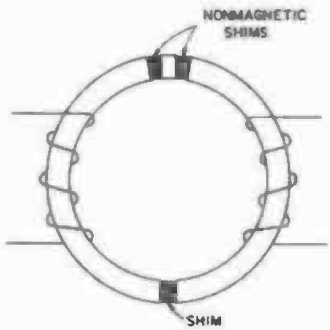


Fig. 17-80B. A three-gap ring-type erasure head.

When two gaps are used (Fig. 17-80A), a closed-core-type construction is used. For a three-gap head (Fig. 17-80B) the *ring-type* construction is used with one gap at the lower portion of the magnetic structure.

17.81 What is wrap around?—The amount of curvature the magnetic tape or film makes in its passage over the pole pieces of the magnetic heads. This term is also applied to sprockets and idler rollers.

17.82 What is the value of erasure current used in the average recorder?—For an erasure head with an impedance of 500 ohms, from 150 to 300 milliamperes.

17.83 What is a dc erasure head?—An erase head to which direct current is applied rather than alternating current. This method of erasure is not gen-

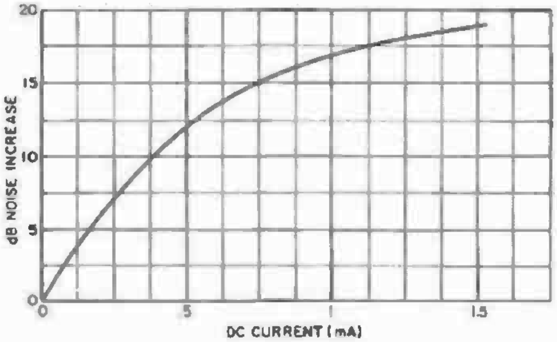


Fig. 17-83. The effect of dc current versus noise for a dc erasure system.

erally used as it introduces a high noise level in the tape with considerable harmonic distortion. Fig. 17-83 shows how the noise increases with an increase of direct current.

Dc erasure is often used in inexpensive tape recorders, particularly those using transistors. In this instance, current from the batteries is applied to the erasure head. This eliminates the necessity for a heavy bias oscillator. Erasure of this nature increases the distortion and is similar to the permanent-magnet-type erasure described in Question 17.85.

17.84 How is an erasure coil connected to the bias-current oscillator?—As shown in Fig. 17-84. The output of the oscillator feeds the erasure head in series with a variable capacitor and a 10-ohm resistor. After the oscillator has been set to its correct frequency, the head circuit is resonated for a maximum current through the head by adjusting the variable capacitor C. The bias current is read by the meter across the 10-ohm resistor R. In commercial recorders, the current is sometimes read by the VU meter.

17.85 Can permanent magnets be used for erasure?—Yes, but like the dc method described in Question 17.83, it induces a considerable amount of noise and distortion into the tape. Permanent magnets are sometimes used with very small inexpensive portable magnetic recorders to eliminate the current drain of a heavy high-frequency oscillator.

The method of placing the permanent magnets along the tape travel for erasure purposes is shown in Fig. 17-85. By the use of permanent magnets, erasure is easily obtained but permanent magnets leave the tape magnetized in one direction. A single-pole magnet will leave the tape fully magnetized to saturation, resulting in a very high noise level with a high degree of even-order harmonic distortion. To minimize this effect, more than one magnet is used

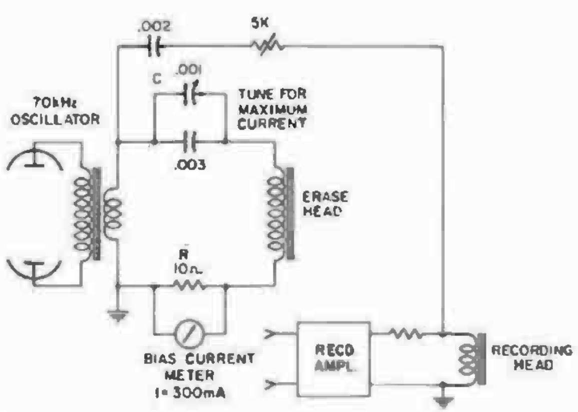


Fig. 17-84. Erasure and recording heads connected to the output of a high-frequency bias current oscillator.

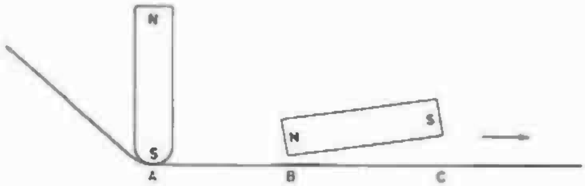


Fig. 17-85. An erasure system using permanent magnets. (Courtesy, Minnesota Mining and Manufacturing Co.)

and so placed that the tape is left in a nearly demagnetized condition.

To secure a completely satisfactory erasure, a large number of magnets of opposite polarity, gradually decreasing in strength, would be equivalent to an ac erasure, but impractical. A simple permanent-magnet erasure system is shown in Fig. 17-85. Successively opposite field maxima are experienced by the tape at points A, B, and C. At A the tape contacts the magnet and is saturated, which obliterates any previous recordings on the tape. The function of the fields B and C is to leave the tape in such a condition that it is essentially demagnetized. To accomplish this in a satisfactory manner, the fields must be of the correct strength and properly spaced for the linear speed of the tape. Practical spacings for 7.5 inches per second are: Point A bears on the tape; at B the spacing is approximately 0.003 inch; and at C approximately 0.028 inch.

17.86 Show the schematic diagram for a high-frequency push-pull bias current oscillator.—A typical push-pull high-frequency bias current oscillator is shown in Fig. 17-86. The coil in the grid circuit is tuned to the resonant frequency by the capacitor C. The coil is of high Q design.

The circuit shown is designed for an output current of 35 mA and is used for supplying the recording bias only. If the oscillator is to be used for both recording and erasure service, a tube with a greater output should be used.

17.87 What is the height of the gap in the average erase head?—The average height is from 2 to 7 mils, with a length slightly greater than the width of the tape. As a rule, magnetic film recorders do not employ erase heads; however, when they do, the erase-head gap is made slightly longer than the width of the sound track, which is 200 mils.

17.88 Why are magnetic film recorders for motion picture recording designed without erase heads?—For two reasons, (1) to prevent accidental erasure of sound tracks, and (2) to eliminate the need for a heavy current oscillator. Erasure is accomplished by the use of a bulk eraser, discussed in Question 17.65. Erasure heads are sometimes included in magnetic film recorders when they are used for looping (see Question 17.223) and are automatically energized and turned off. Erasure heads are also employed with special magnetic pickup recorders as discussed in Question 17.238.

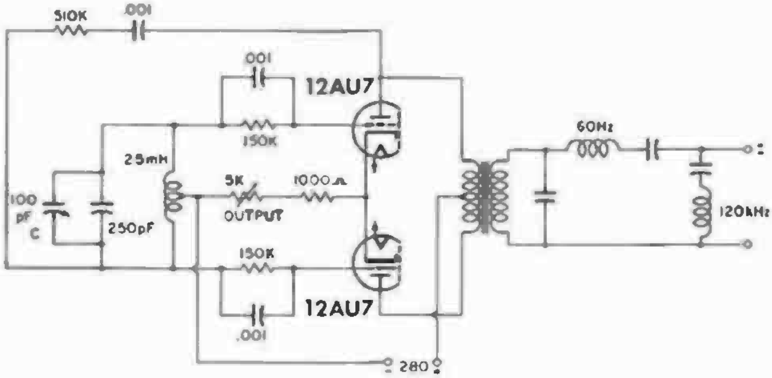


Fig. 17-86. A low distortion 60-kHz recording-bias oscillator circuit.

17.89 What causes a recording head to become magnetized?—Working around the head with magnetized tools; removing tubes when the machine is in the record position; disconnecting the head in the record position; testing the head with a volt-ohmmeter; surges created by the motor system, if operated too quickly after throwing to the record position from playback; and an unsymmetrical waveform in the bias oscillator. A magnetized recording head will decrease the signal-to-noise ratio 6 to 10 dB. Also, it will gradually erase the high frequencies during playback.

17.90 How are magnetic heads degaussed?—By the use of a hand-type degausser as shown in Fig. 17-90A. The most convenient type degausser is that made from a 250-watt soldering gun. The normal soldering tip is removed and in its place is connected a coil of heavy wire wound large enough to slip over the head or other parts to be degaussed.

Three coils are generally required. One consists of 12 turns of number 6 enameled wire about 3½ inches in diameter wound in a single layer with the turns closely taped together. Two other coils of number 8 and 14 wire each containing 12 to 15 turns, respectively, are constructed in a similar manner. The large coil is used for demagnetizing the impedance drum of a recorder or projector and the smaller coils are used to degauss sprockets and recording and reproducing heads. Erase heads do not need degaussing as they are continuously being demagnetized.

A degaussing tool is shown in Fig. 17-90B. The poles are coated on the ends to prevent marring the poles of the recording head.

17.91 How is a degaussing tool used?—By passing the coil of the degaussing tool slowly over the device to be demagnetized and then slowly removing it. The degaussing tool or coil must be energized before slipping it over the part to be demagnetized and must not be shut off until removed a distance of several inches away from the demagnetized part.

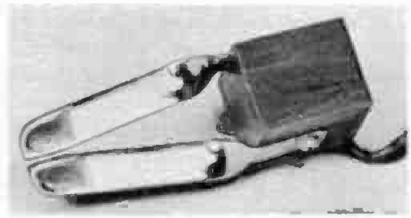


Fig. 17-90B. A degaussing tool for ¼-inch tape recorders.

If the above precautions are not observed and the degaussing coil is energized or de-energized near the part, it may magnetize the part rather than demagnetize it. *The amplifier system should be OFF to prevent a heavy signal from being applied to the amplifier input and possibly damaging the internal components.*

17.92 How many times can magnetic tape and film be degaussed and reused?—The number of times a tape or magnetic film can be used depends entirely on how the tape has been handled or cared for. If the base has not been stretched or warped or the magnetic oxide abrased, it can be reused many

Fig. 17-90A. A degaussing tool made from a 250-watt soldering gun.

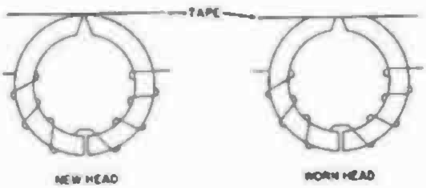


Fig. 17-93A. Top view of a new and a worn head with a flat. (Courtesy, Audio Devices, Inc.)

times. To enable the tape or film to be used over and over requires that the recording machines be in perfect condition relative to their transport systems and head alignment. If the magnetic oxide is damaged in any manner, dropouts and changes in level may be encountered. Tape or film having high-level recording should be erased as soon as possible; if not, it may be rather difficult to erase them down to their original noise level. (See Question 17.66.)

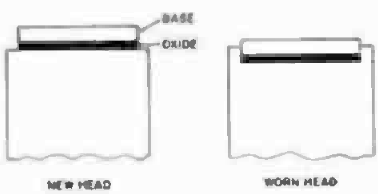


Fig. 17-93B. End view of a new and a worn head with a slot. (Courtesy, Audio Devices, Inc.)

17.93 Show the effects of wear on a ¼-inch tape head.—The abrasiveness of the tape emulsion wears a flat in the pole pieces as shown in Fig. 17-93A, or a groove as shown in Fig. 17-93B. If the tape is slightly oversize, it will climb the sides of the groove, reducing the contact and affecting the frequency response with possible dropouts, as

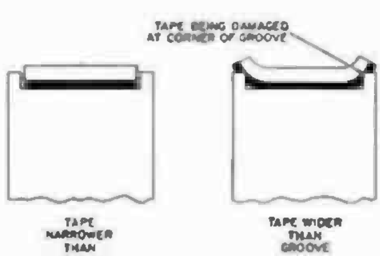


Fig. 17-93C. The effect of tape width on worn heads. (Courtesy, Audio Devices, Inc.)

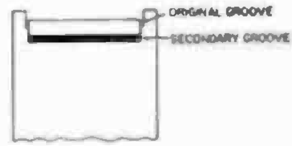


Fig. 17-93D. A worn head with two slots. (Courtesy, Audio Devices, Inc.)

shown in Fig. 17-93C. On the other hand, if the tape is undersize, it will slide against the side of the groove wall and kink, inducing flutter and variations in both recording and playback, with possible dropouts.

In some instances, two grooves are worn in a head as shown in Fig. 17-93D.

Some manufacturers of ¼-inch tape recorders use heads with a slight rise in the center, to prevent slotting the head (Fig. 17-93E). It is assumed the head will be replaced when it has worn even with the sides of the pole pieces. The use of an extended pole piece requires that the tape guide system must be in continuous good alignment. One effect of head wear is that the inductance of the head falls off with the wear of the pole pieces.

17.94 What may be done to salvage a worn recording or playback head?—It may be honed down by using an Arkansas stone, slightly oiled. However, if the honing is done with too great a pressure, the head may be completely ruined. Polishing tapes may also be used to hone the head if the tape is of the same base thickness as the tape or film that is normally used for recording. If stoning is not practical and polishing tapes are not available, a loop of the same tape that is normally used for recording may be used, with polishing compound applied to the surface of the tape. This loop is then run until the desired amount of honing is accomplished. The tape or film that is used to make the loop must be free from warping or

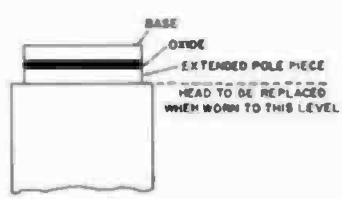


Fig. 17-93E. Extended pole-piece head. Head is replaced when worn down to the level of the main pole piece.

twists. A satisfactory polishing compound is Carborundum Polishing Compound AA 600-V5-OS, an Aluminum Oxide with a grit of 600. (See Question 17.125.)

17.95 How important is tape contact with the head?—If a ring-type head is used, it is essential that the tape be in intimate and unvarying contact with the surfaces of the head pole pieces. The principal loss because of poor contact takes place at the high frequencies, because the field of the pole pieces falls off more rapidly as the tape is separated from the pole-piece surface. It has been demonstrated that when the tape is separated from the pole pieces as little as 0.00075 inch, the response at 5000 Hz will drop 30 dB compared to the level at 1000 Hz. For a separation of 0.0015 inch the loss at 2000 Hz is 28 dB. Thus, it may be seen that the contact the tape makes with the head is all-important. The foregoing data is for playback heads, but there is good reason to believe the same effects hold good for recording heads.

One of the reasons for poor head contact is stepping of the head. A stepped head is one in which one edge or pole piece has raised above the other pole piece due to mechanical slippage. The result is a step. Heads that have become stepped will show a very poor high-frequency response. Stepping of heads will generally occur only once. The step may be removed by honing.

17.96 How is the head-wear pattern of a magnetic head checked?—By coating the surfaces of the head pole pieces with layout blue (Prussian blue). The tape is then run over the head at its normal speed long enough for it to wipe off the blue. The area of contact may be clearly seen after removing the tape. The head angle is adjusted for a smooth even contact by securing an even wipe-off of the layout blue. If good contact

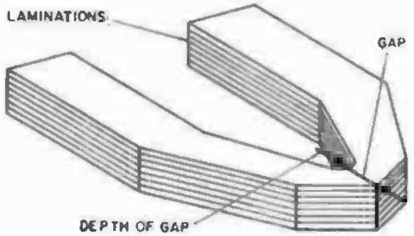


Fig. 17-69A. Magnetic head using tapered pole pieces, before wear.

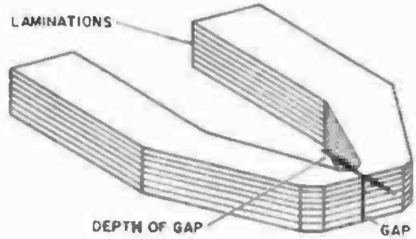


Fig. 17-96B. Magnetic head using tapered pole pieces, after considerable wear.

cannot be achieved by normal adjustment, honing of the head may be done as described in Question 17.94.

If the head construction is of the tapered pole-piece type (Fig. 17-96A), as the wear progresses the pole pieces are worn down as shown in Fig. 17-96B. For a head in this condition there is only one solution—replacement. Head polishing may be carried out as described in Question 17.125. However, it must be done carefully, as it is quite easy to ruin a perfectly satisfactory head by excessive polishing.

17.97 What is contact noise?—Noise generated within the tape by irregular contact with the recording or playback head.

17.98 How may a magnetic recording head be tested for contact?—By recording a 3000-Hz constant-amplitude signal, then playing it back while measuring the output with a VU meter and listening to the reproduction.

Assuming that the tape is of good quality and the machine has a low percentage of flutter, variations in output level may be due to the following causes:

a. Tape bouncing on playback head as it passes over the pole pieces,
b. Tape does not make good contact on recording head but does so on the playback head,
c. Splices in the tape causing bouncing,
d. Scratches or abrasions in tape emulsion,
e. Dirt on emulsion or on pole pieces,
f. Head worn,
g. Drop-outs caused by poor head contact.

High-quality tape when played back on a good reproducer will show a variation in level of approximately plus or minus 0.25 dB and 1 dB from roll to roll.

If the contact pressure on the playback head is too great, high-frequency losses will occur. This is also true if the contact pressure is not great enough. When the contact pressure of the playback head can be adjusted, it is advisable to adjust the pressure until maximum output is obtained and then increase the pressure until the playback level drops 1 dB. This is done to insure that the correct amount of pressure will be present at the surface of the tape.

17.99 Is it possible to have large variations in magnetic tape due to manufacturing defects?—Yes, tape of questionable manufacture will show large variations in output level and dropouts. Also, the noise and distortion can be high, varying from one section to another. High-quality tape will show about plus-minus 0.25 dB and 1 dB from roll to roll.

17.100 How does the manufacturer test magnetic tape for consistency of output level?—By recording a reference signal on the tape as it passes through the dry-box after being coated. The signal is recorded on an automatic-level recorder, and if it varies more than the manufacturing tolerances, a warning signal is sounded.

17.101 What solutions are recommended for cleaning magnetic heads, tape, and film?—Several commercial solutions are available for cleaning magnetic heads, tape, film, and the metal parts around the transport system. Among them are Freon TF, which is nontoxic, noninflammable, and nonexplosive. Also available is a Xylene solution in 0.1-percent Aerosil; however, this solution should not be used for cleaning the capstan or the pinch wheel. Tape guides, capstan, and the pinch wheel should be cleaned with denatured alcohol. *Under no circumstances should carbon tetrachloride be used* because of its toxic nature. In some areas its use is forbidden by law. Cleaning solutions are applied to the head with a cotton swab and wiped off with a lint-free cloth.

17.102 What effect does magnetization of the heads have on recording and reproduction?—The noise and distortion are increased.

17.103 How are the heads of a magnetic recorder shielded from the influence of external magnetic fields?—By placing the heads in shields of Mumetal alternated with copper shields. The head is completely enclosed in the shield, except for a slot to permit the tape to make contact with the head pole pieces. The cases are nested as described in Question 8.51.

17.104 What is an azimuth adjustment?—An angular adjustment in a horizontal plane moving in a clockwise direction. (See Fig. 17-104.) In a magnetic recorder or reproducer, it is the angle the gap in the head deviates from a right angle with respect to the direction of travel of the tape. This adjustment is similar to that of adjusting the recording slit in an optical film recorder.

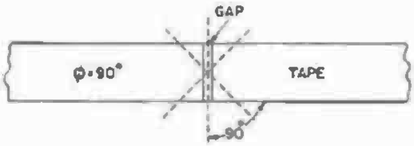


Fig. 17-104. Azimuth adjustment of magnetic head. For correct azimuth the angle of the gap to the direction of tape travel equals 90 degrees.

17.105 How are test tapes recorded?—Test tapes are generally recorded to comply with the NAB Standard (April, 1965) using a 400-Hz reference level tone recorded 10 to 15 dB below a THD of 3 percent, with the tape running at a linear speed of 7¾ ips, and the bias adjusted to its optimum value. These are the levels established for the NAB Primary Reference Tape.

NAB Standard test tapes are designated by the number 65, the linear speed, and recorded full track. Therefore, if they are used with a reproduce head that is less in width than full track, a low-frequency rise may be expected. Instructions are given with the tape for this correction. The frequencies standardized for test tapes are given in Fig. 17-105A. In addition to the frequencies shown, an azimuth adjustment frequency of 5, 10, or 15 kHz is given depending on the linear speed of the tape. The recorded reference level in dB below 100-percent modulation is given at the top of each column, for different linear tape speeds. Each frequency has a duration of 12 seconds, the reference tone 20 seconds, and the azimuth adjustment 60 seconds. Also included is an additional program reference tone of 1000 Hz. In recording test

15 ips 0 dB	7.5 ips −10 dB	3¾ ips −15 dB	1⅞ ips −15 dB
15 kHz	15 kHz	—	—
12 kHz	12 kHz	—	—
10 kHz	10 kHz	10 kHz	—
7.5 kHz	7.5 kHz	7.5 kHz	—
5 kHz	5 kHz	5 kHz	5 kHz
2.5 kHz	2 kHz	2.5 kHz	2.5 kHz
1 kHz	1 kHz	1 kHz	1 kHz
750 Hz	750 Hz	750 Hz	750 Hz
500 Hz	500 Hz	500 Hz	500 Hz
250 Hz	250 Hz	250 Hz	250 Hz
100 Hz	100 Hz	100 Hz	100 Hz
75 Hz	75 Hz	75 Hz	75 Hz
50 Hz	50 Hz	50 Hz	50 Hz
30 Hz	30 Hz	30 Hz	30 Hz
Azimuth 15 kHz	15 kHz	10 kHz	5 kHz

Fig. 17-105A. Standard NAB test frequencies used for recording test tapes. The zero level frequency on tapes for linear speeds less than 15 ips is lower by 10 to 15 dB to prevent overloading the playback system, because of equalization.

tapes for speeds other than 7¾ ips, they are recorded to supply the same ideal head-flux at the same wavelength as the Primary Reference Tape, when measured on an ideal reproducing system.

Because of the multiplicity of equipment in the field, test tapes cannot be completely changed to conform to the NAB Standard of April, 1965. Therefore, most tape manufacturers can supply tapes having characteristics used before the adoption of the present standards. To obtain the accuracy required in the manufacture of test tapes, each tape is an original. It will be observed in Fig. 17-105B that the frequencies for magnetic cartridge test tapes are slightly different than those for the conventional recorder-reproducer.

Manufacturers of standard test tapes hold the recorded frequency level to plus or minus 0.25 dB or less. A voice announcement precedes each frequency for identification.

17.106 What are the effects of improper azimuth adjustment?—The tables in Figs. 17-106A, B, and C show the effects of improper azimuth alignment for one-quarter, one-half, and full-track ¼-inch tape recorders. Azimuth can also be affected by deformation of the tape which is caused by skewing as it travels over the heads and tape-guiding mechanism. Skewing is caused by tape wound too tightly or unevenly on the reel, cramped against the flange of the reel, and moisture absorbed by the base material causing ripples along the edge of the tape. The amount of this bending loss at ½-mil wavelength varies somewhat with the tape, but it is generally on the order of 0.5 dB. It will be observed that the adjustment of the azimuth is quite critical for a full-track head, more so than for one of less than full-track. On a full-track machine, reproducing 15,000 Hz with only two minutes of azimuth error, the loss will range from 0.5 dB at 15 ips, to over 5 dB at 3¾ ips. By study of the tables, losses may be extracted for different wavelengths and sound-track configurations.

400-Hz Calib. Level. 10 dB below Standard Ref.		
400-Hz Standard Ref. Level. plus-minus 0.25 dB		
15 kHz	2.5 kHz	150 kHz
12 kHz	1.0 kHz	75 kHz
10 kHz	600 kHz	50 kHz
8 kHz	300 kHz	30 kHz
5 kHz		

Fig. 17-105B. Standard NAB test frequencies used for recording magnetic tape cartridge test tapes.

1-Mil Wavelength		½-Mil Wavelength		¼-Mil Wavelength	
Loss in dB	*Azimuth Error in Minutes*	*Loss in dB*	*Azimuth Error in Minutes*	*Loss in dB*	*Azimuth Error in Minutes*
0.5 dB	14.86	0.5 dB	7.43	0.5 dB	3.71
1.0 dB	20.90	1.0 dB	10.45	1.0 dB	5.22
2.0 dB	29.21	2.0 dB	14.60	2.0 dB	7.30
3.0 dB		3.0 dB	17.67	30. dB	8.83
4.0 dB		4.0 dB	20.16	4.0 dB	10.08
5.0 dB		5.0 dB	22.16	5.0 dB	11.13
6.0 dB		6.0 dB	24.08	6.0 dB	12.04
7.0 dB		7.0 dB	25.68	7.0 dB	12.84
8.0 dB		8.0 dB	27.09	8.0 dB	13.54
9.0 dB		9.0 dB	28.36	9.0 dB	14.18
10.0 dB		10.0 dB	29.50	10.0 dB	14.75

Fig. 17-106A. Loss due to azimuth misalignment for 43-mil quarter-track. (Courtesy, Ampex Corp. Test Tape Laboratory.)

1-Mil Wavelength		½-Mil Wavelength		¼-Mil Wavelength	
Loss in dB	*Azimuth Error in Minutes*	*Loss in dB*	*Azimuth Error in Minutes*	*Loss in dB*	*Azimuth Error in Minutes*
0.5 dB	8.52	0.5 dB	4.26	0.5 dB	2.13
1.0 dB	11.98	1.0 dB	5.99	1.0 dB	2.99
2.0 dB	16.75	2.0 dB	8.37	2.0 dB	4.18
3.0 dB	20.27	3.0 dB	10.13	3.0 dB	5.06
4.0 dB	23.12	4.0 dB	11.56	4.0 dB	5.78
5.0 dB	25.53	5.0 dB	12.76	5.0 dB	6.38
6.0 dB	27.61	6.0 dB	13.80	6.0 dB	6.90
7.0 dB	29.44	7.0 dB	14.72	7.0 dB	7.36
8.0 dB		8.0 dB	15.53	8.0 dB	7.76
9.0 dB		9.0 dB	16.26	9.0 dB	8.13
10.0 dB		10.0 dB	16.91	10.0 dB	8.45

Fig. 17-106B. Loss due to azimuth misalignment for 75-mil two-track. (Courtesy, Ampex Corp. Test Tape Laboratory.)

1-Mil Wavelength		½-Mil Wavelength		¼-Mil Wavelength	
Loss in dB	*Azimuth Error in Minutes*	*Loss in dB*	*Azimuth Error in Minutes*	*Loss in dB*	*Azimuth Error in Minutes*
0.5 dB	2.5	0.5 dB	1.3	0.5 dB	0.64
1.0 dB	3.6	1.0 dB	1.8	1.0 dB	0.89
2.0 dB	5.0	2.0 dB	2.5	2.0 dB	1.25
3.0 dB	6.0	3.0 dB	3.0	3.0 dB	1.52
4.0 dB	6.9	4.0 dB	3.5	4.0 dB	1.73
5.0 dB	7.6	5.0 dB	3.8	5.0 dB	1.91
6.0 dB	8.3	6.0 dB	4.0	6.0 dB	2.07
7.0 dB	8.8	7.0 dB	4.4	7.0 dB	2.20
8.0 dB	9.3	8.0 dB	4.7	8.0 dB	2.33
9.0 dB	9.7	9.0 dB	4.9	9.0 dB	2.43
10.0 dB	10.0	10.0 dB	5.0	10.0 dB	2.53

Fig. 17-106C. Loss due to azimuth misalignment for 250-mil full-track. (Courtesy, Ampex Corp. Test Tape Laboratory.)

The NAB Standard specifies that the azimuth of an NAB-65 test tape must be within plus-minus 1 minute of arc, with respect to the edge of the tape. This means the slitting of the tape, during manufacture, and the tension and winding of the tape on the reel must be carefully controlled to prevent stretching and warping of the edge. These precautions must also be observed by the user.

17.107 What frequencies are used for magnetic film azimuth alignment?—The Society of Motion Picture and Television Engineers (SMPTE) and the USASI (ASA) Standards specify 7000 Hz for 16-mm magnetic film running at a linear speed of 36 fpm; 8000 Hz for 17.5-mm running at 45 fpm, or 35-mm at 90 fpm. However, most studios employ a frequency of 9000 Hz for both 17.5-mm and 35-mm film.

The azimuth of the playback head must first be established before the recording head azimuth can be set. This is accomplished by playing back a standard azimuth film of either 7000 or 9000 Hz and measuring the amplitude of the azimuth signal at the output of the playback amplifier. The playback head is rotated for a maximum signal and locked into place.

A well degaussed magnetic film is then threaded on the machine, and either a 7000- or 9000-Hz signal recorded while observing its amplitude at the playback amplifier. The recording head is then rotated for a maximum signal and locked in place. The recording head will now have the same azimuth as the playback head. Azimuth may be measured by making the sound track visible, as discussed in Question 17.133.

17.108 To what accuracy are azimuth test tapes and film recorded?—For ¼-inch tape, plus-minus 1 degree of arc; for magnetic film tape, 3 degrees of arc, regardless of size. Because of the extreme accuracy required in such tapes and films, each one is an original recording. In the instance of magnetic film, azimuth test films may be obtained from the SMPTE in various lengths. However, azimuth tests are generally included in most multiple test films.

17.109 What is the procedure for lining up the magnetic heads on a tape recorder?—Assume the recorder has a playback head which may be used for monitoring a signal as it is recorded. The lineup tape is played back and the azimuth of the playback head rotated for a maximum output signal. A degaussed tape is then threaded on the recorder and a signal of the same frequency as the lineup frequency is recorded, while monitoring the signal from the playback head.

The recording head is rotated for a maximum signal at the playback output. When the output is maximum, the azimuth of the recording and playback heads are the same. This adjustment should be made at a level of 10 to 20 dB below the maximum recording level.

17.110 How is the azimuth of a combination record and playback head aligned?—By aligning the head in the playback position. As the same head and gap are used for both playback and record, the azimuth adjustment is not too important, unless the recorded material is to be played back on other equipment.

17.111 Is the azimuth adjustment of the erase head important?—No, as long as it is approximately at right angles to the direction of motion of the tape, it will be satisfactory as the gap is quite wide.

17.112 What are the distortion characteristics of magnetic tape?—Several distortion curves made by the Minnesota Mining and Mfg. Co. on their Type 111 magnetic tape are shown in Fig. 17-112. The third harmonic distortion at 400 Hz is shown for several values of bias and recorded level. These four curves and distributed around the optimum value of bias current.

17.113 What is the relationship of linear tape speed to frequency response?—As it was shown in Fig. 17-46, the high-frequency response increases with an increase in linear speed. Nearly all of the factors affecting the high-frequency response of magnetic tape are not frequency dependent, but rather dependent upon the wavelength of the recorded signal.

Doubling the speed of the tape doubles the highest frequency that may be recorded. It is understood that such factors as the recording and reproducing head-gap heights and azimuth adjustment must all be taken into consideration as well as the linear speed of the tape, for best results.

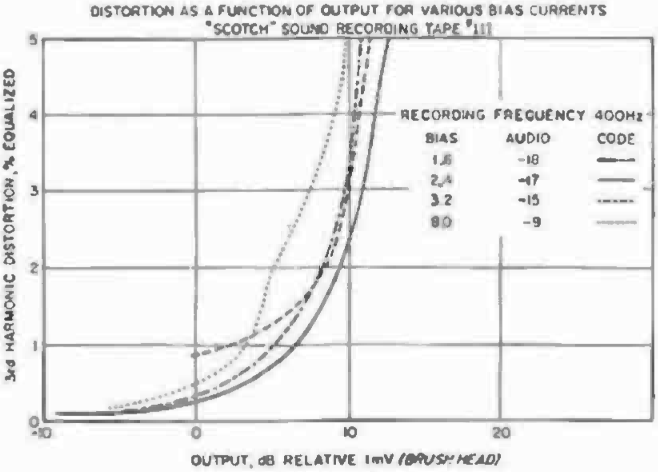


Fig. 17-112. Third-harmonic distortion as a function of output level versus bias current. (Courtesy, Minnesota Mining and Manufacturing Co.)

17.114 What is the lowest linear speed that may be used for magnetic recording?—Speeds down to a fraction of an inch per second are in current use for scientific recording. One commercial recorder uses a linear speed of $^{15}\!/_{16}$ inch per second. The frequency range is from 300 to 2500 Hz with a dynamic range of 35 dB.

17.115 What is the relationship between the width of a magnetic sound track and the output level?—The output level is directly proportional to the width of the sound track. If the original track is 200 mils in width and it is reduced to 100 mils, the output level will drop 50 percent or 6 dB. If the width is reduced to one-quarter of its original width, the output will drop to 25 percent of original value or 12 dB.

17.116 What effect does the particle thickness and physical alignment have on the characteristics of magnetic film or tape?—The thickness of the magnetic emulsion has a very definite effect on the frequency response of the recording medium. If the high-frequency bias current is adjusted for the best frequency response using a medium which employs a thick magnetic coating and then one employing a thin magnetic coating is substituted, a reduction in the high-frequency response will result.

If the high-frequency bias current of a recorder is adjusted for the best frequency response using a thin magnetic coating and a heavy coating is substituted, high-frequency response will remain the same with some improvement in the low-frequency response.

The particles of the gamma ferric oxide are dispersed throughout a resin binder and have a very definite effect on the characteristics, which depend on the shape, size, and orientation of these particles. Frequency response, signal-to-noise ratio and the general sensitivity are all interrelated not just to each other, but to how close to optimum the gamma oxide is handled in manufacture. The particles may be visualized as a group of needles which must be oriented in parallel lines, and their magnetic fields such as to reinforce each other. As these particles measure 1 micron by 0.2 micron (1 micron equals $^{1}\!/_{1000}$ of a millimeter), it is rather a difficult manufacturing operation to get them into alignment.

During manufacture, the particles are put into a ball mill, a massive stainless-steel drum with about two million steel ball bearings, and are tumbled. A binder to act as a suspension is also put into the drum as the particles are tumbled. As the ball bearings tumble, they shear the honeylike suspension, separating the individual particles, and coating them with the binder to prevent them from making contact with one another.

In the early days of tape manufacture, the particles were cubical in shape, requiring that the tape be operated at a linear tape speed of 30 ips for good quality recording. However, cube-type

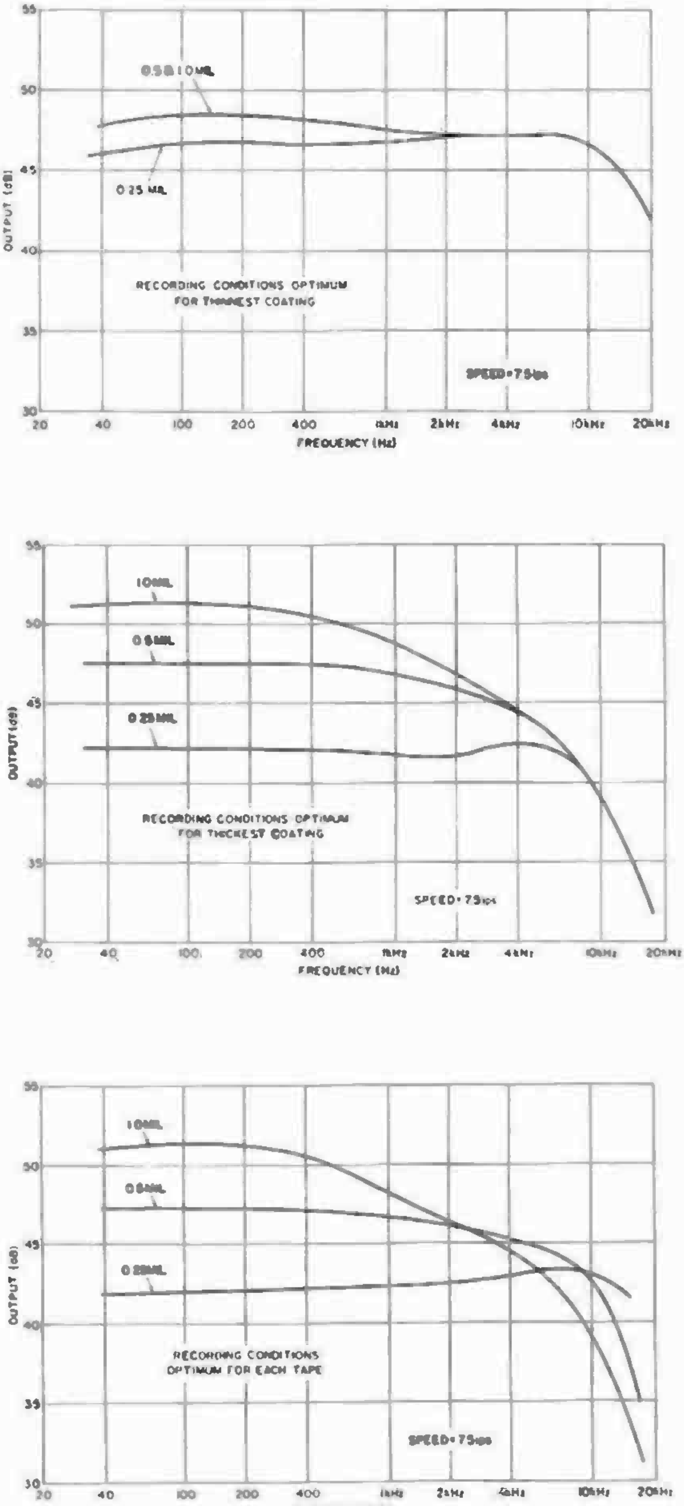


Fig. 17-116. Comparison of frequency response for various magnetic-oxide coatings running at 7½ ips. (Courtesy, Magnetic Products Div., Minnesota Mining and Manufacturing Co.)

particles cause a loss of sensitivity, elongating the hysteresis curve, and increasing the print-through. The milling time (tumbling time) is critical to within 1 percent, with temperature and humidity playing their part, also. When the magnetic oxide is first placed on the base, the particles are in random fashion. While the coating is still soft, the tape is passed through a very strong magnetic field that physically orients the particles in a lengthwise direction, end to end and parallel to each other, and they remain so throughout the life of the tape. Three groups of frequency versus coating thickness for linear speeds of 7.5 ips are given in Fig. 17-116. (See Question 17.120.)

17.117 What causes modulation noise?—Modulation noise can be caused by irregularities in the magnetic coating. It manifests itself by a fuzziness in the reproduction and varies with the percentage of modulation, because the noise is modulated by the signal frequencies. Modulation noise is only apparent when a signal is present. (See Question 17.160.)

17.118 What is the difference between black and red oxide tape?—The black oxide has a higher coercivity than the red oxide, although they both have approximately the same frequency response. The black oxide was originally used with paper base tape, and is now obsolete. (See Question 17.29.) Black oxide should not be confused with the dark green magnetic oxide which is used on high-output tape, described in Question 17.164.

17.119 What causes dropouts in magnetic recording?—Dropouts are caused by discontinuities in the magnetic coating and, if great enough, cause noise pulses which act as a spurious signal. Dropouts are caused by a lack of magnetic coating at the point where a signal is recorded and lead to an explanation based on the more frequent occurrences of small inclusions, called nodules, in the magnetic coating. These nodules can be classified as oxide clumps, acetate particles, embedded filter fibers, etc.

Oxide clumps which protrude from the otherwise flat surface of the tape force the main body of the tape away from the recording and playback gaps. During recording, the presence of a nodule reduces the sharpness and intensity of the recording field. On playback, when the rate of change of the recorded flux is observed, the already reduced steepness of the flux front is observed from a distance which further reduces the rate of change of flux in the reproducing head. This combination results in a decrease in output which is called a dropout.

Two groups of curves showing the effect of dropouts and the attending loss of level are given in Figs. 17-119A and

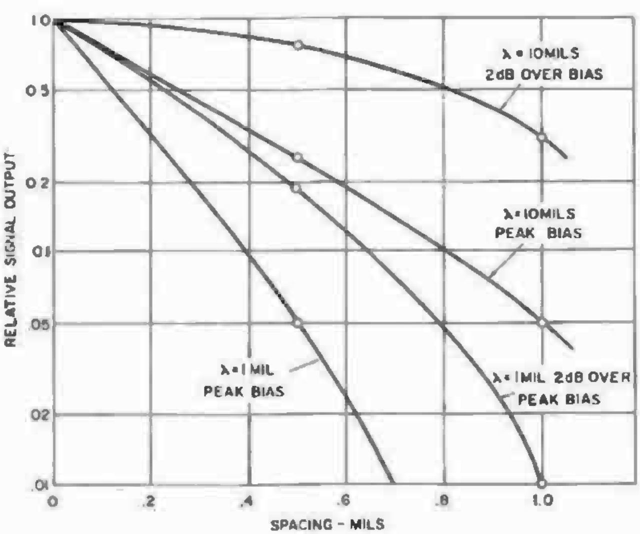


Fig. 17-119A. Curves showing the fall-off of recorded signals versus spacing from recording head. (Courtesy, Minnesota Mining and Manufacturing Co.)

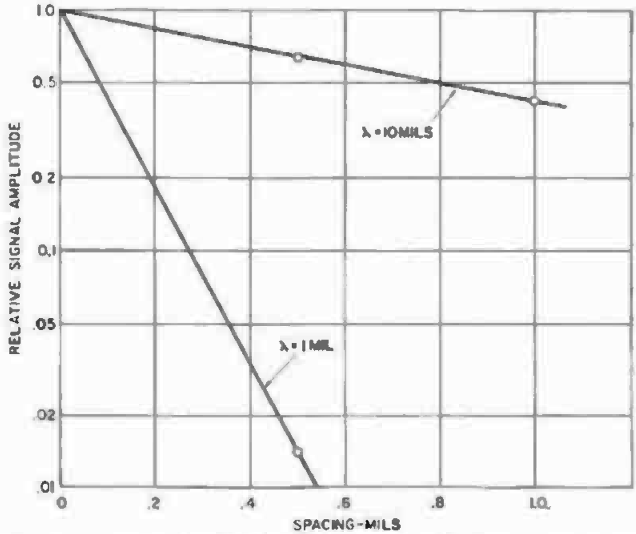


Fig. 17-119B. Curves showing fall-off of reproduced signals versus spacing from reproducer head. (Courtesy, Minnesota Mining and Manufacturing Co.)

B, with the appearance of a typical dropout recorded by means of a graphic level recorder.

It is the general practice in the tape manufacturing industry to record a wavelength of about 37.5 mils on the product at a constant input level. This signal is then played back, filtered, and the output signal at particular critical wavelengths recorded using a graphic recorder. The recorded response for instantaneous dropouts caused by foreign matter in the coating or on the surface is shown graphically in Fig. 17-119C. To evaluate surface smoothness and tape to head contact, a wavelength of 1 mil is used. These two tests used together aid in evaluating the lubrication, slitting of the base material, and the oxide binder characteristics. The smoother the resulting record the more uniform the magnetic sensitivity. The graphic recording in Fig. 17-119D shows typical results using the above test methods. The above data apply equally well to both magnetic film and tape.

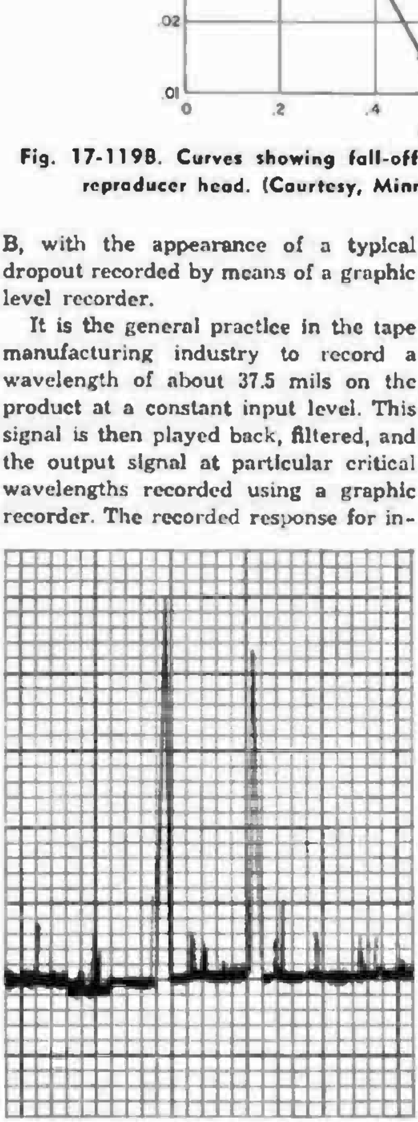


Fig. 17-119C. Graphical appearance of a magnetic tape with dropouts, using a 37.5-mil wavelength. (Courtesy, Eastman Kodak Co.)

17.120 What is meant by orientation of the magnetic emulsion particles? —It is the process of orienting the individual particles of iron oxide so that they face in the same direction, thereby increasing their overall magnetic effect. Orienting of the iron-oxide particles results in lower distortion and an

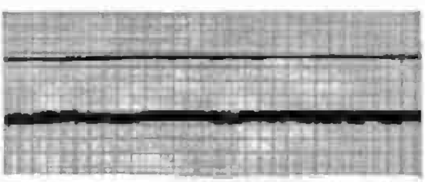
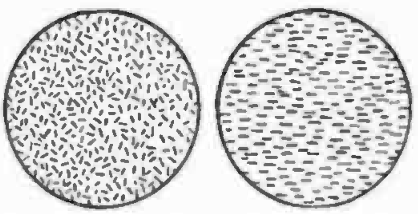


Fig. 17-119D. Upper trace; surface smoothness and tape-to-head contact test, using a 1-mil wavelength. Lower trace; test for foreign matter on tape surface, using a 37.5-mil wavelength. These two tests are for a high-quality coating. (Courtesy, Eastman Kodak Co.)



(a) Nonoriented. (b) Oriented.

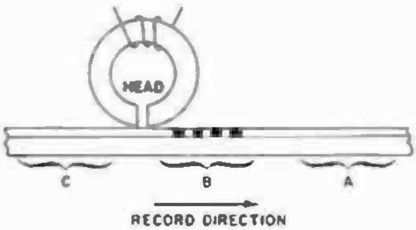
Fig. 17-120. The arrangement of non-oriented oxide particles and oriented oxide particles.

increase of 2 to 3 dB in the output level for a given set of conditions.

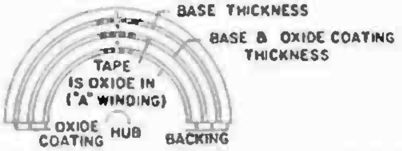
During the manufacture of the magnetic tape, a strong magnetic field is applied while the tape is drying and hardening. The particles line up longitudinally (Fig. 17-120) and have a cooperative magnetic effect which improves the recording characteristics.

17.121 What is print-through?—The unwanted transfer of a signal from one layer of tape to another by magnetic induction. Print-through is caused by overmodulation or recording at excessively high levels approaching saturation.

When tape or magnetic film is wound from reel to reel, each layer of tape is in the magnetic field of its neighbor. Since any magnetic material subjected to a magnetic field tends to become magnetized, each layer of tape will be magnetized to some extent by the adjacent layer. The effect is nonlinear, and is somewhat like recording without bias current. The transfer level decreases about 2 dB for each 1-dB decrease in the recording level. Generally, transfer effect is only noticed for the most highly modulated sections. The transfer effect increases with the time of the storage, the conditions of storage, and with increased temperature. Therefore, tape and magnetic film should be stored in a cool place, and in metal cans if possible. The tape must also be kept away from stray magnetic fields, both ac and dc, as such fields can cause a transfer from a few dB to 30 or 40 dB. Good rules to follow are do not drive the tape into saturation, avoid exposure to magnetic fields, and avoid high temperatures. Several manufacturers can supply tape with a low ratio of print-through, and this type should be used for critical recordings. Magnetic film, because of its heavier base, does not suffer as much from print-through as does the ¼-inch tape; however, the same precautions should be observed relative to magnetic fields. If the head end of the tape or film contains music, for storage it should not be rewound, but left tail out. This will prevent echo effects of print-through.



(a) Direction of tape travel.



(b) Tape stored on take-up reel. This method of storage is recommended for master tapes.

(c) Tape stored on stock reel.

Fig. 17-121. Showing how print-through is generated.

Since the amount of print-through received by a given section depends upon its separation from the section carrying the print signal, the next outer layer will receive more print-through than the next inner layer, in a normal oxide-in winding. This is true because the printing field must reach the top surface of the oxide coating undergoing print-through. Thus, to print the next outer layer the print signal must pass through only the base material. To print the next inner layer, the print signal must pass through the base and an additional thickness of oxide. To illustrate these points, the direction of recording (greatly enlarged) is shown in Fig. 17-121 part (a). Fig. 17-121 part (b) shows a tape stored on the take-up reel, with section C undergoing maximum print-through. As playback section C goes past the playback head, the printing signal is heard as a postprint, or as is commonly called, an echo.

In Fig. 17-121 part (c), the tape is shown rewound and stored on stock reels, with section A receiving the maximum print. Under these conditions, when played back, section A precedes the printing signal past the reproduce head and appears as preprint. The method of storage illustrated in Fig. 17-121 part (b) is recommended for storing master tapes because it results upon playback, of the two print-throughs appearing stronger as post-print and weaker as preprint. Postprint is less troublesome than preprint because it is likely to be masked by the original signal. Preprint may be quite noticeable if it is strong and if preceded by a quiet section of tape.

Print-through is transient. Its level drops quite rapidly upon removal from the printing field and will show a drop of 6 dB or more after removal. By storing tailout and rewinding just before playback, the printed signal intensity is reduced, and generally there is not sufficient time for a new signal to be printed.

17.122 How should magnetic tape be stored?—In a temperature of 60°F to 70°F with a relative humidity of 40 to 60 percent, sealed in a metal container, and stored away from magnetic fields, as even a weak field will cause print-through. Tape stored in a high temperature will experience a life reduction of about 90 percent. Curves showing how temperature affects the transfer of a signal from one layer to another for two different types of magnetic tape are given in Fig. 17-122.

17.123 If a tape of different manufacture is substituted for that normally used, what tests are required?—Although professional magnetic tape and film are held to close tolerances, there is some variation between manufacturers; therefore, if possible, a complete set of measurements of bias, frequency response, distortion and signal-to-noise ratios should be made on sample rolls. It is the practice of several of the major motion picture studios to degauss and measure the bias requirements for each roll of magnetic film just before using. The bias current is measured for *each* roll of film on the recorder it is to be used with. Experience has shown this has its advantages in the long run, as the recording levels must be the same from roll to roll and it is an assurance of the best recording conditions at all times. Generally, a special test circuit is installed as an integral part of the recorder so that the test may be accomplished quickly. Measuring in advance and marking the bias current on each roll is not always satisfactory, as the bias setting could be overlooked by the operator.

17.124 What causes volume compression in magnetic tape or film?—Excessively high recording levels which drive the tape into saturation. However, this may be easily detected because as the tape approaches saturation, the high-frequency distortion is increased, resembling heavy sibilance.

17.125 What is lapping tape?—Lapping tape is used for applying a high polish to the surface of finished products for many devices. It may be obtained with grit so fine that as little as 3 microns (3 thousandths of a millimeter) can be removed. Such tape can be very useful in resurfacing magnetic recording and reproducing heads. Lap-

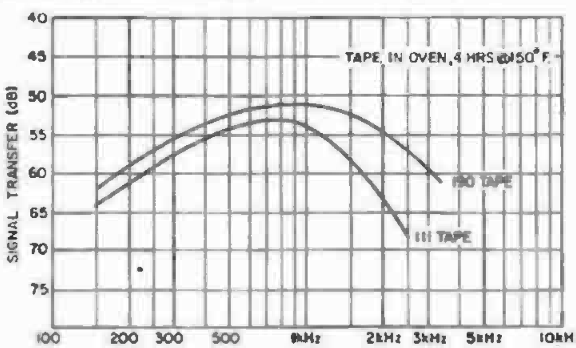


Fig. 17-122. Layer to layer signal transfer. (Courtesy, Minnesota Mining and Manufacturing Co.)

ping tape is available in standard magnetic film and tape widths and is obtainable from most manufacturers of magnetic tape. Before purchasing or using such tapes, the tape manufacturer should be consulted for recommendations as to the grit size. (See Questions 17.93 and 17.96.)

17.126 Explain the use of lapping tape and what precautions must be observed in its use?—If a magnetic head has been reset because of an uneven wear pattern, provided precaution is observed, lapping tape may be used to resurface the pole pieces by splicing a piece of lapping tape about 10 inches in length into a loop of leader stock of about 3 feet in length. After two or three pases of the lapping tape over the head, the loop is lifted and the pole pieces are inspected under a magnifying glass and the wear pattern is noted. Only a very minimum amount of lapping should be used, as it can wear the pole pieces down quite rapidly. The new wear pattern can be made visible by the application of machinist blue to the pole pieces, then a clean leader is run over the head. The wear pattern will become visible by the removal of the machinist blue. (See Question 17.96.)

17.127 What is the relationship between output voltage and frequency from a magnetic reproducer head?—The voltage generated in the playback head coil is proportional to the rate of change of the magnetic flux in the tape linking the coil.

17.128 How are coils in a magnetic head connected?—They are connected *series aiding* or as one continuous coil, or *hum-bucking* as described in Question 8.98.

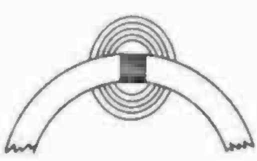


Fig. 17-129. Magnetic field distribution around the pole pieces of a magnetic recording head.

17.129 How do the lines of force appear between the poles of a recording head?—As shown in the exaggerated cross-sectional view in Fig. 17-129. The strength of the field falls off quite rapidly even a small distance from the pole pieces.

17.130 What is fm tape noise?—Frequency modulation of the recorded signal as it passes over the recording head because of the varying frictional forces acting on the tape in its contact with the pole pieces of the head.

The effect of fm noise is similar to that generated by drawing a bow across a violin string; however, the tape being highly damped does not produce a tone, only the noise. Frequency modulation of the signal may also be caused by minute irregularities in the oxide coating of the tape and variations in the speed of the transport system of the recorder. In the latter case, it is referred to as flutter.

There are two ways of identifying frequency modulation noise from that of magnetic noise. Fm noise is sensitive to the frequency of the recorded signal and as a rule, low frequencies are relatively free of this effect, whereas the higher frequencies are quite sensitive. The second method of test is to alter the tape tension, or the tape guides, and to note the effect on the high frequencies. As a rule, increasing the tension increases the noise, while a decreased tension lessens the friction and noise is reduced. The application of an oscillator to the input of the channel, set to mid-range high frequency, will generally be of assistance in identifying and locating the source of trouble.

17.131 What is the effect when a magnetic recording head is overloaded?—The core saturates and produces the same effect as if the gap had been widened. Overloading also increases the harmonic distortion and reduces the signal-to-noise ratio.

17.132 What causes beats or tweets in the reproduction from a magnetic recorder?—The beats are only heard when the tape is reproduced and are caused by the frequency of the bias-current oscillator beating with the applied audio frequencies when the tape was recorded. Beats may be the result of cross talk between the oscillator and recording circuits or poor filtering in the bias-current circuits. If the frequency of the bias oscillator is less than five times that of the highest frequency to be recorded, beat notes may result.

17.133 Can magnetic sound tracks be made visible?—The sound track re-

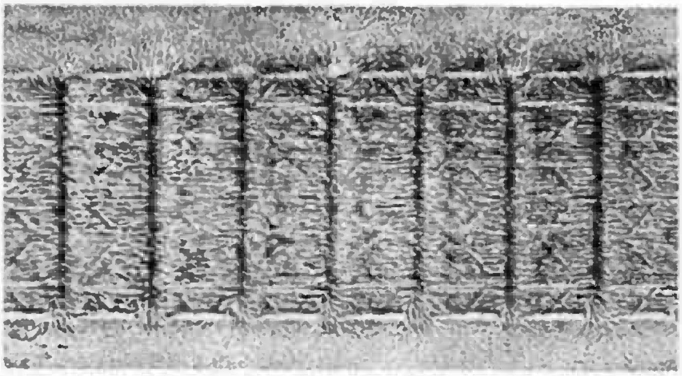


Fig. 17-133A. A magnetic sound track made visible by the use of carbonyl iron in heptane. (Courtesy, Minnesota Mining and Manufacturing Co.)

corded on magnetic tape may be made visible by means of a method similar to that used in the mapping of fields around magnets. The iron particles must, of course, be much smaller than iron fillings. The particles which are used are carbonyl iron with a diameter of about 3 microns, or about 3 thousandths of a millimeter. In addition to being small, the particles must be able to move about so that they can settle in regions where the tape is strongly magnetized. In order to provide the desired mobility, the carbonyl iron may be dispersed in a light oil or in a volatile substance such as heptane. Even ordinary water may be used. The simplest method is merely to pass the recorded tape through a suspension of carbonyl iron in heptane. The heptane quickly

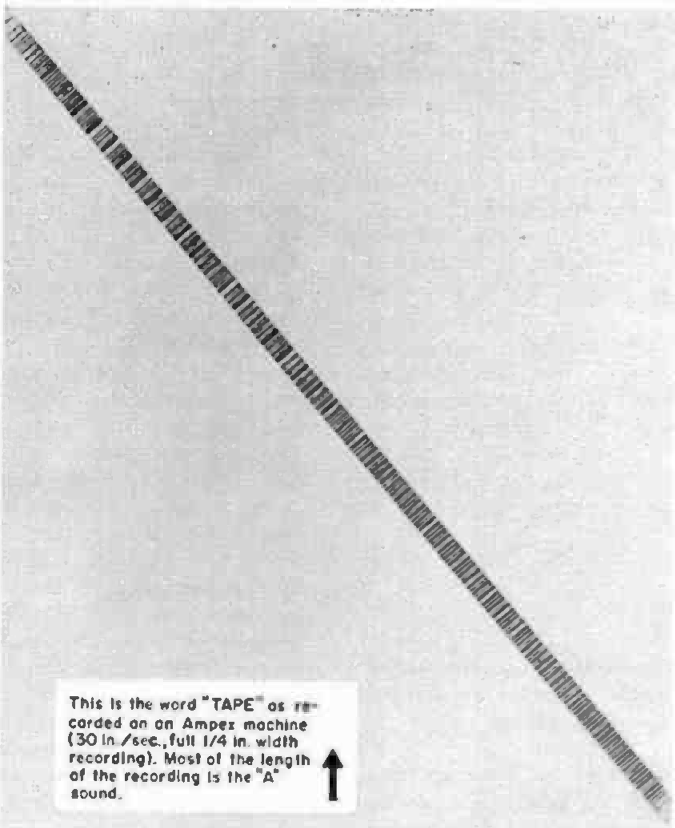


Fig. 17-133B. A magnetic sound track with the word "tape" made visible by the use of carbonyl iron in heptane.

evaporates leaving the carbonyl iron particles settled on the regions which are most strongly magnetized. The photographs in Figs. 17-133A and B illustrate typical sound-track patterns which have been made visible in this manner. The longer wavelengths recorded on the tape are evident to the naked eye. Very short wavelengths, as short as 0.001 inch, may also be observed. However, in the case of the short wavelengths, more satisfactory results are obtained if the carbonyl iron is dispersed in a light oil instead of being suspended in heptane.

The track made visible in the manner described above has a number of uses. One application of the method is to permit an evaluation of the degree of alignment of the magnetic heads. This requires a study of very short wavelengths. One can determine with a microscope whether a recorded track is perpendicular to the direction of tape travel. Relative positioning of the sound tracks in multitrack recording may also be examined. Defects in the gaps of recording heads may be revealed and some idea of fringing effects may be obtained. Occasionally, by using a microscope to examine the visible track, one may find tape defects which contribute to noise. The method cannot be used to reveal weakly recorded signals and for best results a fairly high signal level is necessary.

Two commercial solutions are available, under the names of Visa-Mag and Magna-See. A mechanical device, manufactured by Minnesota Mining and Manufacturing Co., and sold under the trade name of Scotch Magnetic Viewer, permits the *direct viewing* of the sound track. The viewer is about 2 inches in diameter and contains a mobile solution in which are dispersed finely powdered magnetic particles. This solution is sealed between a thin nonmagnetic diaphragm on the bottom of a glass top. When placed over the sound track modulations, the particles suspended in the liquid arrange themselves in accordance with the modulations, thus the track becomes visible. The pattern is removed by rubbing the back of the diaphragm with a circular motion.

17.134 What effect does mechanical vibration have on a magnetic recorder?—If the vibration is great enough the recording head is vibrated, causing mechanical modulation of the recording head laminations. The resulting modulations are recorded as noise, and sound similar to running gears. Head vibration may also take on the sound of rubbing noise, and in most instances appears only during modulation periods. This type of noise or distortion is quite hard to detect and has the effect of distorting the program material in such a manner that it is not easily associated with any particular portion of the transport system. At times the noise may be heard between portions of dialogue; however, as a rule, it is only heard when modulation is present.

Such noises can be detected by sweeping an audio oscillator slowly from 200 to 3000 Hz through the system, while simultaneously monitoring the signal on the playback head. The noise will appear behind the oscillator signal. The frequency of the oscillator signal will many times supply a clue as to which section of the transport system is causing the vibration.

If the noise is below 800 Hz it may be caused by gears, motor unbalance, belts, or a combination of all. If it is of a high-frequency nature, it may be caused by the transport system vibrating the tubes in the recording or playback amplifiers causing microphonics. (See Question 11.58.)

Each recorder will exhibit its own characteristics, depending on the number of motor poles and the method of coupling the motor to the transport system. Also, the tension of the supply and take-up reels must be such that they do not induce vibration.

Microphonic tubes will induce noise in the modulations during recording because of mechanical vibration. A simple test is to tap the tubes in the recording amplifier while recording. Noise due to microphonics will be recorded as a ringing sound. If no noise is heard it may be assumed that the tubes are not contributing to the noise. If microphonic noises are heard when playing back an unmodulated tape, the playback amplifier tubes are microphonic.

17.135 What is the cause of cross talk in a multitrack recorder-reproducer?—Cross talk between recording heads is caused by magnetic coupling and the transformer action between the head-

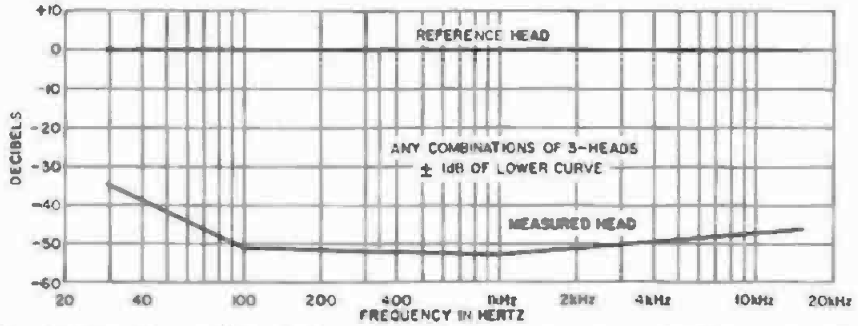


Fig. 17-135. Cross-talk levels between heads 1 and 3 of a 3-track 35-mm magnetic film recorder using a track width of 200 mils. Recording level plus 10 dBm.

coil windings. This may be reduced by separation of the windings and introducing a grounded shield between the windings. In reproducing heads, cross talk is caused by leakage between the windings and the fringing effect at the lower frequencies. (See Question 17.199.)

It can be observed in the curve of Fig. 17-135, that the cross-talk ratio is the greatest at the midrange frequencies and lowest at the lower frequencies. For frequencies above 1000 Hz, only a slight reduction in the ratio is noted.

The measurements shown were made by applying a constant-amplitude signal to head 1, and measuring the cross talk or leakage across head 3, and various combinations of the three heads. The results shown are typical for a three-track 35-mm magnetic film recorder, using a sound track of 200 mils in width.

17.136 Describe a magnetic time-delay and a magnetic reverberation unit. —Magnetic time-delay units are used in auditoriums and theaters to induce a time delay between the original sound and certain areas in the house. Time-delay units are used for eliminating confusion or interference. Basically, the device is a magnetic tape recorder, using a continuous loop of tape, which is first erased then recorded on. Playback heads are placed along the path of the tape loop, and the recorded signal is picked up at various intervals, inducing a time delay between the original sound and the reproduced sound. As an example: assume it is not possible to cover the entire floor of a theater or auditorium having a balcony from a single or group of loudspeakers concentrated at that single point. Loudspeakers placed high over the stage can be adjusted to cover most of the main floor and the balcony, with a minimum of acoustic feedback. However, they will not cover the under area of the balcony. Adding speakers at the sides of the stage only increases the acoustic feedback. This difficulty can be overcome by placing speakers under the balcony. In this

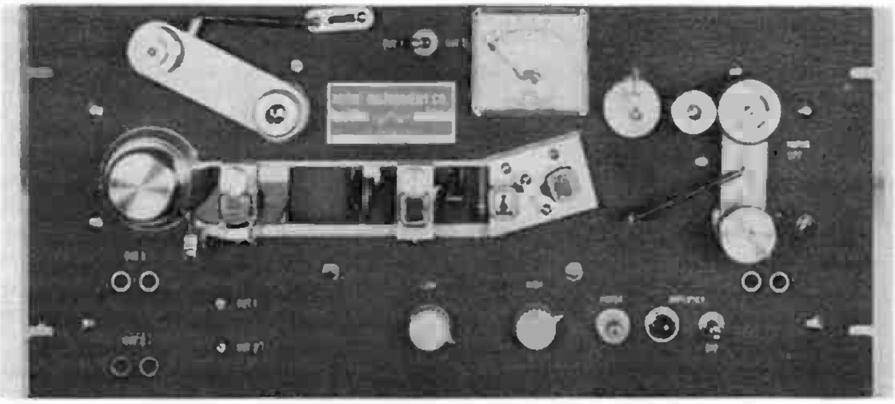


Fig. 17-136A. Model 301 magnetic time-delay unit manufactured by Audio Instrument Co.

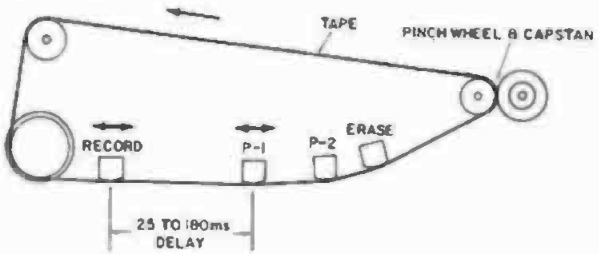


Fig. 17-136B. Transport system and head layout for Audio Instrument Co., magnetic time-delay unit Model 301.

instance, the coverage at the back wall will be satisfactory, but an area of confusion and unintelligibility will exist when the stage speakers and those under the balcony have the same sound power level (SPL). This is true because the sound from the stage speakers is delayed and arrives a considerable time after the sound from under the balcony speakers is heard. This condition may be overcome by delaying the sound from the speakers under the balcony up to 50 milliseconds beyond the time it takes the sound arriving at the front of balcony from the stage speakers.

The Haas effect states: The first souud to be heard takes command of the ear, and sound arriving up to 50 milliseconds later seems to arrive as a part of and from the same directions as the original sound. This restores intelligibility and eliminates confusion.

To achieve the required time delay, a magnetic unit such as Model 301 time delay, manufactured by the Audio Instrument Co. (Fig. 17-136A) may be employed. This device uses a 33-inch loop of hardened surface tape, running at a linear speed of 30 ips (other models may run 20 and 90 ips). The recording and reproducing heads at the left are movable and may be adjusted for a time delay of from 25 to 180 milliseconds, at 60 Hz. Such a tape loop will give about 20 hours of service at 30 ips, or about 8 hours at 90 ips.

Delay devices may be obtained for either monophonic or stereophonic use. Equalizers are included for increasing the midrange high frequencies and reducing the low-frequency response. The transport and head placement is shown in Fig. 17-136B. (See Questions 2.35 to 2.37.)

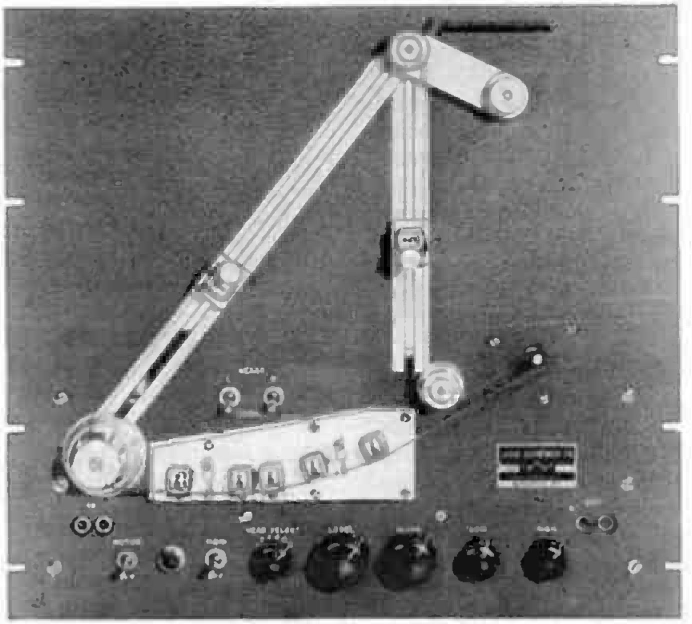


Fig. 17-136C. The Model 42A magnetic reverberation unit manufactured by the Audio Instrument Co.

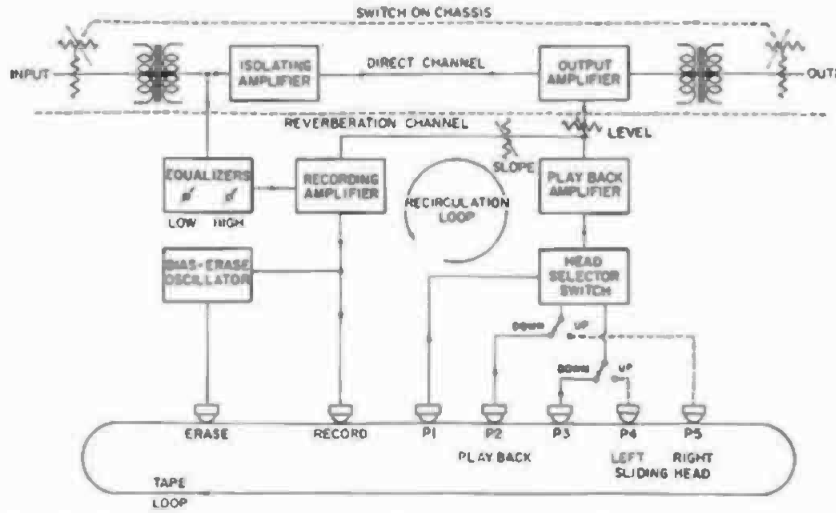


Fig. 17-136D. Block diagram for Audio Instrument Co. Model 42A magnetic reverberation unit.

A magnetic reverberation unit, using a ¼-inch magnetic tape and 7 playback heads that may be used to add synthetic reverberation, manufactured by the Audio Instrument Co., is shown in Fig. 17-136C. The signal to be reverberated is first recorded, then reproduced by one of the several heads at the desired delay time. A portion of the delayed signal is then returned to the recording head and passed around a recirculation loop again, with diminished amplitude. Thus, the signal is caused to drop in amplitude in the same manner as a sound wave diminishes during a multiple reflection from the interior walls of an enclosure. The ratio of reverberated sound to direct sound may be widely altered, corresponding to a shift in the microphone-to-orchestra distance. Included are high- and low-frequency equalizer circuits for correcting and attaining the desired frequency response. An elementary block diagram of the internal connections is given in Fig. 17-136D.

A magnetic reverberation unit of different design is shown in Fig. 17-136E, manufactured by Bauer Electronics Corp., and it utilizes a rotating recording head around a circular drum. Two playback heads, one variable and one fixed, are also a part of the tape loop system. Reverberation effects are accomplished by sampling the incoming program material, processing and recording this sample on a closed-loop magnetic tape with a variable time-delay repeat, amplifying the signal, then mixing it with the original program material. A short loop of magnetic recording tape encircles the tape drum and passes over erase, record, and playback heads. Selection of the tape speeds and the head spacing determines the delay time of a specific audio signal recorded on the tape. The physical sepa-

Fig. 17-136E. Bauer Electronic Corp. Model S-1000 surrounding sound reverberation unit.

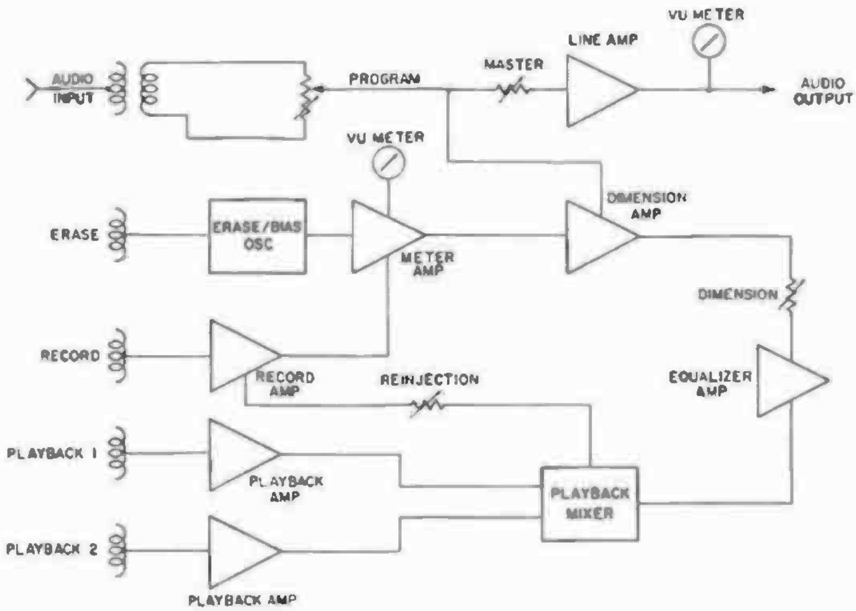


Fig. 17-136F. Block diagram for Bauer Electronics Corp. magnetic tape reverberation unit.

ration on the record and playback heads makes possible the variable dimension effect. A block diagram of the system appears in Fig. 17-136F.

The record head is mounted on an arm that may be rotated about the tape drum to vary the distance between the record and number one channel playback head. In addition, the number two channel playback head, mounted after the first head, is also movable, making a controlled multiple delay possible. Any portion of this controlled signal or other variation made within the electronic portion of the surrounding sound unit can be mixed with the incoming program material to achieve a desired result. The tape is fed from a Cousino Audio Vender which holds a quantity of tape, and is pulled downward and upward by the action of two pinch wheels and a capstan below the tape supply, as used in the conventional tape recorder. Time delays ranging from 0.1 to 1.30 second are possible by rotating the recording head (under hand) around the tape drum. Speeds of 10 and 20 ips are available by throwing a switch.

Referring again to the overall view of Fig. 17-136E, variable equalization is provided using an equalizer similar to that described in Question 6.12. Two VU meters provide monitoring of the record, playback, and program channels. The electronics are solid state, using a cadmium cell for switching to eliminate the possibility of noise. The power supply is mounted at the lower portion of the cabinet.

It is quite common practice to use reverberation units in radio broadcasting to increase the signal strength at the fringe areas of transmission and for enhancing certain types of records and effects in commercials.

17.137 Describe the electron cloud magnetic head.—Several different designs of magnetic heads have been developed to reproduce magnetic sound tracks. Since most of these designs require rather elaborate construction, the simple coil and core construction used in the majority of tape recorders has been retained. In July, 1939, a novel design was suggested by A. M. Skellett (patent 2,165,307) which resembles a cathode-ray tube, in which pole pieces are substituted for the beam-deflecting coils. At the face end of the tube, the beam strikes a pair of target plates. The magnetic flux from the tape in passing over the head deflects the beam and unbalances the current flow to the target plates. The output of this transducer is directly proportional to the flux recorded on the tape, rather than to the rate of change. Therefore, signals at extremely low speed can be reproduced even if the tape is at rest; thus, the low-frequency response is uniform and does not require 6 dB per octave equalization. The principal objection to this form of a transducer is its sensitivity to extraneous magnetic fields, even to the earth. The transducer requires extensive magnetic shielding which makes it rather bulky. Also, magnetic pole pieces must be included within the glass envelope.

17.138 Describe a staggered sound track recorder.—In the design of early model stereophonic recorders and reproducers, because the technique of building in-line magnetic heads had not been mastered, the heads were staggered, as shown in Fig. 17-138, to increase the cross talk-to-signal ratio. Modern recorders and reproducers now use in-line heads and shielding between heads to reduce the cross talk. It is not uncommon for telemetering equipment to use 12 or more heads in a single in-line stack (cluster). (See Questions 17.135 and 147.)

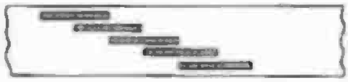


Fig. 17-138. Staggered sound tracks used in early recorders to reduce cross-talk. (Now obsolete).

17.139 What is the standard NAB reference and program level for ¼-inch tape recorder-reproducers? — The NAB Standard (April, 1965) specifies that the reference level shall be 400 Hz and equal to the recorded level on the NAB primary reference tape. The standard recorded program level shall produce the same reference deflection on a Standard Volume Indicator Meter (VU meter USASI (ASA) Standard C16.5-1961) as that produced by a 400-Hz sine-wave signal recorded at the standard NAB reference level. The NAB standard reference level is 2.2 dB below that formerly used. It will be found that some test-tape manufacturers use the older higher level. In this instance, the gain must be reduced 2.2 dB to equal the present NAB level.

The NAB primary reference tape is recorded at 7.5 ips, with the bias current adjusted for a maximum output level on an average good quality tape. The reference level is recorded 8 dB below a level that will produce 3-percent third-harmonic distortion. All NAB test tapes contain this reference level and are within 0.25 dB of the staudard reference level.

The use of an 8-dB level below the 3-percent third harmonic does not imply a failure to meet the 10-dB overload margin, but is rather a practical, convenient method of specification consistent with magnetic recording and reproducing systems.

The motion picture industry uses a somewhat different method in determining the recording level. As most of the sound track produced in a studio is used internally on their own equipment, the recording level is based on the maximum practical signal-to-noise level for a given percent harmonic distortion. As a rule, most studios operate with a maximum distortion of 0.5 to 1 percent THD, for a signal-to-noise ratio of 60 to 65 dB.

If a composite print for theater use is to be released using magnetic sound track, the maximum distortion on the release print may be increased to 2 percent, and in some instances even 3 percent. Increasing the maximum distortion also permits a higher level to be recorded on the release sound track. This higher distortion is permissible because the theater reproducing systems are generally adjusted for an 8000- to 10,000-Hz cutoff frequency, with some installations using a cutoff frequency of 6500 Hz. However as the frequency range is increased, the distortion must be lowered if good clean reproduction is to be had. It should be pointed out that some theater systems use no cutoff at all, but let the system drop off with the head characteristics. As most magnetic heads in projectors employ a head gap of 0.25 mil, the fall-off starts about 10,000 Hz, then falls quite rapidly. For magnetic tape cartridges, the same reference levels are used as given in the first paragraph.

17.140 What is the standard for distortion relative to ¼-inch tape?—The NAB Standard specifies the overall record-reproduce total harmonic distortion (THD) including the tape shall be less than 3 percent rms for 400-Hz sine wave, recorded to achieve a reproducing level 6 dB above the NAB Standard Reference Level. Recording levels for motion picture recording equipment are established somewhat differently. The standard is the same for magnetic tape cartridges. (See Question 17.139.)

17.141 What is a split-film recorder?—A magnetic film recorder using 17.5-mm full-coat film, with 35-mm perforations, running at linear speeds of 45 fpm. The term split-film is a carry-over from the early days when split photographic film was used in recorders and reproducers. The term is now obsolete.

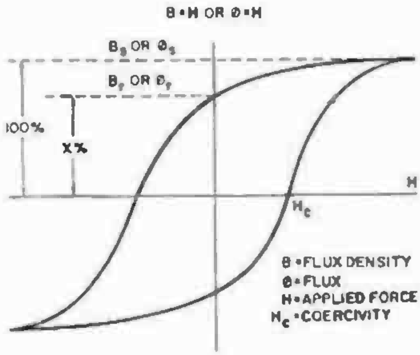


Fig. 17-142A. Typical B-H curve for magnetic tape or film.

17.142 Explain squareness factor, relative to magnetic tape and film.—Squareness factor for magnetic tape and film is the shape of the B-H curve of the magnetic-oxide coating and has a shape similar to that shown in Fig. 17-142A. It will be observed in this illustration, the upper and lower portions (termed the "knee" and "toe") are quite rounded and the sides are widely spread. An ideal curve would be one with straight sides (Fig. 17-142B); however, this is not achieved in actual practice. The term, squareness ratio, is a coined term, used by tape manufacturers, to provide a quantitative description of the hysteresis loop. Multiplication of ϕ_r/ϕ_s or B_r/B_s by 100, gives the squareness ratio in percentage, or X percent of the saturation magnetism which remains as useful remanent flux after the applied field is removed. This may be taken as a ratio of efficiency for tape and magnetic material. In reality, it is a plot of ϕ-H versus H. Since the permeability of a magnetic tape or film is quite low, it is more satisfactory in analyzing the magnetic properties of such materials to subtract or balance

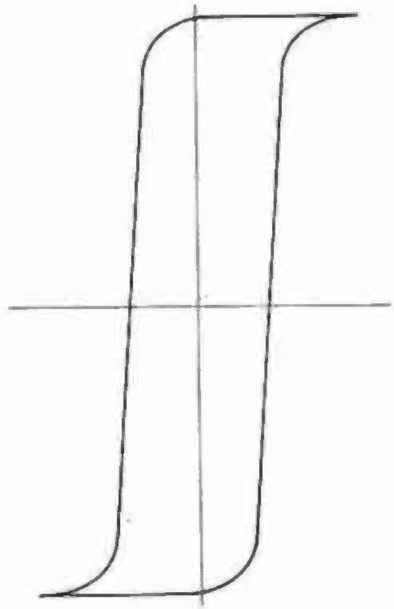


Fig. 17-142B. Ideal squareness factor.

out the applied field and measure only the flux increase caused by the sample.

$$\text{Squareness Ratio} = \frac{\text{Remanent flux } (B_r)}{\text{Saturation flux } (B_s)} \times 100$$

Quality tape and magnetic film generally have squareness ratio factors on the order of 75 percent.

Magnetic materials are characterized by the nonlinear characteristics between the magnetizing force and the resulting state of magnetization. This at first glance would tend to eliminate them for high-quality sound recording. Fig. 17-142C shows the relationship existing between the magnetization force H applied to the tape over the recording head gap and the resulting magnetization B, starting at 0 with a completely demagnetized tape. The magnetizing force H is proportional to the product of the number of turns in the coil of the recording head, and the current flowing through the windings. Assuming this force is H_1 at a given instant as the tape is about to pass out of the recording head gap, its magnetization will be B_1 while still in the gap, but will drop to B_r as it leaves the gap. B_r is the remanent magnetization on the tape after the magnetizing field has been removed, or the tape has left the head gap.

Plotting the values of remanent magnetization B_r corresponding to various

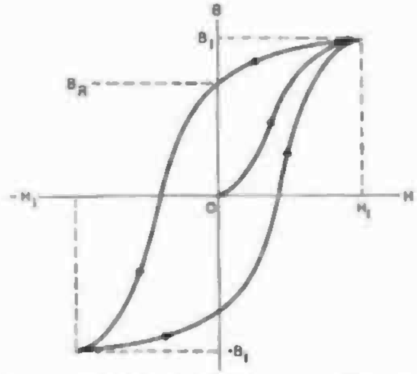


Fig. 17-142C. Typical magnetization curve existing between the magnetizing force applied to a magnetic tape or film, over the recording head-gap and the resulting magnetization.

values of magnetizing force H results in a curve, the actual recording characteristic of the recording medium (Fig. 17-142D). This plot illustrates the nonlinearity of the recording medium; however, portions of the characteristics are linear. By the application of a high-frequency bias current, the recording signal is moved into the more linear portion of the characteristic. This subject is discussed in Question 17.140. (See Question 23.158.)

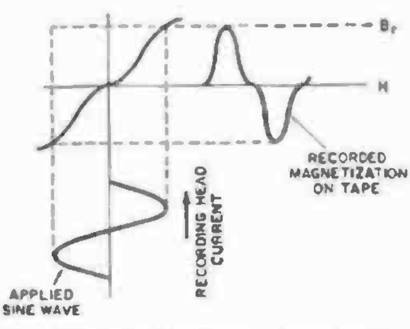


Fig. 17-142D. Nonlinear characteristics of magnetic tape or film without high-frequency bias.

17.143 What is a dual or two-track recorder?—In the early days of stereophonic recording, recorders were manufactured using two heads placed side by side. However, this method is now obsolete, with three and four tracks machines now taking their place. (See Question 17.147.)

17.144 What are the standards for flutter, relative to ¼-inch tape recorders?—The NAB Standard states that in the reproduce mode, the unweighted and weighted flutter content, when reproducing an essentially flutter-free recording of 3000 Hz at any portion of the reel, shall not exceed the values given in Fig. 17-144A. The standard further specifies that unweighted flutter measurements are to be made over a range of 0.5 to 200 Hz, falling off at a rate of 6 dB per octave, above and below these frequencies.

At the lower frequencies, where the meter movement follows the waveform, the maximum deflection shall indicate the rms value. The indicating meter is to have standard VU meter characteristics, using a full-wave rectified average measurement law, and must be calibrated to read the rms values of a sinusoidal variation. This latter specification conforms to VU meter USASI (ASA) Standard C16.5-1961, discussed in Question 10.3. The meter is read for random periods throughout the length of the tape, and noting the average peak readings, excluding random peaks that do not reoccur more than three times in any 10-second period. This is the average flutter.

Weighted flutter measurements are made in the same manner, using the same measurement system, except that a weighting network is used with the frequency characteristic given in Fig. 17-144B. In some type flutter meters the weighting network may be included in its circuitry, therefore the manual for a particular flutter bridge should be consulted. (See Question 5.98.)

The unweighted values of flutter given in Fig. 17-144A are the maximum values and are generally less in most professional equipment. Although the value of maximum flutter is not given for 30 ips, these machines usually have 0.10-percent flutter or less, using a

Tape speed	*Flutter (rms)*
Unweighted flutter	
15 ips	0.15 percent
7½ ips	0.20 percent
3¾ ips	0.25 percent
Weighted flutter	
15 ips	0.05 percent
7½ ips	0.07 percent
3¾ ips	0.10 percent

Fig. 17-144A. Maximum flutter content for NAB Standard recorders.

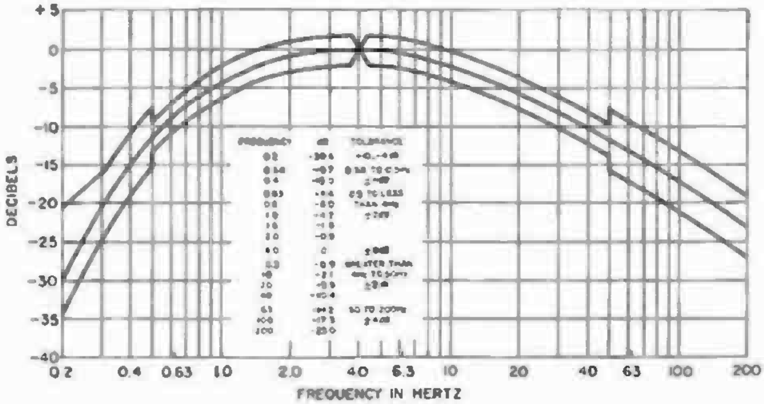


Fig. 17-144B. Frequency characteristics for weighted network used with flutter measurements. (Courtesy, DIN 45-507 Oct. 1962 National Association of Broadcasters.)

weighted response, while for $1\frac{7}{8}$ and $\frac{15}{16}$ ips, the unweighted flutter must not exceed 0.25 and 0.30 percent respectively. For magnetic tape cartridges the total flutter shall not exceed 0.3 percent. The frequency response for a weighted response network with the frequency tolerance is given in Fig. 17-144B. Magnetic film equipment may be measured in a similar manner.

17.145 Describe the purpose of a weighted curve flutter measurement.—Unweighted flutter measurements are made using a flutter bridge, with a flat frequency characteristic. Weighted flutter measurements are made using a weighting network, with a frequency characteristic as shown in Fig. 17-144B. This network provides a frequency characteristic similar to the average human ear hearing characteristic and for certain types of flutter is more realistic. However, when hearing is not involved, the flat characteristic is desirable.

17.146 What are the standard widths for magnetic film?—The standard widths for magnetic film are: 16 mm, 17.5 mm, and 35 mm. The 16 mm is run at 36 fpm; the 17.5 mm at 45 fpm; and the 35 mm at 90 fpm. The 16 mm and 17.5 mm are generally used for dialogue recording on motion picture production, although 16 mm may also be used at times for music recording. The 17.5 mm is a 35-mm film split down the center. Running 17.5-mm film at a speed of 45 fpm doubles the time of recording over 35 mm run at 90 fpm and has the same frequency response, but reduces the initial cost of stock.

17.147 Describe the various type heads used for magnetic tape and film recording.—In the manufacture of magnetic recording and reproducing equipment, one of the most closely guarded proprietary secrets is the manufacture of magnetic heads. Therefore, only the basic principles of design and construction will be discussed.

Fig. 17-147A. A 16-mm magnetic recording head.

Fig. 17-147B. A 35-mm magnetic recording head.

A group of recording and reproducing heads are shown in Figs. 17-147A to G manufactured by Lipps Inc. In Figs. 17-147A and B are edge-track heads for 16-mm and 35-mm magnetic film recording. The shoe at the upper end of the head supports the film parallel to the pole-piece surfaces and prevents warping of the film in its passage over the head.

Fig. 17-147C. A 3-track 35-mm head.

A three-track cluster for 35-mm film appears in Fig. 17-147C. Heads shown in Figs. 17-147D and E are for ½-inch magnetic tape. A four-track head cluster designed for 16-mm magnetic film is pictured in Fig. 17-147F, with a prealigned 8-track assembly, consisting of an erase, record and reproducing heads is shown in Fig. 17-147G.

To reduce cross talk between heads (see Question 17.149), Mumetal shields are inserted between the heads and grounded, with end shields used to prevent pickup from stray magnetic fields.

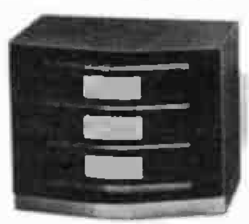


Fig. 17-147D. A 3-track head for ½-inch magnetic tape.

After the heads are assembled and aligned, they are encased in an epoxy-like compound, ground, and polished to shape. The round cases of the heads shown in Figs. 17-147A and B are Mumetal shields.

Recording and reproducing heads for motion picture recording equipment use sound tracks of 150 and 200 mils in width. The recording gap is on the order of 0.50 mil and the reproducer 0.25 mil.

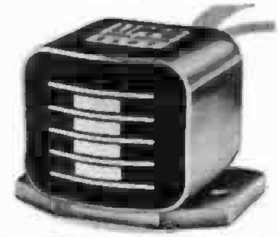


Fig. 17-147E. A 4-track head for ½-inch magnetic tape.

These larger gap heights may be used since the frequency response is generally not over 10,000 Hz. Head gaps for ¼-inch tape vary in both height and length, depending on the nature of the head, and vary also if for half or quarter track.

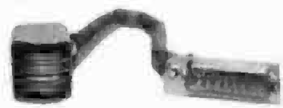


Fig. 17-147F. A 4-track head for 16-mm magnetic film.

The characteristics of a head are generally stated for a given bias and recording current with its inductance. Older type heads used high inductance; however, the inductances for newer heads are considerably lower and range from 400 to 500 mH for record-playback heads, and 30 to 65 mH for a single recording head. The inductance of a magnetic head for motion picture recording equipment is generally of low value, on the order of 6 mH, and is operated into an impedance-matching transformer.

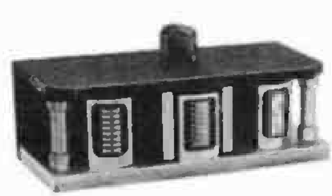
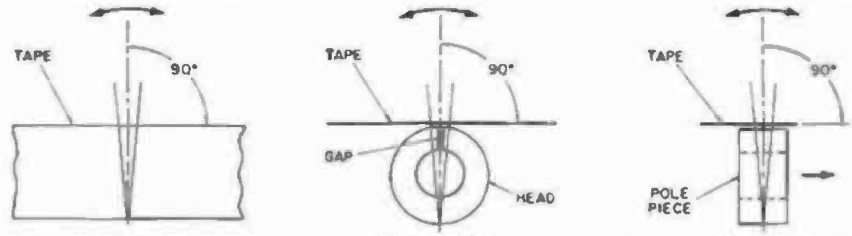


Fig. 17-147G. Prealigned 8-track assembly with record, reproduce, erase heads.

The use of this low impedance reduces the pickup from stray magnetic fields as the amplifiers in rack-mounted equipment are generally some distance from the head. Typical values for such heads are 6 to 30 ohms impedance at 400 Hz.

17.148 Describe the three basic adjustments for aligning magnetic heads.—Three basic adjustments are required to



(a) *Azimuth.* (b) *Zenith.* (c) *Parallelism of the pole pieces with the surface of the tape.*

Fig. 17-148. Three basic adjustments for aligning magnetic heads.

align a magnetic recording or reproducer head. They are: azimuth, zenith, and paralleling of the surfaces of pole pieces to the surface of the tape or film, as shown in Fig. 17-148. In some type machines all three adjustments have to be considered, and in others only the azimuth and zenith are adjusted. The parallel adjustment of the pole pieces is relative to the tape surface being fixed. For head assemblies such as that shown in Fig. 17-147G, the heads are prealigned and require no further adjustment. They are installed or removed by the locking lever at the top of the assembly.

17.149 What are the standards for signal-to-cross-talk ratio in 1/4-inch magnetic recorders?—The NAB Standard for cross talk specifies: for 1/4-inch two- and four-track stereophonic recorders, the adjacent signal-to-cross-talk ratio shall not be less than 60 dB in the range of 200 to 10,000 Hz. When making these measurements, the bias voltage is turned off. In the reproduce position, the separation between channels shall not be less than 40 dB between 100 and 10,000 Hz. For motion picture recording, signal-to-cross-talk level should be 60 dB or greater, and for reproduction, a separation of 45 to 55 dB. For 1/4-inch tape, the NAB standard reference is used; for motion picture recording, the system is set up as discussed in Questions 17.135 and 17.139.

For magnetic tape cartridges, the signal-to-cross-talk ratio for monophonic reproduction is such that cue-tone to program-channel system cross talk at the NAB standard reference level is to be not less than the following:

At 150 Hz, 50 dB
At 1000 Hz, 55 dB
At 8000 Hz, 50 dB.

For stereophonic reproduction under the same conditions, not less than 50 dB.

17.150 Show the sound track placement for dual-track 1/4-inch tape recorders.—The dimensions for 1/4-inch multi-sound track recorders are shown in Figs. 17-150A, B, and C. In the four-track arrangement, the tracks are centered on the tape with the heads evenly dispersed across the tape width. The

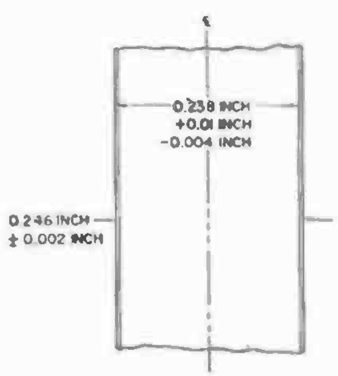


Fig. 17-150A. Full-track monophonic or stereophonic 1/4-inch tape recording.

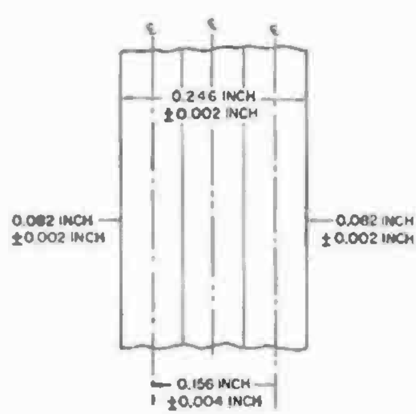


Fig. 17-150B. Two-track (half-track) monophonic or stereophonic 1/4-inch tape recording.

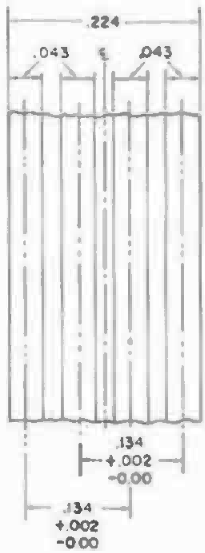
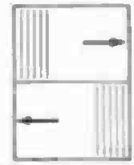
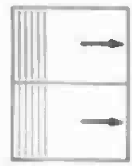


Fig. 17-150C. Sound track dimensions for four-track, ¼-inch tape, monophonic or stereophonic recording.

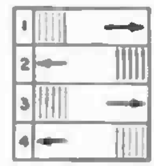
(a) *Full-track.*



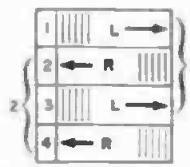
(b) *Half-track monophonic.*



(c) *Half-track stereo.*



(d) *Four-track monophonic.*



(e) *Four-track stereo.*

Fig. 17-151. Sound track combinations used with ¼-inch magnetic tape recorders. These positions are in accordance with the NAB Standard April 1965.

outer edges of tracks 1 and 4 are coincident with the edges of the tape. The alignment is based on a tape width of 0.244 inch.

17.151 Give the terminology used with ¼-inch multitrack recording.—Much confusion has arisen in the terminology used with ¼-inch tape recorders sold to the public for home recording. Such recorders are referred to as half-track, quarter-track, and four-track machines. This terminology can be clarified by referring to Fig. 17-151. The direction of tape travel is indicated by the arrows. In the stereophonic versions, the heads are made in clusters of two.

At part (a) is a full-track monophonic sound track recorded by using a single head covering the full width of the tape. At part (b) is a half-track or two-track monophonic recording, using two heads, each slightly less in width than half the tape width. At part (c) is a half-track stereophonic recording. Referring to part (d), a quarter-track monophonic recording is shown using four heads slightly less in width than one-quarter the actual tape width. At part (e) is shown a quarter-track stereophonic recording, using four heads. The illustrations show the appearance (if they could be seen) of the sound tracks as viewed from the back of the tape (smooth side).

It will be observed at part (b) that the sound track is recorded first to the right and then to the left, while at part (c) the sound tracks are both recorded to the right. For four-track monophonic at (d), tracks 1 and 3 are recorded to the right and tracks 2 and 4 to the left. Thus, long playing selections can be played back by running the tape through the machine in the usual manner, rather than transferring from the take-up position to the feed-position and rewinding.

It is standard in two-track stereophonic recorders for track 1 to carry the left channel as viewed from the audience, and track 2 the right hand channel. The recording head-gaps are to be of the in-line type phased for reproduction on equipment so connected that when a full-track recording is reproduced, it produces in-phase signals in the two-channel reproducer outputs.

For stereophonic four-track recordings, sound tracks 1 and 3 are used simultaneously for one direction of travel, and tracks 2 and 4 for the reverse. Tracks 1 and 3 are used first as the tape is unwound from the supply reel. Tracks 1 and 4 carry the left channels and 2 and 3 the right channels.

Equipment designed for the rerecording of motion picture use multiple sound tracks and heads; however, the terminology is quite different. Such recorders do not use reverse sound tracks. Three- and four-track head machines may be used for either monophonic or stereophonic recording. The input of each sound track recording amplifier appears in a patch-bay and may be selected for any combination of recording. For three-track monophonic recording, the dialogue, sound effects, and music are recorded on separate sound tracks, through a three-section mixer console as described in Question 9.46. Four-track machines used for stereophonic recording, carry left, center, and right sound tracks in that order, with the fourth track carrying miscellaneous material. Later, these tracks may be combined in other combinations and a submaster made for transferring to theater release prints, as each print is an original recording transferred from the submaster (second generation master). This method of making release prints is discussed in Question 17.202.

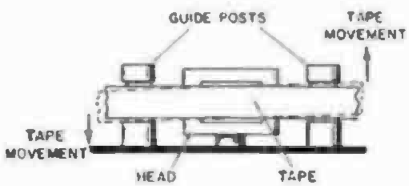


Fig. 17-152A. Worn guide posts permit the tape to wobble or skew in its travel across the head.

17.152 What are the factors affecting head wear?—With the components guiding the tape and the heads properly aligned, head wear will be at a minimum. However, even under the best conditions, and using high quality tape, the heads will require replacement in time. Fig. 17-152A shows the effect of worn guide posts on the tape travel. Here the post is worn to a point where the tape can skew or wobble as it passes over the head. This condition will cause the tape to go in and out of azimuth, even with a properly aligned head. A cross-sectional view of an improperly aligned head cluster is shown in Fig. 17-152B. If the surfaces of the heads are

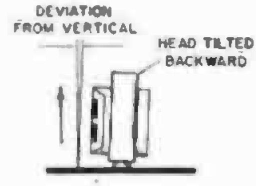


Fig. 17-152B. Head out of vertical alignment.

not perpendicular to the tape travel, the tape will tend to skew in the direction indicated by the arrow.

Head wear is generally indicated by the appearance of the wear pattern and the necessity to increase the bias current (assuming the same brand and type tape is being used as when the heads were originally installed), and the loss of high-frequency response. Typical wear patterns are given in Figs. 17-152C and D. In Fig. 17-152C the wear pattern indicates that the vertical alignment is such that the head is tilted too far back at the top and more pressure is being applied to the lower portion of the tape. In the two-head cluster in Fig. 17-152D, the keystone wear pattern indicates that the alignment in the vertical plane is tilted too far forward, applying more pressure to the upper head than to lower.

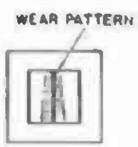


Fig. 17-152C. Typical wear pattern encountered for head tilted back at top.

Worn recording heads are generally manifested by one or more of the following: peak bias current decreases actual bias current increases, inductance of the head decreases, recording sensitivity increases with an increase of recording current, and if the head is permitted to wear enough, the linearity of the gap decreases and opens. If the recorder employs pressure pads greater wear may be expected.

In the playback position the sensitivity may increase with an increase of high-frequency response, with respect to the low-frequency response. In the recording process, the increase of bias current reduces the high-frequency response by overbiasing.

Head life is principally governed by the design of the machine, head alignment, and the quality of the tape used. Factors such as humidity and dust also have their effect. Life expectancy will run from 400 hours to as high as 4000 hours of use, again depending on the design, alignment, and local environment. A tape recorder used 5 hours a day, running at a linear speed of 3.75 ips, passes 640 miles of tape over the head surfaces; this is a good illustration of why the head wear increases so rapidly.

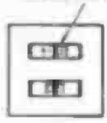


Fig. 17-152D. Typical wear pattern encountered for heads that are tilted forward at top.

17.153 Give recording time versus linear speed and length of tape.—Recording time for various tape speeds and reel size is given in Fig. 17-153. By the use of a 1-mil base, the footage capacity for a given reel size is increased by 50 percent.

17.154 Describe a noise reduction circuit as used in ¼-inch tape recorders.—A method sometimes used in ¼-inch tape recorders to cancel the effects of residual magnetization in recording heads is shown in Fig. 17-154A. A dc voltage is applied to the recording head. Residual magnetism is cancelled by running an unmodulated tape through the machine with only the recording bias current on. A vacuum-tube voltmeter is connected across the playback output and the noise cancelling potentiometer is adjusted for a minimum noise indication on the meter. It should be pointed out that the lowest noise setting is not always the point of lowest distortion; therefore, the distortion should be measured after making the noise-reduction adjustment, and a compromise made if necessary. Magnetization of the recording head can be caused by an unsymmetrical waveform in the bias oscillator.

A second method of noise adjustment is shown in Fig. 17-154B. Here a control is connected in the grid returns of the oscillator tubes and is used to adjust the symmetry of the oscillator waveform. It is similar to the circuit used in most of the modern recorders employing transistors.

17.155 Describe the procedure for adjusting the equalization of a ¼-inch tape recorder.—The azimuth and bias current are adjusted first, and the playback equalization is set by the use of a standard test tape. After adjusting the playback equalization for as near a uniform response as possible, a good quality tape is threaded on the machine and a series of test frequencies recorded using the same frequencies as on the test tape. This tape is then played back and the frequency response noted. If it does not fall within the limits of the test tape, the recording equalization is adjusted until such a response is obtained. The final playback curve is then the record-playback characteristics of the machine. The recording level should be at least 10 dB below 100-percent modulation of the normal recording level. If the machine is equipped with a separate reproduce head, the effect of changing the recording equalization may be monitored.

17.156 How is the time delay between a recording and monitoring head calculated?—The time delay in seconds may be calculated:

$$\text{Time} = \frac{D}{V}$$

where,

D is the distance between the gaps of the two heads in inches,

V is the linear velocity of the tape in inches per second.

17.157 What is the equation for calculating the wavelength of a recorded frequency?

$$\text{Wavelength} = \frac{V}{F}$$

where,

V is the linear velocity of the tape in inches per second,

F is the frequency in Hz.

In magnetic tape or film recording, the velocity is constant; therefore, it has little effect on the frequency response except to limit the highest frequency which may be recorded. A graph from which the wavelength for a given frequency and velocity may be read with-

REEL SIZE →	3"	4"	5"		7"		10½"		14"
TAPE FOOTAGE →	150'	300'	600'	900' *	1200'	1800' *	2500'	3600' *	5000'
TAPE SPEED IN. PER SEC.	SINGLE TRACK RECORDING								
$\frac{15}{16}$"	32 min.	1 hr. 4 min.	2 hr. 8 min.	3 hr. 12 min.	4 hr. 16 min.	6 hr. 24 min.	8 hr. 48 min.	12 hr. 48 min.	17 hr. 44 min.
1⅞"	16 min.	32 min.	1 hr. 4 min.	1 hr. 36 min.	2 hr. 18 min.	3 hr. 12 min.	4 hr. 24 min.	6 hr. 24 min.	8 hr. 52 min.
3¾"	8 min.	16 min.	32 min.	48 min.	1 hr. 4 min.	1 hr. 36 min.	2 hr. 12 min.	3 hr. 12 min.	4 hr. 26 min.
7½"	4 min.	8 min.	16 min.	24 min.	32 min.	48 min.	1 hr. 6 min.	1 hr. 36 min.	2 hr. 13 min.
15"	2 min.	4 min.	8 min.	12 min.	16 min.	24 min.	33 min.	48 min.	1 hr. 6 min.
30"	1 min.	2 min.	4 min.	6 min.	8 min.	12 min.	16 min.	24 min.	33 min.
	DUAL TRACK RECORDING								
$\frac{15}{16}$"	1 hr. 4 min.	2 hr. 8 min.	4 hr. 16 min.	6 hr. 24 min.	8 hr. 32 min.	12 hr. 48 min.	17 hr. 36 min.	25 hr. 36 min.	35 hr. 28 min.
1⅞"	32 min.	1 hr. 4 min.	2 hr. 8 min.	3 hr. 12 min.	4 hr. 16 min.	6 hr. 24 min.	8 hr. 48 min.	12 hr. 48 min.	17 hr. 44 min.
3¾"	16 min.	32 min.	1 hr. 4 min.	1 hr. 36 min.	2 hr. 8 min.	3 hr. 12 min.	4 hr. 24 min.	6 hr. 24 min.	8 hr. 52 min.
7½"	8 min.	16 min.	32 min.	48 min.	1 hr. 4 min.	1 hr. 36 min.	2 hr. 12 min.	3 hr. 12 min.	4 hr. 26 min.
15"	4 min.	8 min.	16 min.	24 min.	32 min.	48 min.	1 hr. 6 min.	1 hr. 36 min.	2 hr. 13 min.
30"	2 min.	4 min.	8 min.	12 min.	16 min.	24 min.	33 min.	48 min.	1 hr. 6 min.

* Magnetic tape using a Mylar polyester base will provide 50 percent more footage on a given reel size, because of the thinner base material.

Fig. 17-153. Recording time for various tape speeds and reel sizes. (Courtesy, Audio Devices, Inc.)

out calculation is given in Fig. 17-157. As an example: the wavelength of 1000 Hz at a velocity of 1.75 inches per second is 0.00175 in. In other words, one wavelength of 1000 Hz recorded at a velocity of 1.75 inches per second will cover a space of 0.00175 in. on the tape.

At a velocity of 18 inches per second (35-mm film running at 90 feet per minute), the same wavelength will

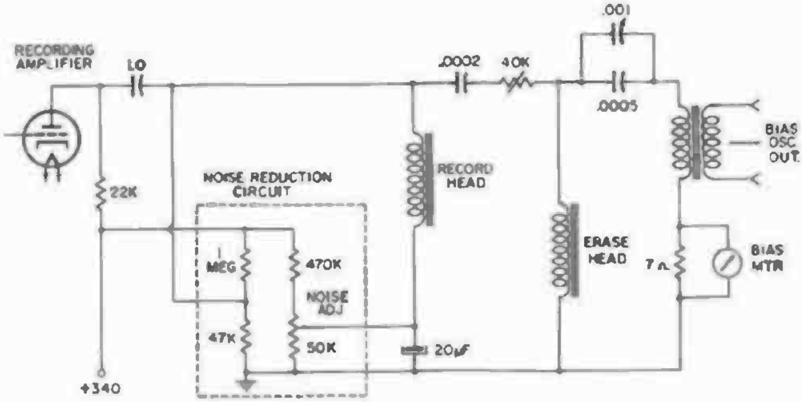


Fig. 17-154A. Noise reduction and erase head coupling circuits used in a ¼-inch magnetic tape recorder.

cover a space of 0.018 inch and at 30 inches per second, a space of 0.03 inch.

17.158 How are tape speeds measured?—The NAB Standard speed is 7½ ips, plus-minus 0.2 percent. Supplementary speeds 15 and 3½ ips, plus-minus 0.2 percent. The tolerances are applicable to any portion of the reel in use. The linear speed is measured by applying a precision pulley on precision bearings one-quarter inch in width to the surface of the tape, between the capstan and the head assembly. The rotational speed of the pulley as it is driven by the tape, may be measured by an ac tachometer or by a stroboscopic disc, mounted on the upper surface of the pulley. (See Question 13.116.) The stroboscopic bars are illuminated by a neon light supplied from the power line. Measurements are made at normal room temperature. The tape thickness is to be 0.0019 inch, plus-minus 0.0002 inch, corresponding to the thickness of a tape using a normal base thickness of 1.5 mils.

The design calls for a pulley with a diameter of 1.4305, plus 0.0002 inch minus 0.000-inch, upon which is attached a stroboscopic disc, having 72 and 36 equally spaced dots or lines. A neon lamp operating from a 60-Hz power source flashes at 120 Hz. When the disc is illuminated by this lamp, it will indicate the 7½- and 15-ips tape speeds. For 3¾ ips, a diode is connected

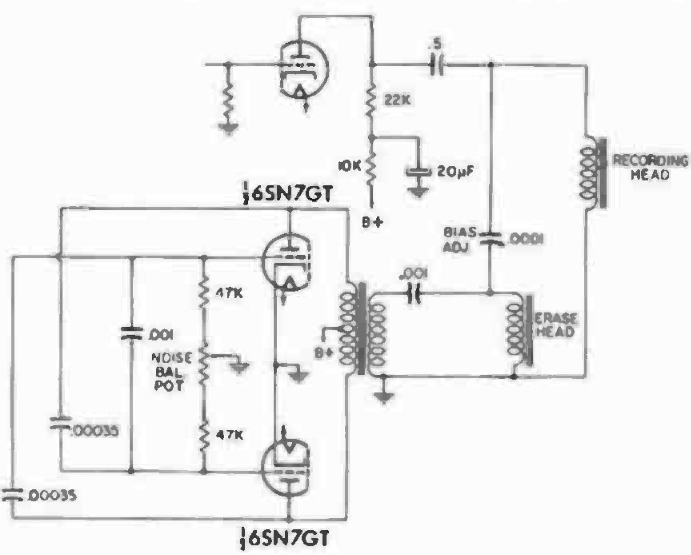


Fig. 17-154B. A noise reduction circuit with the noise adjustment potentiometer in the grid circuit of the push-pull bias oscillator.

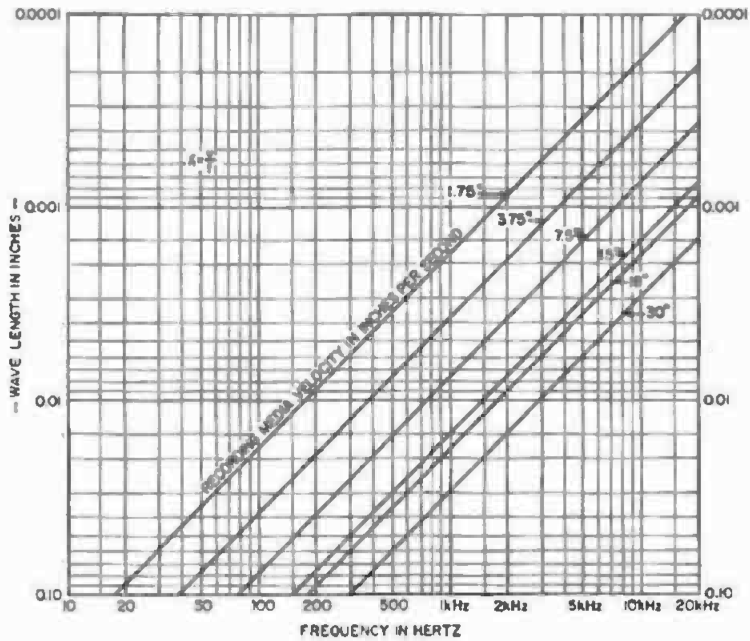


Fig. 17-157. Velocity of recording media versus recorded wavelength in inches for a given frequency.

in series with the lamp, causing it to flash at 60 Hz. No more than 14 dots per minute are permitted to drift past a fixed reference point in either direction for 7½ or 15 ips; for 3¾ ips, the drift per minute is not to exceed 7 dots on the 36-dot disc. The limits given correspond to a speed tolerance of plus-minus 0.20 percent for the specified speed.

17.159 What are the standards for ¼-inch tape relative to signal-to-noise ratio?—The NAB Standard specifies that the noise be measured over a frequency range of 20 to 20,000 Hz. The response of the measuring circuit is to be plus or minus 0.3 dB 30 to 15,000 Hz. 20,000 Hz is to be 3 dB below the 400-Hz level, and falling off at a rate of 12 dB per octave thereafter. The method used is to record a 400-Hz signal, cut the signal off, and let the tape continue to run with only the bias current on. The tape is then played back, and the noise level is measured in reference to the standard signal, with or without a weighting network.

Weighted curves are made using the network given in Question 5.98. Unweighted curves are made using the measuring system with uniform frequency characteristics as described above. A weighted network simulates the human ear characteristics at frequencies below 1000 Hz. (See Question 2.93.)

Signal-to-noise ratio, when given in terms of the NAB Standard, is a figure of merit for the comparisons of noise between recording systems. It does not take into account the program level which may be recorded without excessive distortion. Therefore, the noise level relative to the program level peaks may be better by 8 to 10 dB than the figures given. It will be noted that the signal-to-noise ratio is lower at 15 ips than at 7½ ips, using a weighted curve. This is due to the playback equalization

Unweighted signal-to-noise measurements.			
Tape speed	*Full-track*	*Two-track*	*Four-track*
15 ips	50 dB	45 dB	Not used
7½ ips	50 dB	45 dB	45 dB
3¾ ips	46 dB	46 dB	45 dB
Weighted signal-to-noise measurements.			
15 ips	58 dB	53 dB	Not used
7½ ips	60 dB	55 dB	52 dB
3¾ ips	57 dB	54 dB	52 dB

Fig. 17-159A. NAB Standard recorders, using unweighted and weighted measurements.

being the same for both speeds, while the tape noise increases with tape speed.

The use of a weighted signal-to-noise ratio measurement is desirable as it results in a more realistic indication of the subjective signal-to-noise ratio than does the unweighted response. Minimum values of signal-to-noise ratio for various types of recorders are given in Fig. 17-159A. It should be remembered that these values are the minimum values and are generally exceeded. It is not uncommon for a certain type of equipment to have a signal-to-noise ratio of 80 dB.

Special purpose limited performance systems, unweighted.			
Full-track		46 dB	
Two track		43 dB	
Four track		40 dB	
Professional equipment, weighted.			
Number of tracks	*Tape width*	*Track width*	*15 or 7½ ips*
1	¼ inch	0.234 inch	70 dB
2	¼ inch	0.074 inch	65 dB
2	½ inch	0.200 inch	69 dB
3	½ inch	0.100 inch	66 dB
4	½ inch	0.070 inch	65 dB
3	1 inch	0.250 inch	70 dB
4	1 inch	0.180 inch	69 dB
6	1 inch	0.095 inch	66 dB
8	1 inch	0.070 inch	65 dB

Fig. 17-159B. Signal-to-noise ratios for NAB special purpose limited-performance recorders and for professional equipment, using ¼-inch to 1-inch tape, with multiple-track heads.

Fig. 17-159B shows a table of signal-to-noise ratios for recorders using ¼ to 1-inch tapes and multiple heads. It will be observed the minimum value is 65 dB and the maximum is 70 dB.

For cartridge magnetic tape reproduction, the signal-to-noise level for monophonic recording is 45 dB, and for stereophonic it is 42 dB, measured in the manner given above.

17.160 What is Barkhausen noise in a magnetic recorder?—Barkhausen noise is modulation noise or behind-the-signal noise. It may be caused by tape which has become magnetized or has an uneven coating. Barkhausen noise, viewed on an oscilloscope, has an appearance as shown in Fig. 17-160.

17.161 What effect do warped reels have on a magnetic recorder?—The wandering of the tape, caused by the side of the reel striking the tape, can cause poor frequency response, noise, and distortion. Also, it may induce a considerable amount of flutter in the transport system.

17.162 What are the frequency-response limits for ¼-inch tape recorders?—The reproducing tolerances for NAB Standard recorders are given in Fig. 17-162A, with the recording limits given in Fig. 17-162B. The reproducing limits in a positive direction are not to exceed those shown when using a NAB Standard test tape or its equivalent. It is recommended that the response above and below these limits be rolled-off at a rate of 6 dB per octave. Since all NAB test tapes are recorded full-track, a low-frequency rise may be expected when they are played back on a two-track, half-track, or four-track machine. The rise at the low-frequency end is due to the fringing effect and head bumps discussed in Question 17.199. The tolerances shown in Fig. 17-162B are the maximum limits the recording characteristic can vary and still meet the reproducing characteristic shown in Fig. 17-162A.

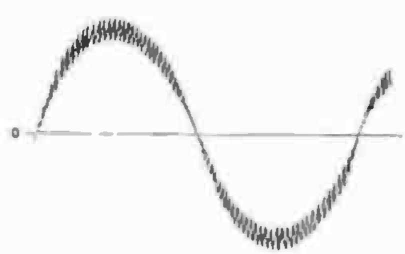


Fig. 17-160. A Barkhausen, or modulation pattern. This is also called behind-the-signal noise.

If the recorder meets the requirements of these characteristics, relative to signal-to-noise, distortion, cross-talk-to-signal ratio, flutter, and other requirements as set forth in the Standard, the machine may be classified as a NAB Standard recorder. For portable and special-service systems, the Standard is as given in Fig. 17-162C. Recorders within these limits are classified as *Special Purpose Limited Performance Systems.* It will be observed that for this class machine, the tolerances are not as stringent as for the first classification.

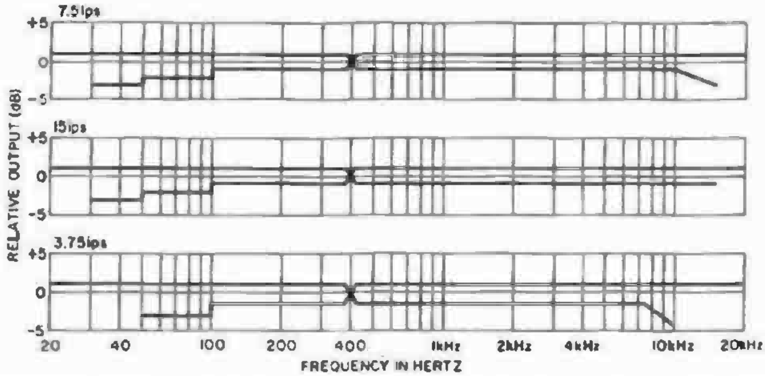


Fig. 17-162A. NAB playback frequency tolerance limits for equipment classified as NAB Standard recorders.

For this latter classification, tape speeds have a tolerance of plus-minus 2 percent, flutter not to exceed 0.5 percent, signal-to-noise ratio of 46 dB for full-track and 40 dB for a four-track machine. The measurements are performed as for the NAB Standard recorder.

For magnetic tape cartridge reproduction, the response falls within the limits of Fig. 17-162A for 7.5 ips. The recording tolerances are those of Fig. 17-162B for 7.5 ips.

17.163 Describe the procedure for setting VU meter lead in recording systems.—Because of the complex nature of dialogue, music, and sound effects, such waveforms cannot be recorded at the same level as a sine wave. In complex waveforms, peaks may be encountered that are 8 to 12 dB higher than those indicated on a standard VU meter (USASI (ASA) C16.5-1961). These peaks can cause severe overloading of the recording system, with consequent

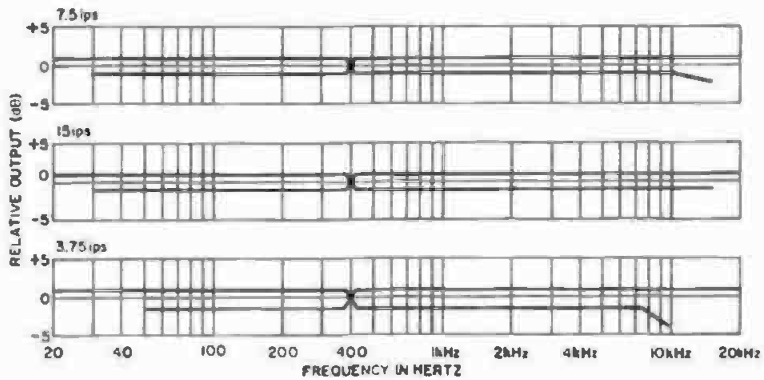


Fig. 17-162B. NAB recording frequency tolerance for equipment classified as NAB Standard recorders.

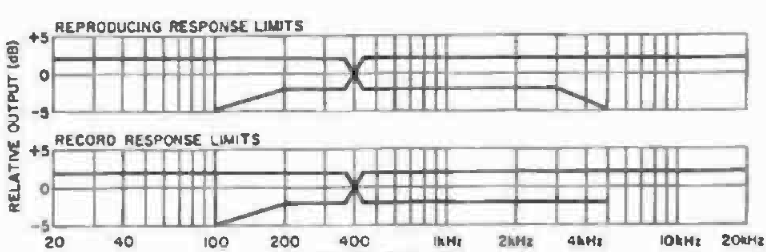


Fig. 17-162C. NAB record-reproduce limits for special purpose limited-performance systems, running 1⅞, 3¾, and 7½ ips.

distortion. Very often, when this is heard on playback it is taken for excessive sibilance, while it is actually high-frequency distortion. To prevent overloading, the VU meter is given a lead or made more sensitive to compensate for the unseen peaks. As an example, a recording channel is lined up for a bridging bus level of plus 12 dBm (100 percent on the VU meter) using a sine wave of low distortion (Fig. 17-163). The distortion at plus 12 dBm is 1 percent and is the desired operating level. An 8-dB lead is inserted in the VU meter by turning back the meter attenuator to plus 4 dBm; thus, the meter is now 8 dBm more sensitive and compensates for the unseen peaks. A lead of 6 dB is the very minimum.

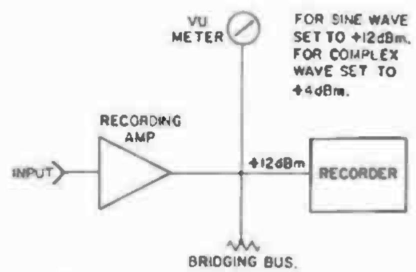


Fig. 17-163. Procedure for setting VU meter.

If only dialogue is being recorded and it is an original sound track to be used for a 16-mm picture, the lead is the same and the dialogue is recorded at an average level of 75 to 80 percent, or about 2 to 3 dB below the 100-percent indications of the meter. For 35-mm recording, the average level is 50 to 60 percent or 6 dB below 100 percent. For straight narration tracks, the average level is approximately 75 percent.

The same reasoning may be applied to any type recording system. However, in some recording activities, a lead of 10 to 12 dB is used. It is suggested that the reader consult Section 10 before making such adjustments. If a peak-indicating meter is used rather than a VU meter, no lead is used; 100 percent on the meter equals 100-percent modulation of the recording channel. (See Questions 17.139, 17.159, and Fig. 18-82.)

17.164 Describe the difference between standard and high-output oxide coatings.—Two types of magnetic film are available; the redish-brown standard oxide, and the dark green high-output oxide. The high-output type will permit a greater recording level to be used, thus increasing the dynamic range and the signal-to-noise ratio for the same percentage distortion. The average recording level between the two types is 8 dB. The frequency response of the two coatings is approximately the same except the high-output type has a greater sensitivity at the lower frequencies, and thus has a rising characteristic. The magnetic coating is 0.65 mil as compared to the 0.55 mil of the red oxide. Bias current requirements are approximately the same. High-output magnetic film is principally used with 16-mm recording eqnipment and with 35-mm full-coat for music recording.

17.165 What is a boundary displacement or borderline magnetic recorder?—A magnetic recorder which records on either tape or magnetic film, but magnetizes the tape to saturation at all times. The conventional high-frequency bias oscillator is not used.

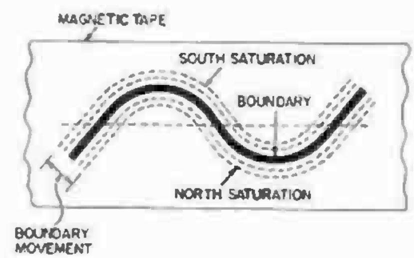


Fig. 17-165. A magnetic sound track recorded with a boundary displacement recorder.

With no modulation applied to the recording head of a boundary displacement recorder, one-half the tape is polarized opposite to the other half by means of two small permanent magnets mounted on each side of the recording head. An unmagnetized boundary in the center of the tape divides it into North and South polarized areas. (See Fig. 17-165.) The boundary is shown as a heavy black line between the two polarized areas. When the audio-frequency currents are applied to the recording head, the unmagnetized boundary is displaced in proportion to the modulating currents applied to the recording head. Standard reproducer heads are used to reproduce the sound tracks, the head scanning the entire boundary displacement area. The signal voltage is generated, when reproducing, by the side to side motion of the boundary line and may be likened to a variable-area optical film recorder described in Section 18.

The advantages claimed for this type recording are:

a. The maximum output corresponds to full tape saturation.
b. A high degree of amplitude linearity is obtained.
c. The dynamic range is equal to a magnetic recorder using the conventional high-frequency bias oscillator.

The disadvantages of a boundary recorder are:

a. The harmonic distortion is greater than for the conventional magnetic recorder.
b. Special recording heads are required.
c. The noise level is higher than in the conventional recorder and is comparable to a dc erasure system as described in Questions 17.83 and 17.85.

17.166 What are the dimensions for 16-mm center-track recording?—Although the center-track placement has been superseded by edge-track recording, many 16-mm magnetic film recorders still employ this track placement. Center-track placement may be used with either single- or double-perforated stock; however, the double perforation is preferred. Dimensions for the center-track placement are given in Fig. 17-

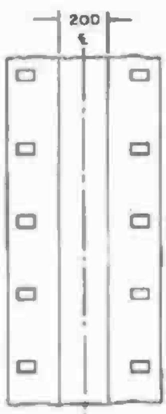


Fig. 17-166. Center sound-track placement with track dimensions for 16-mm magnetic film.

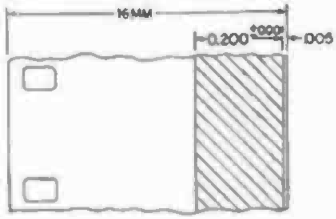


Fig. 17-167. Sound track placement dimensions for 16-mm magnetic film.

166. Center-track placement is now considered to be obsolete.

17.167 What are the sound-track placement measurements for a 16-mm magnetic film recorder?—They are as shown in Fig. 17-167.

17.168 What are the sound-track placement dimensions for 17.5-mm magnetic film?—They are shown in Fig. 17-168.

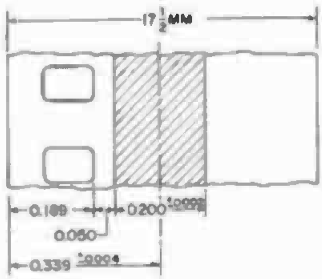


Fig. 17-168. Sound track placement dimensions for 17.5-mm magnetic film.

17.169 What are the sound-track placement dimensions for 3- and 4-track magnetic film recorders?—For studio equipment, dimensions are as given in Figs. 17-169A and B. Sound-track placement for release prints is given in Section 19.

17.170 Show the frequency response for magnetic recording on 8-mm film.—The frequency response specified in USASI (ASA) PH 22.134-1962 Standard, sponsored by the SMPTE, is shown in Fig. 17-170. The film speed is to be 24 perforations per second, approximately 18 feet per minute, with a mean film-speed tolerance of plus-minus 0.5 percent.

17.171 Where is sprocket tape used?—Sprocket tape is standard ¼-inch magnetic tape, with 16-mm film perforations along one edge, as in Fig. 17-30E. This tape was developed for use with small, light battery-operated recorders, similar to any other sprocket type recorder, before the advent of the ¼-inch sync-pulse recorders described in Question 3.78 and 17.179. The sound track has a width of 90 mils and records on the edge opposite from the sprocket holes at a speed of 36 fpm. The equalization characteristics are those of 16-mm magnetic film, given in Fig. 17-174.

17.172 Show typical recording pre-equalization characteristics for tape recording, using NAB, AME, and CCIR

American Standard Dimensions for

200-Mil Magnetic Sound Records on 35mm and 17 1/2mm Motion Picture Film

ASA
Reg. U.S. Pat. Off.
PH22.86-1962
Revision of
PH22.86-1953
*UDC [illegible]

1. Scope

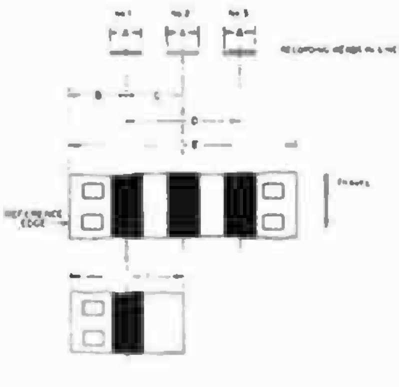
1.1 This standard specifies the locations and dimensions of magnetic sound records, both single and multiple tracks, on 35mm motion-picture film, and single tracks on 17½mm motion-picture film.

1.2 The sound records are determined by the lateral dimensions and position of the magnetic recording head.

1.3 This standard relates the placement of the magnetic coating on the film to the direction of film travel.

2. Dimensions

The dimensions and position of magnetic track No. 1 shall be determined by magnetic head No. 1, as specified in the diagram and table. The positions of tracks two and three shall be determined by the lateral dimensions of magnetic recording heads No. 2 and No. 3, as shown in the diagram.



Dimensions	Inches	Millimeters
A	0.200 ± 0.002	5.08 ± 0.05
B	0.339 ± 0.002	8.61 ± 0.05
C	0.350 ± 0.002	8.89 ± 0.05
D	0.700 ± 0.002	17.78 ± 0.05
E	1.377 nom	35.00 nom
F	0.689 nom	17.50 nom

3. Magnetic Coating

With the direction of film travel shown in the diagram, the magnetic coating shall be on the upper side of the film base.

4. Preferred Track Position

Track No. 1 is the preferred position for 35mm single-track recording and is the only position for 17½mm recording.

5. Recording and Reproducing Speed

Recording and reproducing speed shall be 96 perforations per second (see American Standard 35mm Photographic Sound Motion-Picture Film, Usage in Camera, PH22.2-1961). This is equivalent to 24 frames per second (approximately 18 inches per second).

6. Revision of American Standard Referred to in This Document

When the following American Standard referred to in this document is superseded by a revision approved by the American Standards Association, Incorporated, the revision shall apply:

American Standard 35mm Photographic Sound Motion-Picture Film, Usage in Camera, PH22.2-1961

NOTE: The dimensions in the inch system are the fundamental standard. The dimensions in the metric system are practical approximations based on American Standard Inch-Millimeter Conversion for Industrial Use, B48.1-1933, reaffirmed in 1947, which provides a conversion factor of 1 inch = 25.4 millimeters.

Approved April 25, 1962, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers

* Universal Decimal Classification

Printed in U.S.A.
ASA3/[illegible]

Fig. 17-169A. USASI (ASA) Standard for the 35-mm and 17.5-mm sound track placement.

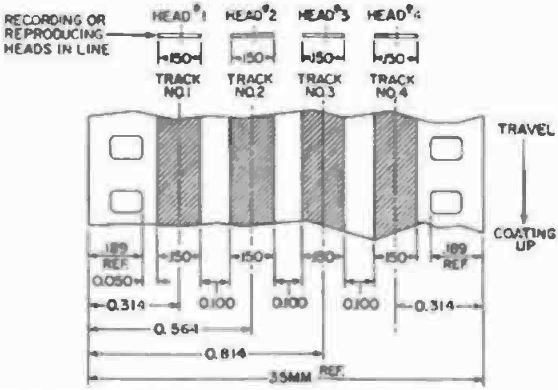


Fig. 17-169B. USASI (ASA) Standard PH22-108-1958 reaffirmed in 1965.

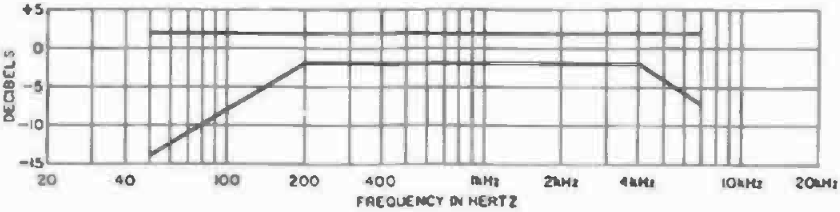


Fig. 17-170. Frequency tolerance limits for recording on 8-mm magnetic film.

characteristics.—Typical recording pre-equalization characteristics for tapes of 3¾ to 15 ips are given in Figs. 17-172A to C. As it was pointed out in Question 17.155, the reproducing equalization must be adjusted first by the use of a Standard test tape. After playback equalization has been set, the recording equalization is adjusted to produce a characteristic that conforms to the NAB Standard reproducing characteristic. The recording equalization curves shown in Figs. 17-172A to C are only a guide to what is generally required and represents no particular recorder. The equalization varies slightly from recorder to recorder, depending on the head characteristics, electronics, tape speed, and many other factors. However, the recorder reproducing response must fall within the NAB Standard limits to start, then the recording equalization must be adjusted for a record-playback response that falls within the reproducing limits. Under these conditions the recorder, insofar as the frequency response is concerned, meets the NAB Standard reproducing characteristic. (See Question 17.162.)

17.173 Show the electrical characteristics required in a ¼-inch tape recorder playback amplifier using NAB, AME, and CCIR equalization.—Such electrical characteristics are given in Figs. 17-173A to C. In adjusting the equalization in a preamplifier, it is convenient to adjust the equalization by use of an injection measurement discussed in Question 17.217. The CCIR response is used in Europe, and the AME (Ampex Master Equalization) is used for special recordings in the

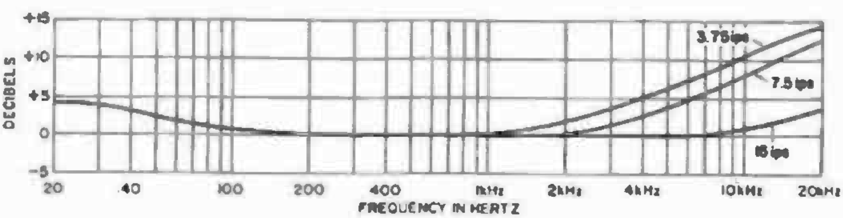


Fig. 17-172A. Typical recording (pre-equalization) for ¼-inch tape recorders using NAB characteristics.

United States. Commercial recorders are sometimes equipped with plug-in equalizer boards having these characteristics.

17.174 What are the pre- and post-equalization characteristics for 16-mm magnetic film? — Typical equalization requirements are given in Fig. 17-174 and are held to within plus-minus 1 dB, with reference to 1000 Hz. (See Question 17.175.)

17.175 What are the pre- and post-equalization characteristics required for 35-mm magnetic film?—Typical characteristics are given in Fig. 17-175. It will be observed that only a small amount of high frequency pre-equalization is employed, while considerable low fre-

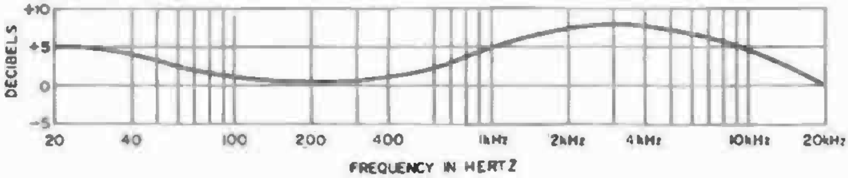


Fig. 17-172B. Typical pre-equalization for Ampex Master Recoding Equalization (AME) at 15 ips.

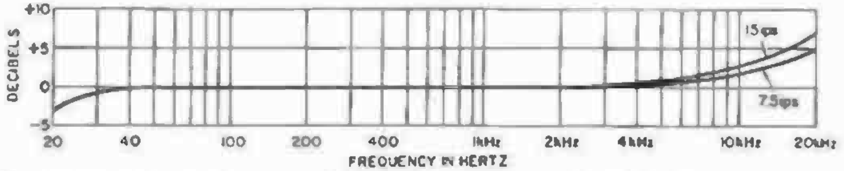


Fig. 17-172C. Typical pre-equalization characteristics for ¼-inch tape recorders running 7.5 and 15 ips using the CCIR (DIN) Standard.

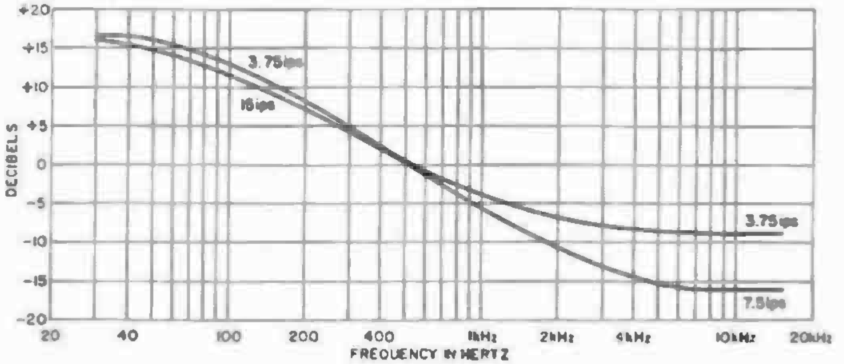


Fig. 17-173A. Typical post-equalization for ¼-inch tape recorders using NAB characteristic.

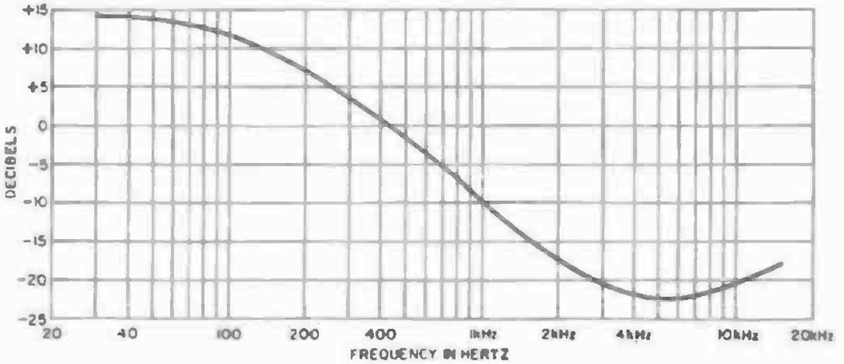


Fig. 17-173B. Typical post-equalization for ¼-inch tape recorders using Ampex Master Recording Equalization (AME) at 15 ips.

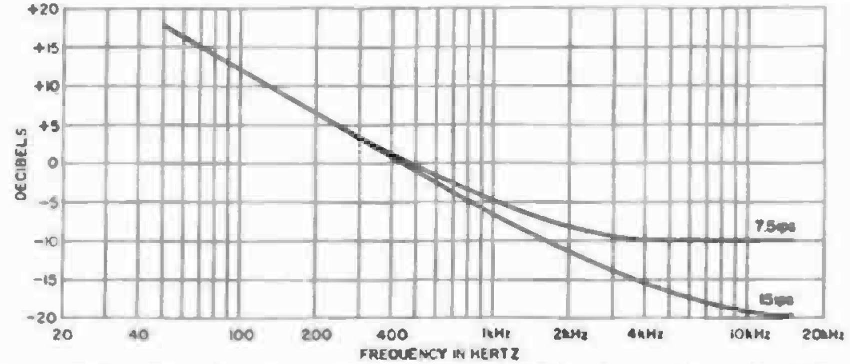


Fig. 17-173C. Typical post-equalization curves for ¼-inch recorders using CCIR characteristics, at 7.5 and 15 ips.

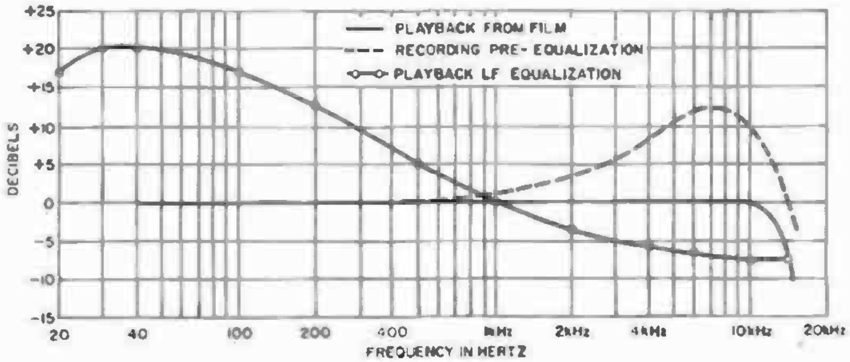


Fig. 17-174. Record-playback characteristics for a typical 16-mm magnetic film recorder running 36 fpm.

quency post-equalization is used, and is greater than that generally used with ¼-inch magnetic tape.

When recording dialogue, the low end is gradually rolled-off for reasons discussed in Questions 4.114, 6.122, and 18.81. Midrange high-frequency equalization also is used, when necessary.

17.176 Show a block diagram for a magnetic film production recording channel.—A block diagram of a typical two-position magnetic film recording channel for motion picture production is shown in Fig. 17-176. At the left are two inputs for microphones, followed by two dialogue equalizers. These equalizers provide low-frequency rolloffs for dialogue equalization, and a flat position for music. Following the equalizers is a two-position mixer network feeding a booster amplifier. A slating or talkback microphone is fed into one side of the mixer network for communication with the recordist and the boom man.

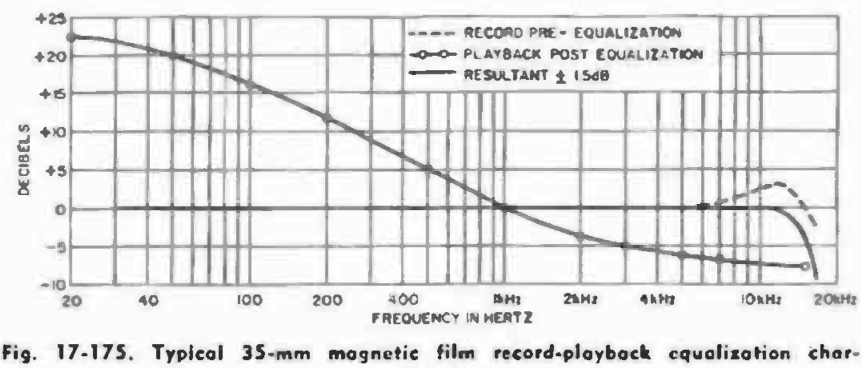


Fig. 17-175. Typical 35-mm magnetic film record-playback equalization characteristics.

At the output of the booster amplifier is an internal high frequency-low frequency equalizer for adjusting the recording characteristics to the particular linear speed of the recorder, 16, 17.5, or 35 mm. This equalizer network is in a negative-feedback loop, between the booster and recording amplifier, and when once adjusted is not readjusted. From the output of the recording am-

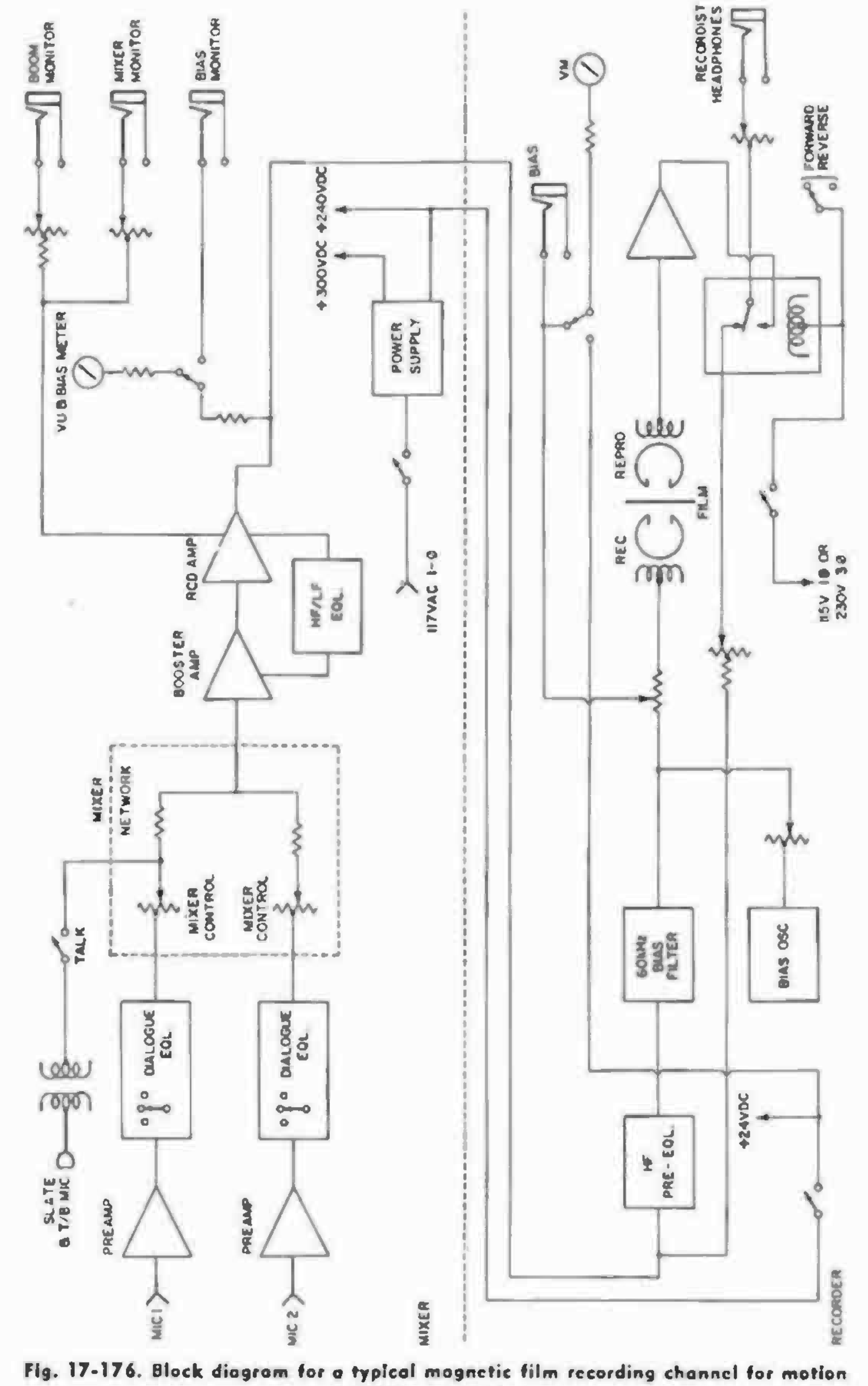


Fig. 17-176. Block diagram for a typical magnetic film recording channel for motion picture production.

plifier, the signal is applied to the recorder. The VU meter is connected at the output of the recording amplifier and is employed for reading both the bias current and the recording levels. Monitoring for the mixer and the boom man are also taken at this point through isolation pads for reducing the level to the recipients. From the output of the mixer, the signal is fed to a high-frequency pre-equalizer used for compensating the head characteristics of the recorder, and then to a 60-kHz bias filter used to prevent the bias-oscillator current from feeding back into the recording circuits. At this junction, the bias-current oscillator is also connected. After passing through the filter, the current is then applied to the recording head. A second meter is mounted on the recorder for monitoring the bias current and the 24-Vdc supply to the system.

After the signal has been recorded on the magnetic film, it is picked up by the reproduce head, amplified, and fed to a relay that automatically connects the output of the reproduce amplifier to the recordist's headphones. When the recorder is in motion, the relay connects the recordist's headphones to the output of the recording amplifier, thus providing direct monitoring. When the recorder is in operation, there is a small time delay between the direct sound and the recorded sound; however, the delay is of little consequence to the recordist, as he is primarily interested in the quality of the sound and not the action. If the mixer wishes, he may monitor directly from the film, but in this instance, the sound would be out of sync with the action. Switches on the recorder provide for reversing the direction of the film travel for rewind or playback. Equipment similar to the block diagram is pictured in Fig. 17-177A.

17.177 Describe the construction of portable magnetic film recording channels.—The Westrex Series 1200 portable magnetic film recording channel pictured in Fig. 17-177A consists of a mixer and magnetic recorder. The basic film-transport system is similar to that in Fig. 17-42B and is available for 16-, 17.5-, and 35-mm film operation. The film-transport system includes a timing belt, gear-reduction box, a 32-tooth stainless-steel film sprocket (35 mm), filter arms, fluid damping dashpot, and two impedance fly wheels.

The signal for the recorder is taken from the LRA-1592 mixer unit at the left. Also contained in the recorder case is a transistor bias oscillator, reproducing amplifier, bias-rejection filter, and a variable high-frequency pre-equalizer. Bias current is indicated on a meter mounted at the right of the front panel. The input and output impedance is 600 ohms. Frequency response is plus or minus 1 dB, 60 to 10,000 Hz. Signal-to-noise ratio is 60 dB for a plus 16 dBm signal at the output of the recording amplifier. The block diagram is quite similar to that of Fig. 17-176.

A portable magnetic film recording channel, Model M-2, manufactured by Amega, appears in Fig. 17-177B. The transport system employs a tight-loop

Fig. 17-177A. Westrex Series 1200 portable magnetic film recording channel, manufactured in Rome, Italy.

Fig. 17-177B. Amega Model M-2 portable magnetic film recording channel.

system using three motors, one for driving the film sprocket, and two torque motors for the supply and take-up reels, using a differential circuit to assure a tight wind. A ball bearing damping system is used to reduce flutter to less than 0.15 for 16-mm, and 0.10 for 35-mm film. The amplifier system is transistorized, using module-type construction which, if required, may be replaced from the front of the mixer case. The view in Fig. 17-177C shows the interior construction. The transport system may be obtained for 16-, 17.5-, or 35-mm film. Frequency response is plus-minus 2 dB, 45 to 10,000 Hz, with 0.10-percent total harmonic distortion from the film. Recording level is 10 dB below saturation. Signal-to-noise ratio at 0.10-percent distortion is 58 dB. Key switches at the lower portion of the amplifier case provide for reading the bias current, recording levels, monitoring direct or from film, dialogue and music equalization, and the selection of either the recording or playback modes.

A second magnetic film channel manufactured by Amega, pictured in Fig. 17-177D, is of particular interest, as it uses 16-mm magnetic film and is either internal battery or ac operated. The complete unit including the battery weighs 22 lbs, and may be carried by a shoulder strap. Synchronism with camera is achieved by the use of a sync-pulse generator at the camera. The 60-

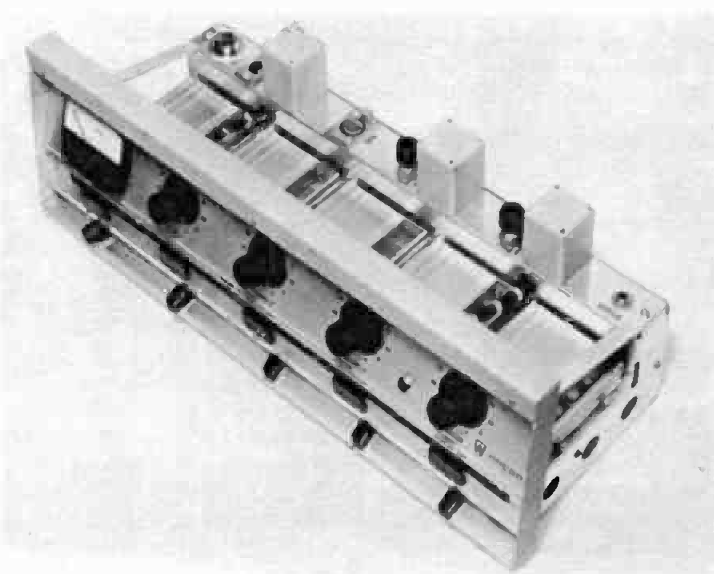


Fig. 17-177C. Interior view of Amego Model M-2 portable recording channel amplifier showing modular construction.

Hz sync pulse from the camera is amplified through an inverter and controls the speed of a salient-pole synchronous motor gear-coupled to the sprocket drive. A built-in 60-Hz frequency generator permits using the recorder for synchronous sound with no external power or frequency regulation. Torque motors are used for both take up and rewind. The audio facilities provide for a microphone input, bridging input, dialogue and music equalization, headphone jack, and monitor gain control. Switching is provided for record, playback, forward and rewind, and internal and external power operation. A simplified block diagram of its basic operation is shown in Fig. 17-177E. The width of the sound track is 200 mils. All electronics are solid-state. Frequency response is plus-minus 2 dB, 50 to 10,000 Hz, with less than 1-percent total harmonic distortion from the film. Flutter is less than 0.15 percent in any band, and 0.2 percent from 0 to 200 Hz (overall). Signal-to-noise ratio is 50 dB.

In Fig. 17-177F is shown a Model PM-64, 17.5-mm, 45-fpm magnetic film recorder manufactured by RCA. The various components are indicated and are typical of a magnetic film recorder used for shooting motion picture production. This unit is used in conjunction with a portable mixer containing the amplifying equipment. The various components indicated are: A, AA supply and take-up reels; B, BB guide rollers; C, CC guide rollers; D, DD feed and take-up sprockets; E, EE tight-loop rollers; F, FF magnetic head guide rollers; G, GG recording and playback heads; H head compartment cover; I bias current meter; J, JJ locking idlers; K footage counter; L carrying handle; M magnetic film; N external connection panel; O record-playback switch; P forward-reverse switch. The foregoing recorder may be operated from 117 Vac or from a camera motor supply system.

17.178 Describe the construction of studio recording equipment.—A group of magnetic film recorder-reproducer machines are illustrated in Fig. 17-178A, manufactured by Magna-Tech Electronics Co. These machines are designed for motion picture rerecording use, with several features that make these machines quite suitable for small or large studio operation. Starting at the left, the first four racks house dual machines that are mechanically and electrically independent of each other, and employ dual-film sprockets for running either 16/35-mm or 17.5/35-mm magnetic film. In the illustration, machines one and four are 16/35-mm, machine two is 35-mm single-track, machine three is 17/35-mm, and machine five is 35-mm three-track. The dual machines may be run straight synchronous, or selsyn interlock using a dual selsyn interlock distributor described in Question 3.56.

Fig. 17-177D. Amega Model 3-M portable 16-mm synchronous magnetic film recorder.

The distributor system may be controlled from the rack or from a remote position. When the speed is changed, proper equalization for the speed used is automatically inserted. With such combinations of linear speeds, sound tracks of different sizes and linear speed may be intermixed. If a 17.5-mm operation is required, the distributor is such that the machine runs at 45 fpm. An interlocking phasing circuit prevents leaving a stalled interlock motor on the distributor bus in an incorrect position, or on a point. The phasing control closes

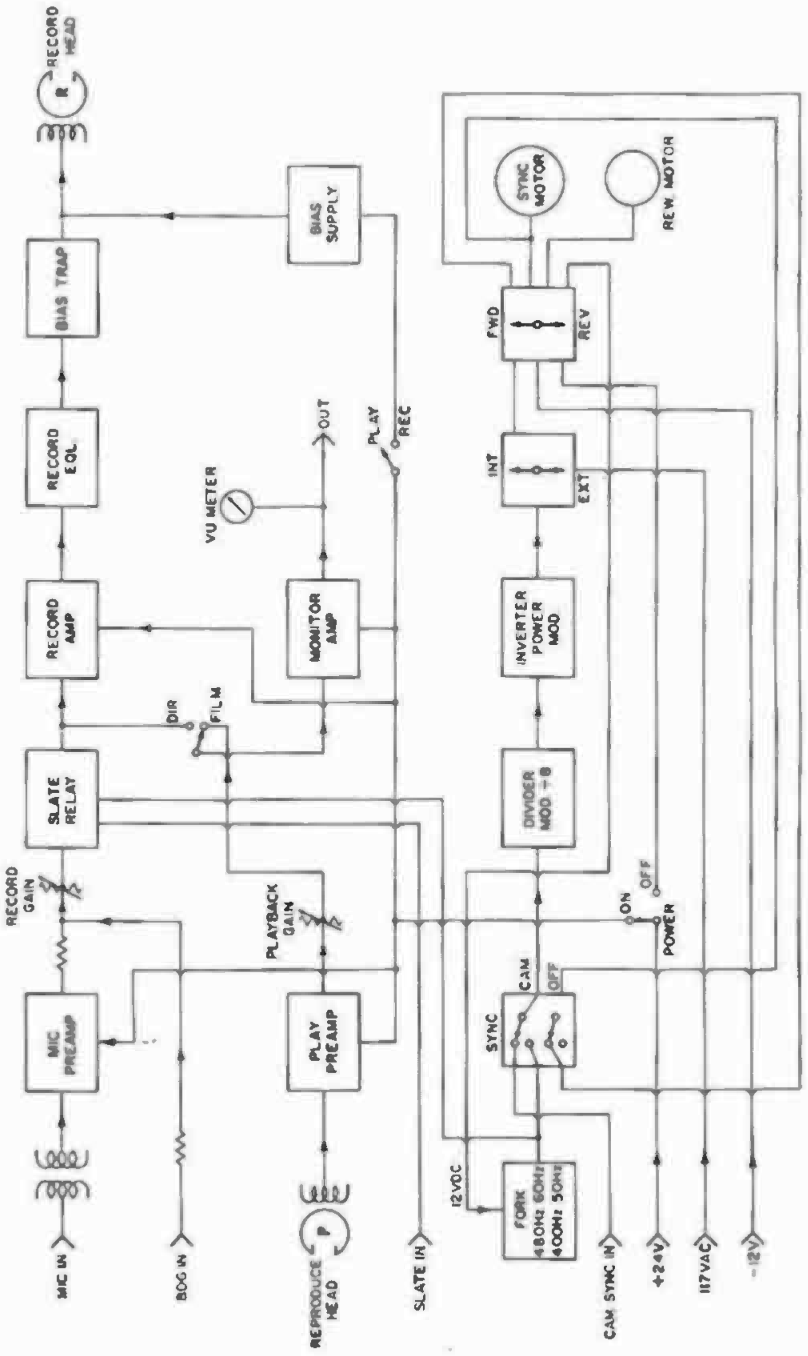


Fig. 17-177E. Block diagram for Amega Model 3-M portable 16-mm synchronous magnetic film recorder.

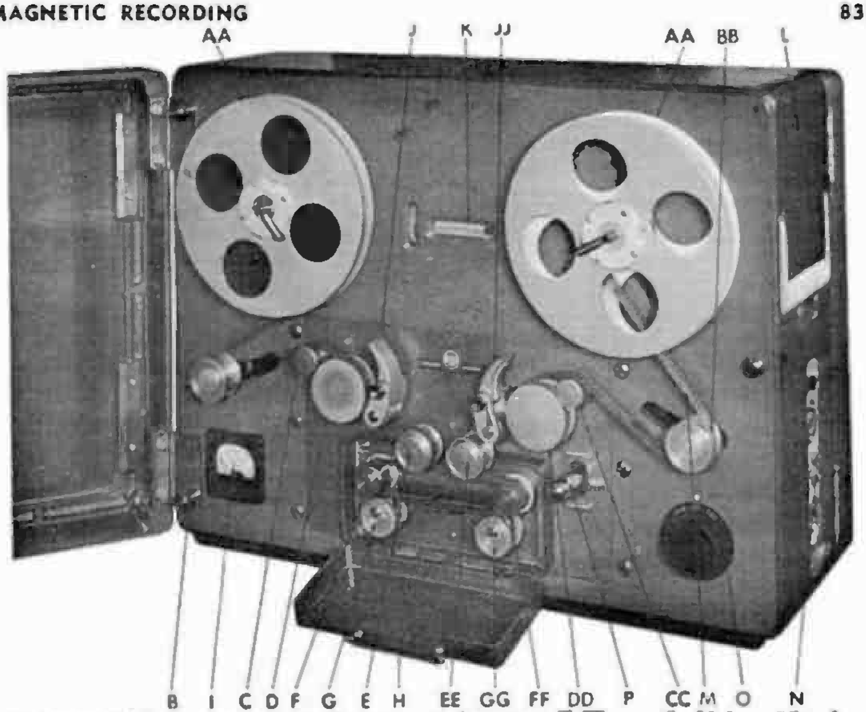


Fig. 17-177F. The RCA Model PM-64 portable 17.5-mm magnetic film recorder for motion picture production.

only after the rotor is correctly aligned. The feed and take-up reels are driven by torque motors which maintain the proper film tension at all times. An electrical braking system prevents the film from unspooling.

The machines are also capable of being reversed for rehearsing down in a reel while still maintaining sync, and for correcting mistakes during recording. The recorder is designed to go in and out of recording without introducing noise in dialogue or music; thus, the necessity of cutting the film into loops is eliminated. The recording and reproducing heads are prealigned and may be interchanged without realignment. Each head is surrounded by a Mumetal shield. The electronics are completely transistorized. The flutter for

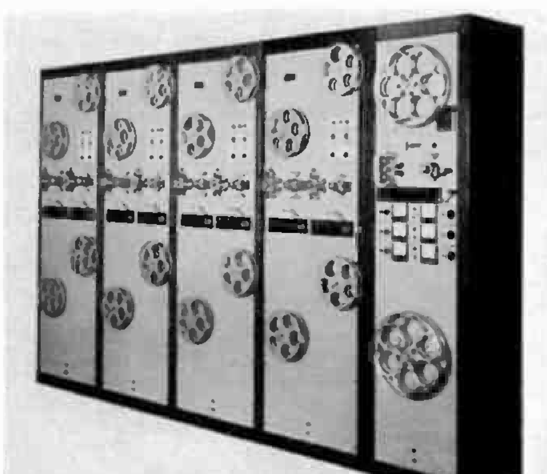


Fig. 17-178A. Combination recorder-reproducer equipment manufactured by Magna-Tech Electronic Co., Inc.

Fig. 17-178B. Westrex Model LRA-1552 recorder-reproducer manufactured in Rome, Italy. Machine records a single track on 35-mm magnetic film at 90 fpm.

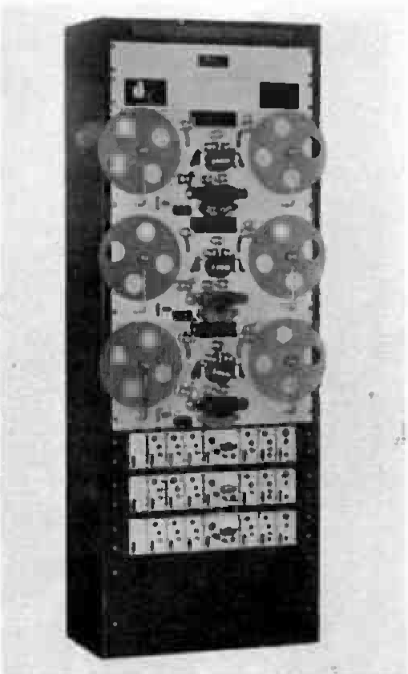


Fig. 17-178D. Westrex Model ST-510 multiple recorder-reproducer manufactured in Rome, Italy. Transistor amplifiers are mounted below the lower transport system.

16-mm equipment is 0.10 percent and 0.08 percent for 35 mm. The frequency response meets the standard of the industry, with a maximum total harmonic distortion of 0.5 percent. Signal-to-noise ratio is 65 dB below 100 percent modulation. The tracks are 200 mils in width; single tracks are recorded in the number 1 position.

A Westrex Model LRA-1552 35-mm single-track recorder-reproducer is illustrated in Fig. 17-178B. The top panel

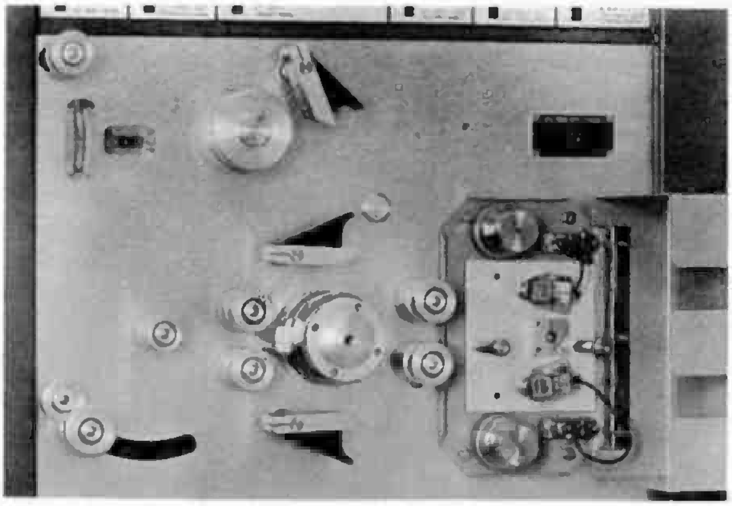


Fig. 17-178C. Close-up of Westrex Model LRA-1552 transport system and head mounts. The threading path is shown in Fig. 17-178B.

contains a torque motor-feed-reel spindle, feed roller and idler, fast forward stop and fast rewind controls, and a film-threading lamp which lights if the film is threaded plus-minus one sprocket from normal. The head assembly is prealigned and may be removed for cleaning and inspection and replaced without realigning. The frequency response, distortion, and signal-to-noise limits meet the industry standards.

A multiple recorder-reproducer, Model ST-510, also manufactured by Westrex, is shown in Fig. 17-178D. This machine incorporates in a single cabinet three reproducers, one of which may also be equipped for recording. The basic speeds may be 16-, 17.5-, or 35-mm magnetic film. A rear view is given in Fig. 17-178E, and shows the mechanical drive coupling system between the three transport systems. A single motor drives the three systems; however, a means of decoupling any one of the three units from the main drive is provided. Each transport system is designed as a complete unit, except for the common drive. If desired, the driving system may be modified for independent drive and control. The transport system uses the Davis tight-loop system, with a single film sprocket and dual impedance or fly wheels. A knob on the front panel of each film sprocket permits a particular transport system to be disconnected from the common drive shaft permitting the alignment of the sync mark without unthreading the film after the motor drive system has been interlocked. Torque motors and electromechanical brakes are employed for the rewind and take-up motors. Independent fast-forward and rewind are provided for also. The transport systems may be driven selsyn interlock or synchronous, depending on the local requirements.

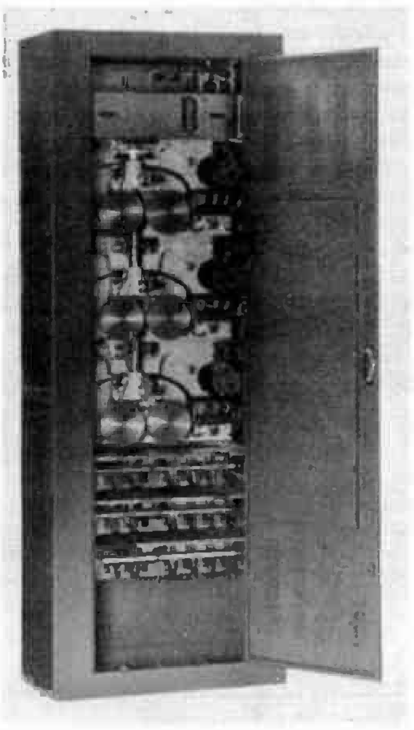


Fig. 17-178E. Rear view of Westrex Model ST-510 multiple recorder-reproducer.

Flutter is less than 0.08 percent at 90 fpm. Amplifier distortion is less than 1 percent at plus 12 dBm. The frequency response and distortion meets the industry standards. Signal-to-noise ratio is 65 dB re: 100-percent modulation. The amplifiers are of modular design, completely transistorized.

17.179 Describe a ¼-tape sync-pulse recorder for motion picture production use.—The recorder pictured in Fig. 17-179A is the well-known Nagara III ¼-inch magnetic tape recorder-reproducer, using the Neopilot (sync-pulse) system, manufactured by Kudelski of Lausanne, Switzerland. This recorder is used by many motion picture studios throughout the world and has many features applicable to motion picture production. Basically, the recorder is completely solid-state, with a ¼-inch tape transport system powered by a dc motor driven by internal batteries, although it may (where applicable) be operated from an external ac power-supply unit. The motor is of the permanent magnet electrodynamic type, resembling a d'Arsonval galvanometer, with a central magnet. Mounted on the motor shaft is a phonic wheel and the tape capstan. The phonic wheel passes in front of a tachometer head which is magnetized. The rotation of the phonic wheel over the magnetized head generates an ac voltage, the frequency of which depends on the speed of the motor shaft. The signal from the tachometer head is amplified, passed through a shaping amplifier, and results in a square wave of constant amplitude. This square wave operates a servo amplifier which, in turn, controls the speed

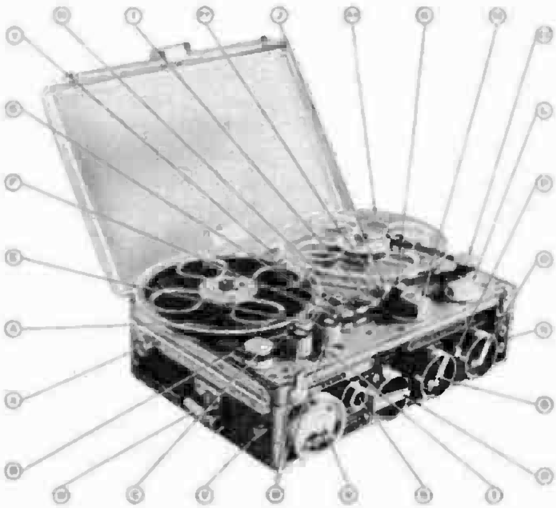


Fig. 17-179A. Nagra III ¼-inch magnetic tape recorder-reproducer, manufactured by Kudelski of Lausanne Switzerland. (Courtesy, Glen Glenn Sound Co.)

of the motor in conjunction with a discriminator-charging capacitor and a discharge diode.

When the recorder is used on motion picture production with cameras, a 60-Hz sync pulse is recorded on the tape, along with the audio signal to establish synchronism between camera and sound. The Neopilot signal (sync-pulse) may be taken from a pulse-generator on the camera, from the house mains, or from the ac voltage of a camera supply unit, as described in Question 3.88.

Referring to Fig. 17-179A, the tape-transport system starts at the supply reel A. The tape B leaves the reel and passes over tension pully C which is movable and supplies a braking action, thus applying a constant tension to the tape. The tape then passes over two flutter-filter rollers D and erase head E.

On the top surface of the flutter-filter rollers are stroboscopic discs for 50 and 60 Hz, one for each filter. The linear speed of the machine is checked by observing these discs under a 50- or 60-Hz light, depending on the particular frequency in use. The speed of the machine may be set accurately for either frequency by internal adjustment. The movement of one dot equal to the width of one dot at 50 Hz represents a change in speed of 1 percent. Variations in speed of 0.5 percent are quite common without the use of a sync-pulse signal.

Items F and FF are tape-alignment guide posts. The record head G and the Neopilot head (sync-pulse head) H are next, followed by the playback head I. At J is the capstan (on the motor shaft), with a neoprene pinch wheel at K. From this point the tape passes over another tension pulley CC and to the take-up reel AA. The controls consist of a playback-line input control O, a

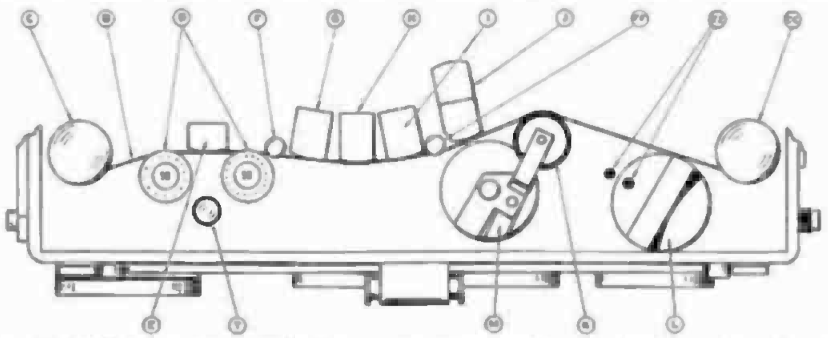


Fig. 17-179B. Plan view of the transport system and heads for Nagra III ¼-inch magnetic tape recorder, manufactured by Kudelski, Lausanne Switzerland.

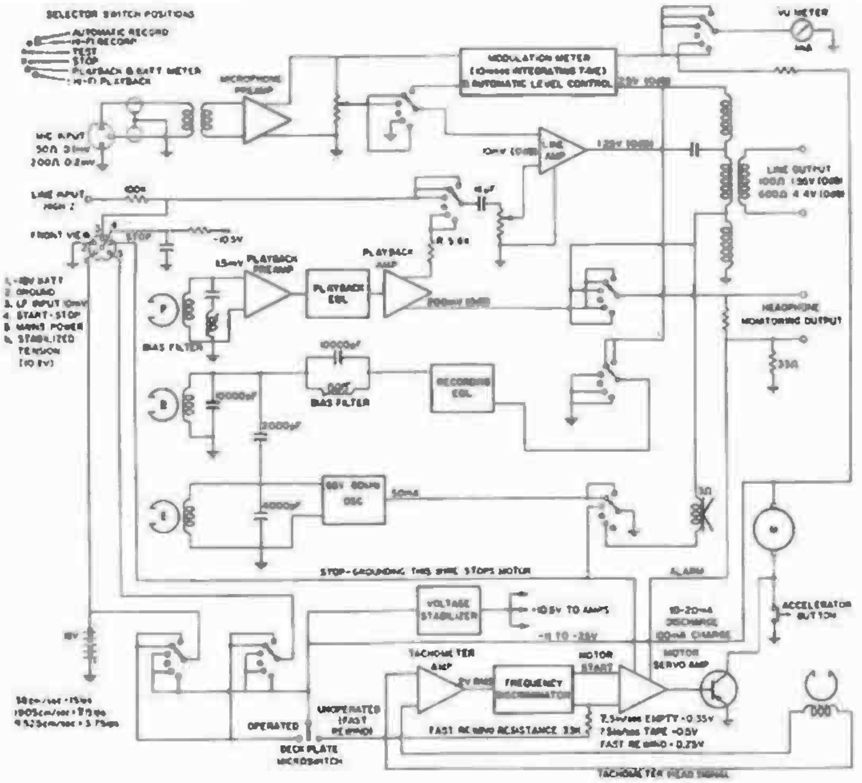


Fig. 17-179C. Block diagram for Nagra III ¼-inch magnetic tape recorder manufactured by Kudelski, Lausanne Switzerland.

microphone input level control Q and a function switch R. An oscillator push button (P) supplies a 3000-Hz signal for recording a reference tone at the head end of the roll for use later in transferring operations. This signal is recorded at the head end of each roll at a predetermined level and adhered to throughout the whole sequence of the sound department operations. Pin jacks are supplied at N for headphones. Returning to the function switch R, this switch has two sets of six positions—six for working from internal batteries and six for operating from an external power supply. The positions include: stop, test, hi-fi record, and automatic record, which provides automatic control of the recording level when using a microphone with attenuation for the low frequencies. However, the quality of the recorded signal in this latter position (according to the manufacturer) is not as good as when used in the hi-fi record position, and is used principally for dialogue recording, not music. This control affects only the microphone input circuit, and not the line input.

To indicate that the sync-pulse signal is present and being actually recorded, an indicator S, with a shutter, turns white when the signal is on and black when the signal is off. An accelerator button T permits the transport system to be speeded up for fast-forward tape movement during playback. VU meter U has two scales, one for measuring the battery voltage and a second for indicating the recording levels. The desired scale is selected by switch R. The VU meter has about a 6-dB lead. (See Question 17.163.) The thumb screw V is for attaching a shoulder strap; a second screw is placed on the opposite end (not shown). The microphone plug W employs a standard XL 3-pin Cannon plug. The input impedance may be set for either 50 or 200 ohms by changing an internal connection. As the machine has three speeds, a switch X provides the speed change and necessary equalization. The equalization meets the CCIR and NAB Standards for 15 and 7½ ips. For 3¾ ips only the NAB Standard is used, as this characteristic provides greater signal-to-noise ratio and less distortion at this speed, as compared to the CCIR.

Control L is a start-stop lever with an interlock push button (ZZ in Fig. 17-179B) for the transport system. Control K closes the pinch wheel against the motor capstan to provide tension between these two, and for pulling the tape over the heads. An internal loudspeaker Y provides a means for monitoring playback. A push button (not shown) allows the selection of direct or tape monitoring. At the rear of the case is a line input, output terminals, and a plug used for the ac pulse signal. At the right hand end (not shown) are connector plugs used for the operation of a resolver unit, described in Question 3.84, an external power supply, bridging input, and other connections.

For battery operation, twelve 1.5-volt flashlight batteries (or rechargeable batteries) are required. A plan drawing of the tape-transport system and heads is given in Fig. 17-179B, with the principal components indicated, using the same callouts for identification as used in Fig. 17-179A. The total weight of the recorder with batteries is about 16 lbs. A block diagram of its several sections appears in Fig. 17-179C.

The frequency response at 15 ips is plus-minus 1 dB, 30 to 18,000 Hz and at 7½ ips the same tolerance, 40 to 15,000 Hz. Harmonic distortion is 2 percent third harmonic, 0.5 percent second harmonic. Signal-to-noise ratios average minus 62 dB below the 100-percent recording level, using a weighting network described in Question 5.98. Erasure is 80 dB below 100-percent modulation. If the machine is to be used for playback on location, the tapes are prerecorded at the studio, including the sync pulse. Flutter measures 0.06 and 0.08 rms for speeds of 15 and 7½ ips, respectively. Operating environment temperature range is minus 20 to plus 50-degrees centigrade. Sync-pulse operation is discussed in Question 3.78.

17.180 Show a transistor plug-in type amplifier for use in recording systems.—An interior view of a 1-watt plug-in monitor amplifier, manufactured by Stancil-Hoffman, mounted on an octal base is pictured in Fig. 17-180A. A mixer-record-reproducer amplifier is shown in Fig. 17-180B. The amplifier combination provides for two microphone inputs, record-playback amplifiers, bias oscillator, and monitor amplifier all of plug-in transistor design. This amplifier is used in the machine pictured in Fig. 17-208A.

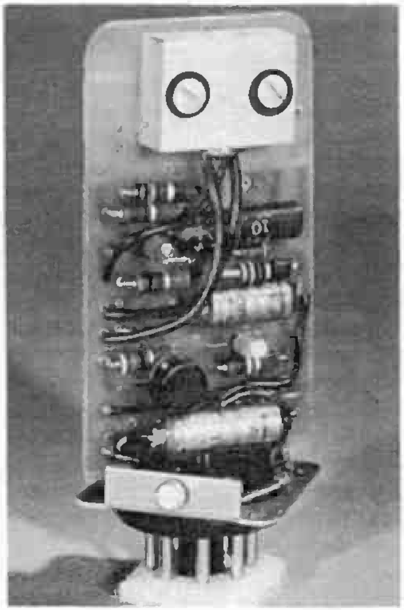


Fig. 17-180A. Interior view of Stancil-Hoffman 1-watt plug-in monitor amplifier Model AW70-1.

17.181 What is pre- and post-striped film?—Post-striped film is an existing motion picture film which has been processed, with a magnetic stripe added. The magnetic stripe may be placed on either the smooth or emulsion side, provided certain precautions are observed during the striping operations. Pre-striped raw stock is used in single-system cameras, similar to news reel, where the sound and camera are operated by one person. A single frame of post-striped film is given in Fig. 18-181. Not all types of photographic film can be pre-striped. Before prestriping any film, the processing laboratory should be consulted. (See Question 17.191.)

17.182 Describe a cassette used in magnetic recorders and reproducers.—A cassette is a removable housing containing a length of magnetic tape, generally with prerecorded program material. They are used in automobile tape reproducers, portable, and home reproducing equipment. Blank tape may also be inserted for recording purposes. The program material may be for monophonic or stereophonic reproduction.

Fig. 17-180B. Stancil-Hoffman Model ARP70 transistor amplifier system for mixing, recording, or reproducing either magnetic tape or film.

The term cassette is taken from the name of the container used with still cameras for holding film.

When a cassette is inserted in the recorder-reproducer, the magnetic head automatically makes contact with the tape and the driving mechanism. These devices are also used for continuous reproduction of program material. Standards and use of such devices are discussed in Question 17.230.

17.183 Give the standard for magnetically striping 8- and 16-mm motion picture film.—The USASI (ASA) Standard for magnetically striping 8- and 16-mm motion picture film is given in Figs. 17-183A to C. A balance stripe is generally placed on the edge without sprocket holes. (See Question 17.191.)

17.184 What is a half-striped magnetic track?—A film base on which there is an existing optical sound track. Half the optical track is covered with a magnetic stripe as shown in Fig. 17-184. The purpose of half-striping the optical sound track is to permit the film to be run on either an optical or a magnetic sound reproducer.

Half-striped sound tracks are generally confined to variable-density types. However, some work has been done using variable-area optical tracks of certain types, such as the double bilateral and duplex. (See Fig. 18-299A.) But, because of the size of the modulations and the fact that the stripe must be placed in the exact center of the modulations, striping variable-area

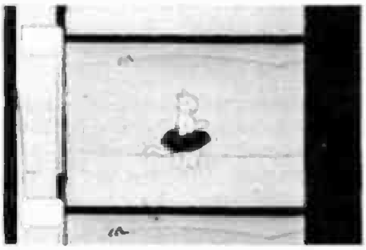


Fig. 17-181. 16-mm single perforated base with 100-mil magnetic stripe and a balance stripe on the opposite edge.

tracks has not met with too much success. Covering half the sound track will also increase the distortion.

Half-striping an optical sound track, regardless of the type, reduces the sound output by 6 dB, a loss of 50 percent in output level. In some instances, this may become serious because the signal-to-noise level is also reduced.

This system of magnetically striping motion picture film was first suggested by George Lewin of the US Army Pictorial Center.

17.185 What is a balance stripe?—A narrow magnetic strip placed on the film base on the opposite edge from the magnetic stripe. The balance stripe equalizes shrinkage of the film and thus prevents uneven winding and warpage of the film when wound or in storage. The balance stripe is applied at the same time the stripe for the sound track is applied and is the same thickness. A typical balance stripe is pictured in Fig.

17-181 and in the Standards in Figs. 17-183A and B. (See Question 17.191.)

17.186 For what purpose is full-coat magnetic film used?—Full-coat magnetic film is used where only the best recording quality is desired and is generally used for original music recording or the recording of a master composite sound track during a dubbing session. This applies equally well to 16-mm magnetic film. Although 17.5-mm magnetic film is also full-coat, it is, as a rule, confined to the recording of dialogue on production recording. Because the film is narrow and has only one line of sprocket holes, it has a tendency to buckle around the sprocket holes and induce flutter. The small amount of flutter induced as far as dialogue is concerned may be ignored. Recording equipment using 17.5-mm film must be carefully maintained as a slight amount of dirt on the film sprocket, idler rollers, or film tension can cause the film to jump the film sprocket quite easily. (See Questions 17.30 and 17.197.)

17.187 What is mag-optical track?—It is a sound track applied to a motion picture release print, using separate photograph and magnetic sound track. The term "mag-optical" is also applied to reproduction machines having facilities for reproducing both magnetic and photographic sound track.

American Standard Dimensions of

100-Mil Magnetic Striping on 16mm Motion-Picture Film Perforated One Edge

ASA
Reg. U.S. Pat. Off.
PH22.87–1966
Revision of
PH22.87-1958
*UDC 778.554.4

1. Scope

This standard specifies the location and dimensions of the magnetic striping material applied to 16mm motion-picture film with perforations along one edge. This film is used for both picture and sound.

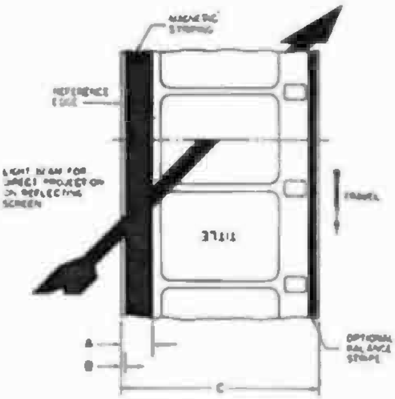
2. Magnetic Striping

2.1 Dimensions. The dimensions shall be as given in the figure and table.

2.2 Material. The magnetic striping material shall be on the side of the film toward the lamp on a projector arranged for direct projection on a reflection-type screen.

3. Film Base

The film base used shall be of the low-shrinkage safety type, cut and perforated in accordance with American Standard Dimensions for 16mm Motion-Picture Film, 1R-3000, PH22.12-1964, or American Standard Dimensions for 16mm Motion-Picture Film, 1R-2994, PH22.109-1965.



Dimensions	Inches		Millimeters	
A	0.100	+ 0.005 − 0.000	2.54	+ 0.13 − 0.00
B	0.005	max	0.13	max
C	0.628	nom	16	nom

NOTE: The balance stripe is optional and may be a magnetic coating or another material of the same thickness.

Appendix

(This Appendix is not a part of American Standard Dimensions of 100-Mil Magnetic Striping on 16mm Motion-Picture Film Perforated One Edge, PH22.87-1966, but is included to facilitate its use.)

The outer edge of the magnetic striping ideally should be coincident with the edge of the film, and for this reason, Dimension B as listed is based upon practical considerations of present striping techniques and film-handling mechanisms. Every effort should be made to reduce this dimension as far as possible, consistent with the best uniformity of stripe thickness and flatness of stripe profile.

Approved June 30, 1966, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers, Inc.

*Universal Decimal Classification

Fig. 17-183A. USASI (ASA) Standard for magnetically striping single perforated 16-mm motion picture film.

American Standard Dimensions of

Magnetic Striping of 16mm Motion-Picture Film, Perforated 2R-3000

ASA
Reg. U.S. Pat. Off.
PH22.101–1963
Revision of
PH22.101-1956

*UDC 778.534.425:771.526

1. Scope

This standard specifies the location and dimensions of the magnetic striping material applied to 16mm motion-picture film with perforations along both edges to be used for both picture and sound.

2. Magnetic Striping

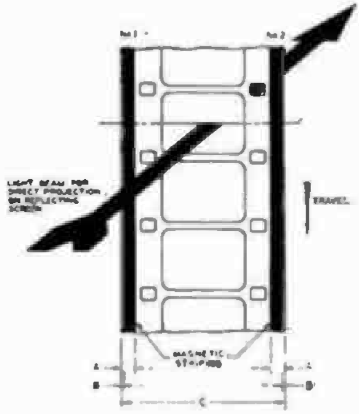
2.1 The location and dimensions of the magnetic striping shall be as given in the figure and table.

2.2 The magnetic striping is on the side of the film toward the lamp of a projector arranged for direct projection on a reflection type screen.

2.3 The No. 1 magnetic stripe is intended for the sound record.

3. Film Base

The film base used shall be of the low-shrinkage safety type, cut and perforated in accordance with American Standard Dimensions for 16mm Film, Perforated Two Edges, PH22.5-1953.



Dimensions	Inches		Millimeters	
A	0.031	max	0.79	max
	0.028	min	0.71	min
B	0.000	+ 0.005 − 0.000	0.00	+ 0.13 − 0.00
C	0.628	nom	15.95	nom

NOTE: The No. 2 stripe is an optional balance stripe and may be a magnetic coating or another material of the same thickness.

APPENDIX

(This Appendix is not a part of American Standard Dimensions of Magnetic Striping of 16mm Motion-Picture Film, Perforated 2R-3000, PH22.101–1963, but is included to facilitate its use.)

The outer edge of the magnetic striping ideally should be coincident with the edge of the film, and for this reason the dimension B is given as zero. The tolerance listed is based upon practical considerations of present striping techniques and film handling mechanisms. Every effort should be made to reduce this tolerance as far as possible, consistent with the best uniformity of stripe thickness and flatness of stripe profile.

Approved March 28, 1963, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers

* Universal Decimal Classification

Fig. 17-1838. USASI (ASA) Standard for magnetically striping double-perforated 16-mm motion picture film.

17.188 Describe an editorial mag-optical sound track.—In the early days of magnetic recording, a combination magnetic and optical striped film was used to aid the film editor in synchronizing the picture as the modulations of the magnetic sound track are invisible. The film base appeared as pictured in Fig. 17-188. The picture which is on a separate film base is placed under the clear portion of the striped film for synchronizing the action with the sound. This system of magnetic recording is now obsolete.

17.189 What is the standard for theater release prints, using four magnetic sound tracks?—The sound track placement is given in the USASI (ASA) Standard, Fig. 17-189.

17.190 Describe a magnetic sound track laminating machine.—Magnetic sound track laminating machines are used to apply a magnetic stripe to an existing optical print, requiring only a single sound track. In a method developed by Siemens and Halke of West Germany, the laminate consists of a polyester foil backing, manufactured by

AMERICAN STANDARD

Magnetic Coating of 8mm Motion-Picture Film

ASA

PH22.88-1956

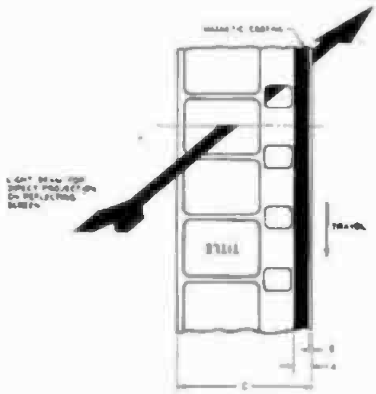
1. Scope

1.1 This standard specifies the location and dimensions of the magnetic coating material applied to 8mm motion-picture film to be used for both picture and sound.

2. Magnetic Coating

2.1 The location and dimensions of the magnetic coating shall be as given in the diagram and table.

2.2 The magnetic coating is on the side of the film toward the lamp of a projector arranged for direct projection on a reflection type screen.



Dimension	Inches	Millimeters
A	0.031 max 0.028 min	0.79 max 0.71 min
B	0.002 max	0.05 max
C	0.315 nom	8 nom

Approved April 24, 1956, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers

Price, 25 Cents

Fig. 17-183C. USASI (ASA) Standard for magnetically striping 8-mm motion picture film.

Agfa-Gevaert, which is preheated to approximately 120 degrees centigrade, then pressed on the 16- or 35-mm film base (Fig. 17-190). After the application, the polyester backing is removed from the magnetic striping material, leaving a smooth deposit of magnetic oxide. After laminating, the film is recorded in the usual manner. The trade name for this system is *Com-Mag*.

Fig. 17-184. A 16-mm half-striped variable-density optical sound track. The half-stripe permits either the optical or magnetic sound track to be reproduced.

17.191 Describe a portable liquid magnetic film striper.—A portable liquid type magnetic film striper is pictured in Fig. 17-191. This machine may be used for striping clear base 16-mm film, or applying a sound track to an existing picture. The machine shown is a Model P-16 liquid magnetic Magna-Striper, manufactured by Reeves Soundcraft Corp. Magnetic stripes may be applied in the standard positions for 16-mm magnetic film. Two types of liquids are available for striping, one for the emulsion side, and the other for striping on the base side. The magnetic emulsion is packaged in plastic bottles, which are mounted vertically on the striping mechanism at the right end of the machine.

Starting at the right of Fig. 17-191, the film is fed from a supply reel A to the striping mechanism consisting of two guide rollers B under a hopper head C supported by a sapphire pressure shoe D then over rollers E to film elevator F. Here, the film is looped 22 times from bottom to top of the elevator. Leaving the bottom of the elevator, the film passes to a capstan G and a pinch-roller H to the take-up reel I driven by spring belt J which is in turn driven by another belt by a motor in the base of the machine. The liquid magnetic emulsion is contained in the plastic bottle K with a shut-off valve L. Two sapphire-faced shoes are supplied, one for 25-mil striping, and one for 100-mil striping. The wider shoe may also be used for both half and full striping. When the machine is transported, the upper portion of the elevator arms are laid left and right. The operating environment is limited to a low value of 60 degrees Fahrenheit, with a relative humidity of 85 percent. If the magnetic stripe must be placed on the emulsion side of the film, special precautions must be taken during the striping operations. The striping equipment manufacturer should be consulted for this procedure.

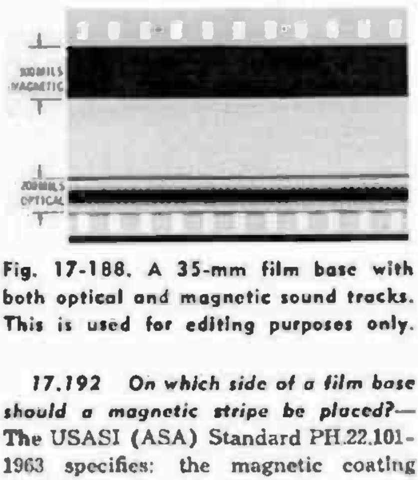


Fig. 17-188. A 35-mm film base with both optical and magnetic sound tracks. This is used for editing purposes only.

17.192 On which side of a film base should a magnetic stripe be placed?—The USASI (ASA) Standard PH.22.101-1963 specifies: the magnetic coating shall be on the side of the film toward the projector lamp of a projector designed for use with a reflective type screen. However, stripes are placed on both sides of the film, because of special considerations. If possible, the stripe should be placed on the smooth side of the base. If it must be placed on the emulsion side, special precautions must be taken in the striping operations. (See Question 17.191.)

17.193 Can unprocessed film be prestriped?—Yes; however, if the film base uses an antihalation coating on the cell side, the stripe may be loosened when passing through the film-processing tanks. The construction of the film base should be ascertained before striping. As a rule, most sound-recording stock contains an antihalation coating. (See Question 17-191.)

17.194 After applying a magnetic stripe to a silent film, how is the sound recorded?—It may be accomplished in two ways. First, the film may be threaded on a projector-recorder and while it is being watched by the narrator, his commentary is recorded. Music and sound effects may be added at the same time. In the second method the film is threaded on a regular magnetic film recorder and is recorded in the usual manner. (See Question 3.86.)

American Standard Dimensions of

Four-Track Magnetic Sound Records for 35mm Release Prints

ASA
Reg. U. S. Pat. Off.

PH22.137-1963

*UDC 778.534.425

1. Scope

1.1 This standard specifies the location and dimensions of the four magnetic sound stripes and of the recording heads to be used thereon for 35mm motion-picture prints.

1.2 This standard specifies the distance between the sound and corresponding picture.

2. Film Base

With the direction of travel as shown in the figure, the emulsion side of the film is up, the base is down, and the magnetic striping is on the base side.

3. Dimensions

3.1 The dimensions shall be as specified in the figure and table.

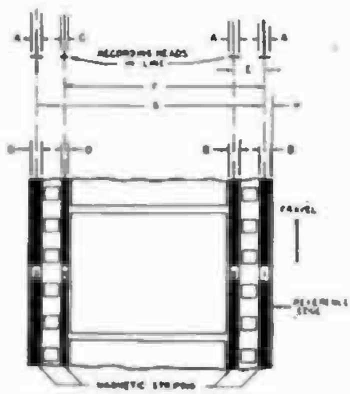
3.2 The cutting and perforating dimensions are specified by American Standard Dimensions for 35mm Motion-Picture Film, CS-1870 PH22.102-1956.

4. Sound Records

Track 1 shall be used for the left (as viewed from the auditorium) loudspeaker channel. Track 2 shall be used for the center loudspeaker channel. Track 3 shall be used for the right loudspeaker channel. Track 4 shall be used for the surround loudspeaker or control signals, or both.

5. Picture-Sound Separation

The sound record shall be separated on the film from the center of the corresponding picture by 28 frames ± ½ frame in such a direction that, in normal film motion, the picture first passes a given reference point and then the corresponding sound passes at a later time.



DRAWING SHOWS FILM AS SEEN FROM THE LIGHT SOURCE IN THE PROJECTOR

Dimensions	Inches	Millimeters
A	0.059 min	1.50 min
B	0.063 ± 0.003	1.60 ± 0.08
C	0.036 ± 0.002	0.91 ± 0.05
D	0.038 ± 0.003	0.97 ± 0.08
E	0.171 ± 0.002	4.34 ± 0.05
F	1.148 ± 0.002	29.16 ± 0.05
G	1.298 ± 0.002	32.97 ± 0.05
H	0.040 ± 0.002	1.02 ± 0.05

NOTE: The dimensions given in this standard are predicated on the use of unshrunk film. It is recognized, however, that in practice, one may encounter some shrinkage when striping a processed print. Specific measurements should take into account the overall width of the film as specified by Dimension A in American Standard PH22.102-1956. Should the film width fall outside the permissible tolerance, all dimensions specified in this standard may be multiplied by the ratio of nominal dimensions determined as follows:

$$\frac{\text{Measured width}}{\text{Specified width}} = \text{Ratio of nominal dimensions}$$

Approved December 10, 1963, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers

* Universal Decimal Classification

Printed in USA
ASA[illegible]

Fig. 17-189. USASI (ASA) Standard for release prints, using four magnetic sound tracks.

17.195 How are multiple magnetic sound tracks on release prints reproduced?—Magnetic sound tracks are used by the majority of the larger studios. Some studios employ four to six sound tracks on the release print, depending if it is 35 or 70 mm in width. For seven-track operation or more, a separate sound head is used running in interlock with the projector. This subject is discussed in more detail in Section 19.

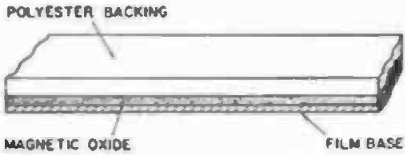


Fig. 17-190. Cross-sectional view of magnetic sound track laminate as used by Siemens and Halske for laminating optical film.

17.196 Describe a combination magnetic and engraved editing sound track.—A form of magnetic modulation writing used for making the modulations visible to facilitate editorial operations. A film base, as shown in Fig. 17-30B, is employed with a device which mechanically engraves a sound track by means of a sharp-pointed stylus. A typical sound track of this nature is shown in Fig. 17-196. In some instances the engraved sound track is played back similarly to a variable-area sound track for cuing purposes. As a rule, the quality is quite poor, but good enough to be used for cuing.

Other methods include writing the modulations with ink on the smaller stripe similar to an oscilloscope pattern or a variable-area optical sound track. This method is now obsolete.

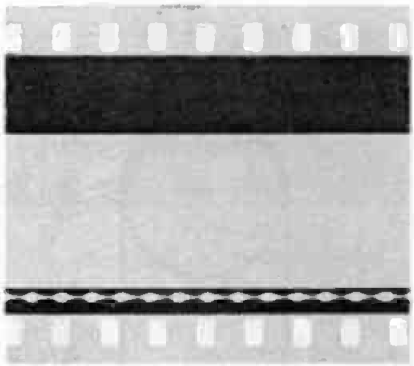


Fig. 17-196. A 35-magnetic film base with a mechanically engraved sound track, for editing purposes, in the 100-mil stripe.

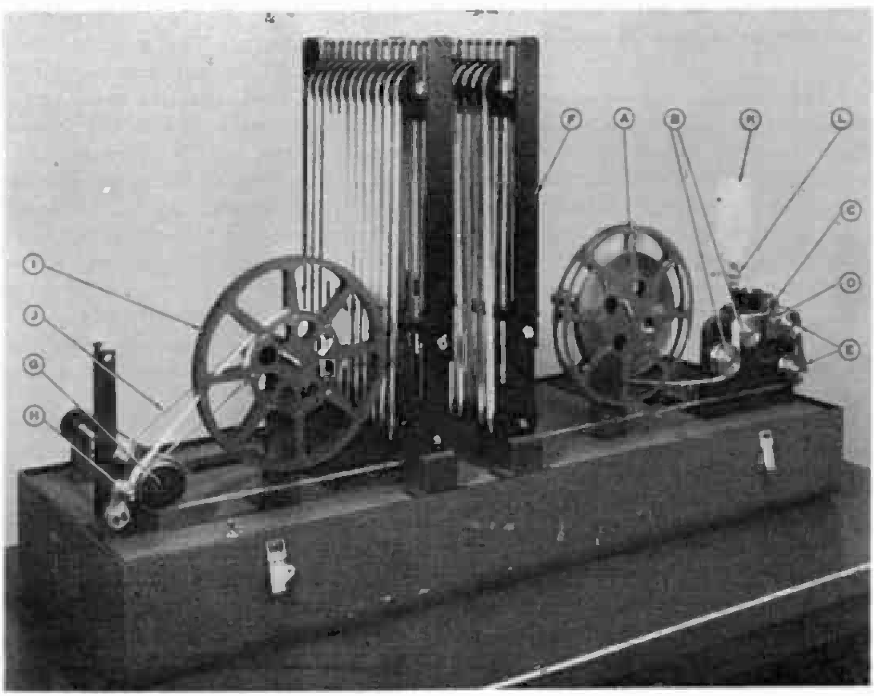


Fig. 17-191. Reeves Soundcraft Corp. Model P-16 16-mm Magna-Striper machine for applying a liquid magnetic oxide to motion picture film base.

17.197 Describe the polygoning effect on 16-mm magnetic film.—Magnetic striping or laminating double perforated optical film has several distinct disadvantages over striping or laminating single perforated film, among these are: double perforated film has only space for a 31-mil stripe, resulting in a low signal-to-noise ratio. Since the stripe is adjacent to the perforations, poor head contact occurs every time a perforation passes over the head. This is caused by the film being more flexible in the perforation area than between the perforations. This flexing is called polygoning and is illustrated in Fig. 17-197.

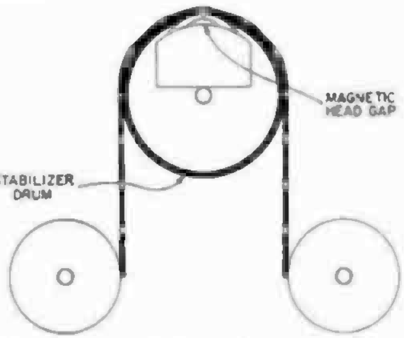


Fig. 17-197. The polygoning effect caused by the lifting of the magnetic stripe from the magnetic head as the sprocket holes pass by the head.

17.198 What is sequential magnetic recording?—The artist first records the melody using a three- or four-track tape recorder. He then records two or more harmony tracks, while listening to the melody track. As the tracks are all recorded and reproduced on the same machine, synchronization is maintained. If a master track is to be made, they are all rerecorded together as a composite track. Generally, the recorders used for this type recording are all especially designed for this operation, or attached to external control equipment. It is the usual practice for each recording activity to develop its own system and methods best suited for its equipment.

17.199 Describe fringing effect in a magnetic reproducing head and its cause.—When using a NAB Standard test tape or its equivalent recorded full-track and played back on a reproducer using a two-track, or a four-track, head, a rising characteristic is generally noted at the low frequencies around 100 Hz and is often mistaken for the tape characteristics. Two factors influence this rising response, fringing effects in the playback head and playback head bumps. Both these effects are the function of the wavelengths recorded on the tape and the linear speed of the tape. Fringing effects are caused by the magnetic flux on each side of the head, and under the head gap, contributing to the current induced in the head-coil winding, and, if great enough, appear as an increase in the signal output. Fringing effects are more noticeable at the lower frequencies and increase with a decrease of frequency because of the longer wavelengths. The magnitude of the fringing effect depends upon the head shielding, therefore a definite correction factor cannot be applied to fit all head designs. Tests made at 7½ ips show that different recorders will exhibit from 1- to 5-dB rise when full-track tapes are reproduced on narrow-track heads, although the recording has a uniform frequency characteristic.

Head bumps are a function of playback head geometry resulting when the pole pieces of the head and the gap pick up magnetic flux and transmit it to the head winding. As the frequency is decreased, bumps and dips in the response become apparent. The greatest bump occurs when one-half the wavelength of the recorded signal is equal to the distance between the pole pieces. Smaller bumps will be noticeable at 1.5 and 2.5 wavelengths and so on, with the largest dip occuring at 1 wavelength, and progressively decreasing at 2 and 3 wavelengths. Some manufacturers of tape recorders provide equalization to smooth out these unwanted effects. It is recommended by tape manufacturers that if full-width tapes are being used, and if low frequency equalization is provided in the playback circuits, it should be adjusted for a flat response at the low frequencies to overcome the effects discussed, since program material is also affected in the same manner. Standard test tapes recorded especially for quarter-track record-reproducers are available.

17.200 Why is 400 Hz used as a reference frequency for 16-mm magnetic film?—The ratio of film running at 90 fpm and 36 fpm is 2.5/1; therefore, 400 Hz running at 36 fpm is equivalent to

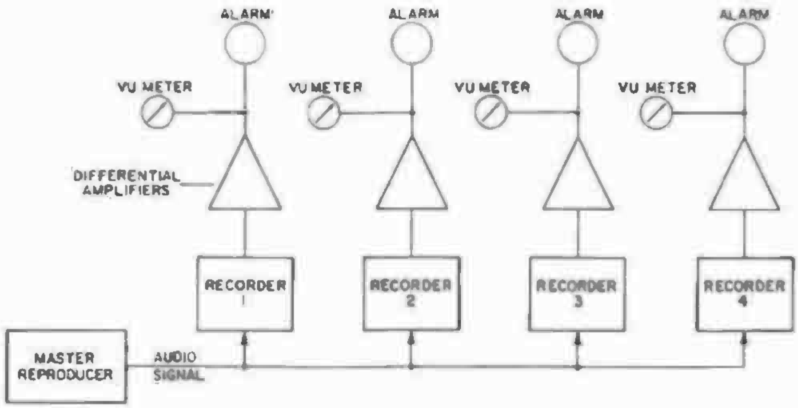


Fig. 17-202. Basic plan for recording four, 4-track release prints. At the output of each recorder is a differential amplifier and an alarm system.

1000 Hz running 90 fpm. For magnetic film, the Standard reference frequency for a linear speed of 36 fpm is 400 Hz, and for 35-mm running at 90 fpm it is 1000 Hz. These reference frequencies are also used for photographic film recording and reproduction.

17.201 Can a photocell be used for synchronizing a ¼-inch tape recorder with camera?—Yes, by the use of white marks along the edge of the tape. The pulse is picked up by the photocell and applied to a resolving device, then to the transfer recorder. This method was used by RCA before the advent of 60-Hz pulse-type synchronization. However, several systems have been since developed for synchronizing ¼-inch tape recorders to cameras and projectors. (See Questions 3.78 and 3.81.)

17.202 Describe how magnetic sound track release prints are recorded, and the overall level monitored.—After the final recording has been accomplished, several submasters are made from the dubbed master for recording of release prints. The frequency response of the submaster is held to within plus-minus 1 dB, with reference to 1000 Hz; however, a slight tilt-up of 2 to 3 dB is given low- and high-frequency ends to increase the signal-to-noise ratio at these frequencies. The submaster is placed on a special reproducing machine and its output applied simultaneously to the input of several magnetic film recorders on which are placed the release prints (Fig. 17-202). In this manner, each release print becomes an original recording.

To listen to each print individually after it has been recorded would be impractical; therefore, when the tracks are being transferred from the submaster, each sound track is monitored as it is being recorded, using a differential amplifier and a bridge circuit, which will sound an alarm if the level varies plus-minus 2 dB from the established level. A light over the machine indicates which track is at fault.

17.203 What precautions must be taken when splicing leader stock to magnetic film?—Leader stock is generally discarded picture stock and out-dated photographic unprocessed sound recording stock. Regardless of the type, the leader must always be spliced to magnetic film with the emulsion away from the magnetic head. If this precaution is not observed, the emulsion from the photographic film will be scraped off by the heads and pile up on the pole pieces and cause dropouts. If the magnetic film is full-coated, magnetically striped film may be used for leader stock; however, the sound track should be turned away from the head to prevent any noise left on the strip from entering the system and be thoroughly degaussed before use.

Two methods of splicing magnetic film are shown in Figs. 17-203A and B. Fig. 17-203A shows a solvent lap splice. In using this method, the splice must be perfectly degaussed to be free of noise. Fig. 17-203B shows a dry butt splice and is considered to be the best to eliminate interruption of the signal. If the splice is made correctly, it may be re-

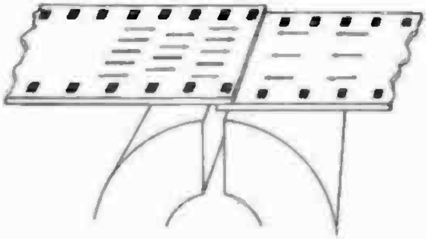


Fig. 17-203A. Butt-splice method of splicing magnetic film.

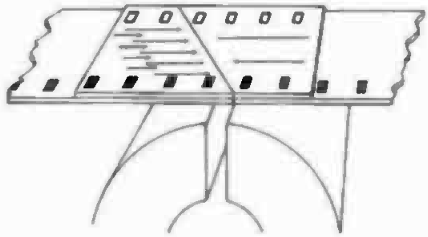


Fig. 17-203B. Diagonal-splice method of splicing magnetic film.

corded over without difficulty. However, if possible, splices should be avoided in a sound track. The arrow in the illustration shows the direction of the flux pattern. It will be noted that there is an abrupt change at the playback head gap when using a lap splice, and a gradual chauge when using a butt-splice.

Regardless of the splice used, all equipment employed in contact with magnetic film or tape must be degaussed. Splicing tape having 16- or 35-mm perforations is available from several sources of supply. Splices on magnetic film may also be made on a Bell and Howell diagonal splicer, as shown in Fig. 18-199B, provided the cutting knives are degaussed daily. This is a must because the splicing machine has an electric heater under the cutting blades, and if the heater is turned off, the knives are magnetically charged; thus, they must be degaussed. A degaussing tool similar to that shown in Fig. 17-90B may be used, except that it must be much more powerful. If the magnetic tape is run through an edge-numbering machine, any parts on this machine that might contribute noise must be degaussed also. The same precautions must be taken for ¼-inch tape.

17.204 ***When making measurements on a magnetic recorder, what precautions must be taken?***—Oscillators used for frequency and distortion measurements must not have a maximum of more than 0.25-percent total harmonic distortion (THD), and perferably less. The output of the playback amplifier must be terminated in a noninductive load resistance equal in value to the manufacturers recommendations. If the input of the recording amplifier is bridging, the oscillator output must also be terminated with a noninductive resistor equal to its output impedance. For frequency measurements, the levels must be set to avoid overloading the recording or playback amplifiers. It should be remembered when making distortion measurements, oscillator distortion cannot be subtracted directly from the measured distortion of the recorder. These and other subjects relative to measurements on magnetic recording equipment are discussed in Section 23.

17.205 ***How may the high-frequency response of a transformer-coupled magnetic playback head be altered?***—By the addition of a resistance in parallel with the output of the head and primary of the input transformer as shown in Fig. 17-205. The added resistance in parallel with the primary winding of the transformer varies the Q of the parallel resonant circuit formed by the transformer secondary and the input capacitance of the tube, thus affecting the high-frequency response.

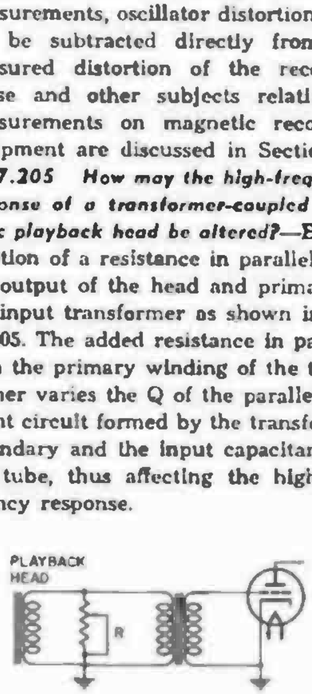


Fig. 17-205. A magnetic playback head shunted with a resistance to permit variation in the high-frequency response.

17.206 ***Describe an ideal magnetic reproducing system.***—According to the NAB Standard, an ideal magnetic reproducing system is a theoretical reproducing system, consisting of an ideal reproducing head, composed of a ferromagnetic ring in which losses are negligible. This means that the head gap is short and straight, the long wavelength flux paths so controlled that no low-frequency contours are present, and the losses in the head material are negligible. The system employs a reproducing amplifier, whose voltage conforms to the frequency response of Fig. 17-206 with

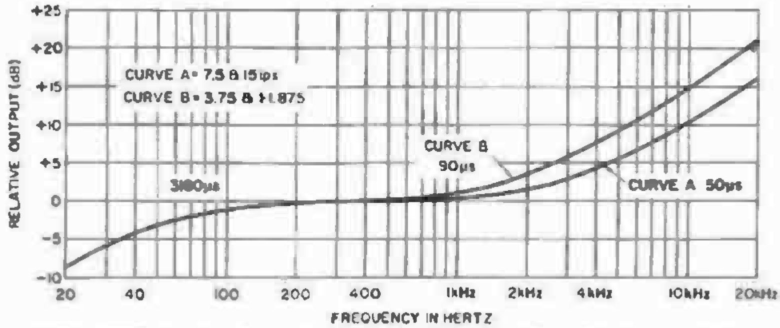


Fig. 17-206. NAB Standard reproducing characteristic. Reproducing amplifier output for a constant flux in the core of an ideal reproducing head.

a constant flux versus frequency in the head core. Because of several reasons, the flux in the core of an ideal head is not necessarily the same as the surface flux on the tape. Since most of the above effects are not easily measured, the NAB Standard is based on an ideal head-core flux, rather than surface induction. (See Question 17.207.)

The voltage versus frequency curve is to be uniform with frequency, except where modified by the equalization time-constant T_1 and T_2. The curve expressed in decibels is:

$$N\ \text{dB} = 20\ \text{Log}_{10}\ \omega T_1 \sqrt{\frac{1 + (\omega T_2)^2}{1 + (\omega T_1)^2}}$$

where,

ω equals $2\omega f$,
f is the frequency in Hz,
T_1 and T_2 are the time constants given below.

Tape Speed	T_1	T_2
15 ips	3180-μS	50-μS
7½ ips	3180-μS	50-μS
3¾ ips	3180-μS	90-μS
1⅞ ips	3180-μS	90-μS

The curve in Fig. 17-206 should not be confused with the pre-equalizer curves of Figs. 17-162A, B and C, as the curve in Fig. 17-206 is for a constant-flux at the reproducer head. The above time constants are necessary to attain the NAB Standard reproducing characteristic, shown in Fig. 17-206, the response for a constant flux using an ideal reproducing head.

17.207 Describe a primary-calibrated reproducing system.—Primary-calibrated reproducing systems are used for the purpose of calibrating standard magnetic test tapes and films. The system must not deviate by more than plus-minus 3 dB from the ideal over a frequency range of interest. The core loss of the reproducing head at the highest frequency of interest and the undamped resonance are not to exceed 3 dB, and the amplifier response should not vary from the ideal more than plus-minus 3 dB. Head-gap losses are not to exceed 3 dB at the highest frequency of interest, and the head-contour effects by not more than plus-minus 2 dB from the average.

Electrical characteristics of the system are made by measuring the frequency response of the amplifier and the reproduce system characteristic, with constant flux versus frequency in the head core. Magnetic losses can be determined by the calculation of gap losses and measurements of the head contour effects. After determining these losses, they are then considered as deviations from the ideal reproducing system.

The first measurement is of the amplifier response, using an input voltage proportional to frequency (the voltage doubles for each octave frequency increase) measured by the conventional methods. The second measurement is the head and amplifier response, made by applying the input signal across a low resistance connected in series with the head, as described in Question 23.52. A third measurement is then made of both the amplifier and head, using a constant flux versus frequency and the signal induced in the core of the head. This latter measurement is made by attaching a small wire firmly fixed over the head gap and feeding a constant-current signal through the wire.

Although the resultant flux distribution will not be identical to that of magnetic tape, it may be considered to be satisfactory for the purpose of measurement. The third measurement should follow the Standard reproducing characteristic given in Fig. 17-206. In actual practice, the curve will vary from the ideal because of the head resonance and losses. The effects of resonance and core losses of resonance are determined by comparing the results of measurements 1 and 2, while the apparent core losses are identified by measurements 2 and 3.

The approximate head-gap losses versus frequency may be calculated using the expression:

$$\text{Gap loss} = -20\,\text{Log}_{10}\frac{\text{Sin}\,[180^\circ\, d/\lambda]}{\pi d/\lambda}$$

where,

d is the null wavelength,
λ the wavelength of the frequency at which the gap is calculated.

Null wavelength is determined by finding the recorded wavelength at which the reproducing-head output voltage reaches a distinct minimum of at least 20 dB below maximum output. This measurement may be made using speeds of one-half and one-quarter the normal speed, using a tuned-voltmeter with no greater than one-third octave bandwidth. To reach a 20-dB null, the head gap edges must be sharp, straight, and parallel. To properly determine that the gap meets the requirements, visual examination using a tool makers microscope with a magnification of about 1000 times is necessary. It has been shown that the null wavelength is about 1.14 times the optical gap height for a perfect head. (See Question 17.67.) For the application being discussed the null wavelength should not be greater than 1.25 times the optical-gap height.

Low-frequency response is measured using a constant current versus frequency signal, and recorded using the normal bias current and the results compared with the curve obtained in measurement three. This will indicate the contour effects. This response should follow ideally the Standard reproducing characteristic in Fig. 17-206, for frequencies below 750 Hz at a speed of 7½ ips. In practice, at the long wavelengths all flux from a tape does not enter the head core, and the current that does enter varies with the wavelength and the length of tape contact at the head and also with the shape of the pole pieces and the shield around the head. It is of extreme importance that

Frequency	*Response*	*Frequency*	*Response*
20 Hz	−8.6 dB	1.5 kHz	+0.9 dB
25	7.0	2	1.45
30	5.8	2.5	2.1
40	4.1	3	2.75
50	3.0	4	4.1
60	2.3	5	5.4
70	1.8	6	6.6
75	1.6	7	7.7
80	1.4	7.5	8.2
90	1.2	8	8.6
100	1.0	9	9.5
150	0.45	10	10.35
200	0.2	11	11.1
250	0.1	12	11.8
300	−0.1	13	12.5
400	±0	14	13.1
500	+0.1	15	13.6
600	0.1	16	14.2
700	0.2	17	14.7
750	0.2	18	15.2
800	0.2	19	15.6
900	0.3	20	+16.1
1 kHz	+0.4 dB		

Fig. 17-207A. NAB Standard reproducing characteristic for 7½-and 15-ips tape speeds. Reproducer amplifier output for a constant flux in the core of an ideal reproducing head.

Frequency	Response	Frequency	Response
20 Hz	−8.8 dB	1.5 kHz	+2.2 dB
25	7.2	2	3.4
30	5.9	2.5	4.6
40	4.2	3	5.7
50	3.2	4	7.7
60	2.4	5	9.4
70	1.9	6	10.8
75	1.7	7	12.1
80	1.6	7.5	12.6
90	1.3	8	13.2
100	1.1	9	14.15
150	0.6	10	15.0
200	0.4	11	15.8
250	0.2	12	16.6
300	0.15	13	17.2
400	±0	14	17.9
500	+0.1	15	18.5
600	0.3	16	19.0
700	0.5	17	19.6
750	0.55	18	20.0
800	0.6	19	20.5
900	0.8	20	+21.0
1 kHz	+1.0 dB		

Fig. 17-207B. NAB Standard reproducing characteristic for 1⅞- and 3¾-ips tape speeds. Reproducer amplifier output for a constant flux in the core of an ideal reproducing head.

the distortion of the oscillator be on the order of 0.10 percent or less. The frequency of the oscillator must be accurately set for each measurement to assure frequency errors are not interpreted as response-curve errors. The slope of the contour effect curve must not exceed 10 dB per octave, so that a frequency error of 0.5 percent will result in a response error of not more than 0.07 dB.

After having once determined the various losses or deviations from the ideal system response, a calibration of the actual system may be made. From the system response measurement 3, subtract the gap-loss curve at the high frequencies and algebraically add the low-frequency portion by the contour effect curve. The resulting curve is the reproducing system response for constant flux from the tape. The difference between this curve and the Standard reproducing system characteristic represents the deviation from the ideal response.

The response for a constant flux in the core of an ideal reproducing head is given in Figs. 17-207A and B covering linear speeds of 1⅞, 3¾, 7½ and 15 ips.

17.208 Describe a loop box.—In rerecording of motion pictures (dubbing), many times it is necessary to have a continuous background sound of rain, crowd noises, traffic, and many other types of noise. To achieve this continuous sound, the particular effect concerned is recorded on a magnetic film 50 to 100 feet in length, and then spliced into a continuous loop. This loop is then threaded on a reproducer and the balance of the film placed in a loop box, where it is automatically fed through the machine as a continuous sound effect. A facility of this type is shown in Fig. 17-208A, mounted on a Stancil-Hoffman S7/ARP-70 35-mm recorder-reproducer. In this particular type loop box, the film is fed from the right end of the transport system A to an idler B, and then to the left side of the box to the right side of sprocket C, which feeds the film D downward into the box. The film returning from the box is pulled upward on the left side of sprocket C, and fed to the transport system again. The film is separated, guided, and prevented from rubbing by webbing, which also applies the proper tension to the film as it enters and leaves the box. In this manner, about 150 feet of continuous sound track may be fed through the reproducer. The left side of the plastic front is cut out to facilitate threading.

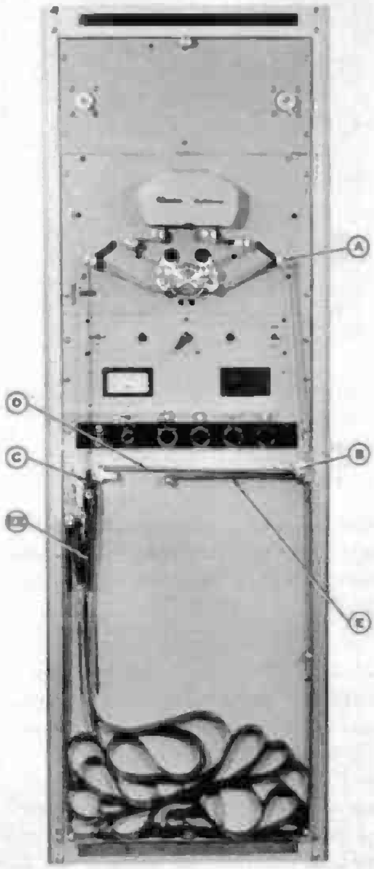


Fig. 17-208A. Stancil-Hoffman Corp. Model S7/ARP-70 magnetic film recorder-reproducer fitted with loop box.

Three loop boxes, mounted on three reproducers of a group of twelve, are shown in Fig. 17-208B. These are manufactured by RCA. Although the appearance of the loop box is similar to that of the first box described, its method of feeding and returning the film to the box is quite different. In this box mechanism, the film is pulled downward by a pinch-wheel arrangement attached to the right take-up spindel. This feeds the film to an idler roller and then back into the box. As the film leaves the box, it is pulled upward by the transport system. The lower portion of the box may be opened for inserting or removing the film. The box may be removed

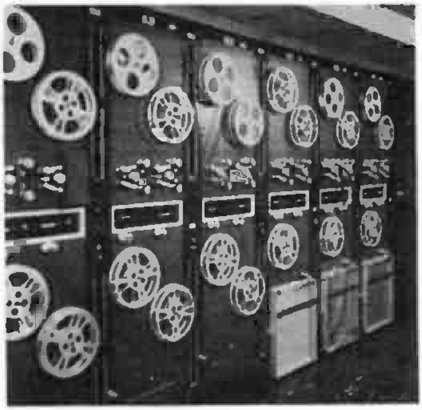


Fig. 17-208B. Six RCA PM-75 dual 35-mm reproducers. Last three machines are equipped with loop boxes.

from the rack by pulling two hinge pins.

The group of machines shown are twleve RCA PM-75 35-mm dual magnetic film reproducers. Each machine is mechanically and electrically independent of each other. All six racks are the same, except for rack 1. Here, the first machine is a single-track reproducer, while the second is a combination single- and three-track reproducer. Each machine is equipped with plug-in preamplifiers housed behind a drop-down door at the bottom of the rack. All machines are selsyn interlock driven, except rack 1. In this rack, dual purpose motors are employed which may be run selsyn or straight synchronous. Torque motors are used for the feed and take-up reels. The motor system is completely reversible. Intercommunication and selsyn interlock controls are placed between each rack for convenience of operation.

17.209 ***What is a magnetic sound-track scriber?***—A device used for inscribing an ink tracing on magnetic sound tracks for editing purposes. Referring to the block diagram in Fig. 17-209A, the device consists of a magnetic tape or film-transport system, a magnetic playback head, a high-frequency loudspeaker unit on which is mounted a special ball pen for scribing the modulations of the sound track on the magnetic tape, an amplifier system, a rectifier for supplying a fixed dc bias voltage to the voice coil of the loudspeaker unit, and a rectifier amplifier which rectifies the signal voltage and

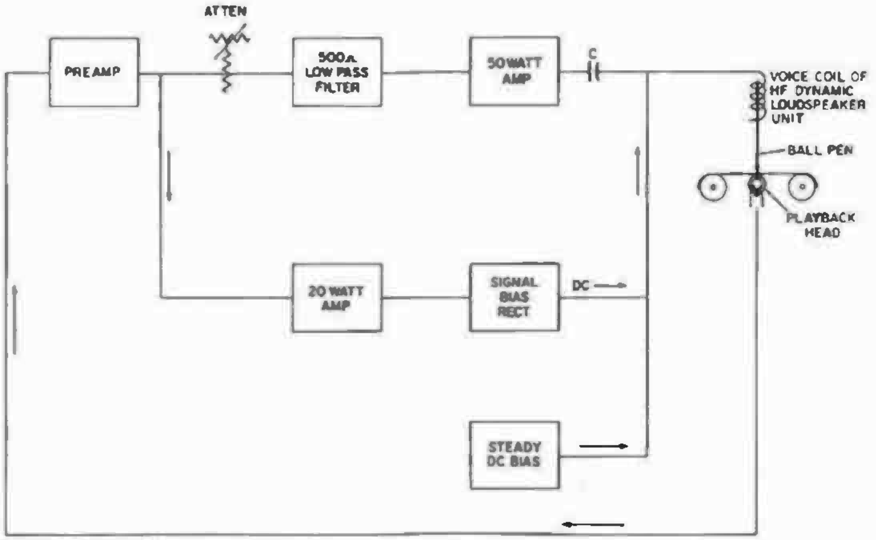


Fig. 17-209A. Block diagram of a magnetic sound track scriber using a ball pen for scribing the sound track on magnetic tape or film.

is used as a variable bias for cancelling a part of the steady dc bias voltage on the speaker unit voice coil.

As may be seen from the drawing, the loudspeaker unit is mounted mechanically in such a manner that the ball pen will bear on the slick side of the magnetic film or tape. The ball pen is directly over the gap in the magnetic pickup head. The signal from the magnetic sound track to be scribed is fed to a preamplifier, then to an attenuator, a 500-Hz low-pass filter, and a 50-watt power amplifier. The loudspeaker unit is coupled to the output of the power amplifier through a large capacitor C. Bridged across the output of the preamplifier is a 20-watt power amplifier which feeds the signal to the bias-rectifier amplifier, whose dc output is connected to the voice coil of the loudspeaker unit. The second rectifier supplies a steady dc voltage to the loudspeaker voice coil, deflecting it to one side during periods of no modulation, inscribing a bias line.

When a signal appears at the input of the preamplifier, the voice coil of the loudspeaker is actuated and the pen inscribes a wavy line following the modulation envelope of the signal. At the same instant, the signal is rectified and the steady bias on the speaker voice coil is cancelled, permitting the pen to scribe the modulations. The 500-Hz low-pass filter limits the frequency response; thus, only the envelope shape of the modulations is scribed on the tape.

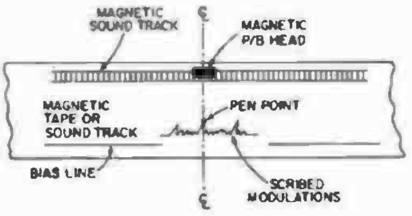


Fig. 17-209B. A magnetic sound track scriber playback head and pen mounted on top of the magnetic film.

An alternate method to the mechanical mounting of Fig. 17-209A is to mount the pen on the top side of the tape across from the pickup head as shown in Fig. 17-209B. This device was developed by the Sound Department of Paramount Productions, Inc., Hollywood, California.

17.210 What is a constant-current recording characteristic of a magnetic recording head?—A frequency characteristic of the head made by applying a constant current through the head, rather than maintaining a constant voltage at the input of the recording amplifier.

The measurement is made using a recording amplifier with a flat frequency response rather than the normal equalization used for recording magnetic sound tracks. A 10-ohm resistor is connected in series with the recording head as shown in Fig. 17-210 and the current is held constant at each frequency of interest by reading the voltage drop across the 10-ohm resistor.

The recorded tape or film is played back using an unequalized playback amplifier but with flat frequency characteristics. The characteristic when plotted should show a 6-dB rise per octave.

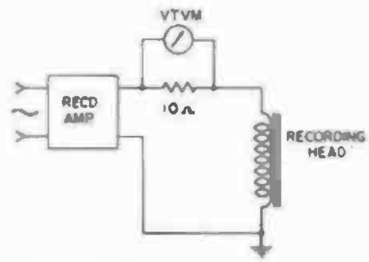


Fig. 17-210. Circuit for measuring the constant-current recording characteristics of a magnetic recording head.

17.211 What is a voltage-recording characteristic of a magnetic recording head?—A measurement of the voltage developed across the head winding with respect to frequency. The measurement is made by connecting a vacuum-tube voltmeter across the head winding and reading the developed voltage for each frequency of interest as shown in Fig. 17-211A.

The test frequencies applied to the input of the recording amplifier are held at a constant amplitude. A typical frequency response, made in the above manner, for a recorder of limited-frequency response is shown in Fig. 17-211B.

17.212 What is a magnetic reader?—A hand-operated device for the purpose of finding the modulations of a magnetic sound track. In Fig. 17-212 is shown such a device manufactured by the Hollywood Film Co.

The magnetic film is pulled by hand or by rewinds across the large drum in the center, making contact with a magnetic reproducing head. The case contains an amplifier and small loudspeaker. The particular unit illustrated will reproduce both photographic and magnetic sound tracks. Such an instrument is quite valuable in finding noisy spots in photographic sound tracks and in locating the exact start or finish of a modulation in a magnetic sound track. The principal components consist of a case A, exciter lamp and housing for optical sound tracks B, sound optical system C, guide rollers D and J, photocell housing E, sound drum F, magnetic reproducing head G, magnetic head lifting arm H, amplifier volume control I, loudspeaker K, and line voltage switch L.

17.213 What is a magnetic synchronizer unit?—A motion picture film-foot-

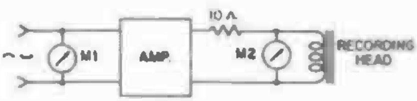


Fig. 17-211A. Circuit for measuring the voltage characteristics of a magnetic recording head.

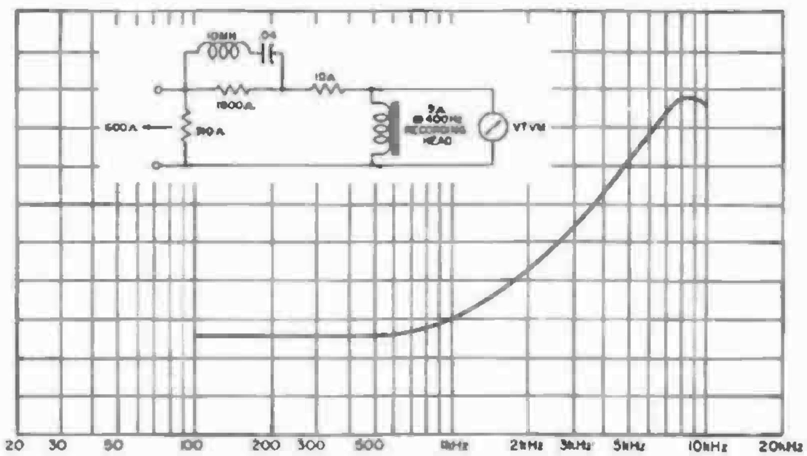


Fig. 17-211B. Voltage response across a magnetic recording head using normal equalization and a constant input to the recording amplifier.

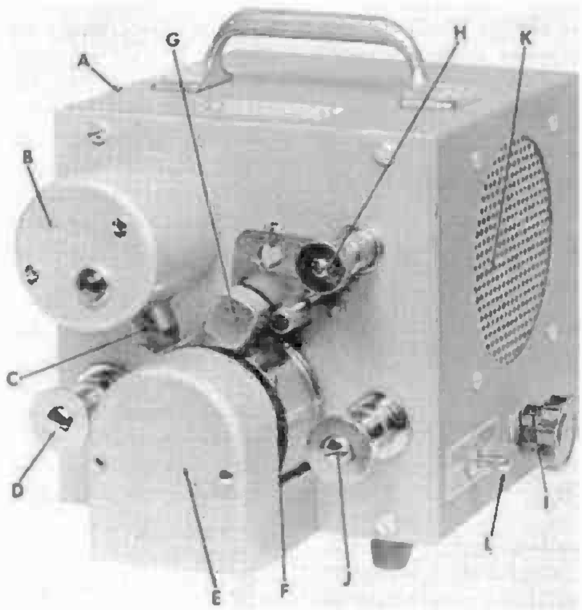


Fig. 17-212. A sound reader manufactured by the Hollywood Film Co. The model illustrated will reproduce both magnetic and photographic sound tracks.

age counter fitted with a magnetic reproducing head as shown in Fig. 17-213. The output of the magnetic head is fed into the amplifier of a magnetic reader such as that illustrated in Fig. 17-212, or to a separate amplifier.

17.214 Show a film splicer suitable for magnetic film.—A magnetic film splicer manufactured by the Hollywood Film Co. is shown in Fig. 17-214. The particular unit shown will splice either 17.5-mm or 35-mm film, using a Mylar base splicing tape. The pins projecting upward from the base hold the film in its proper place while the cutting knife cuts off the surplus film. The splicing tape is then applied over the splice as shown. This type splice eliminates pops when the spliced material is passed over the sound head. The knife and jaws are nonmagnetic.

17.215 How may tapes recorded at a given speed be rerecorded to play back at a different speed?—If two magnetic

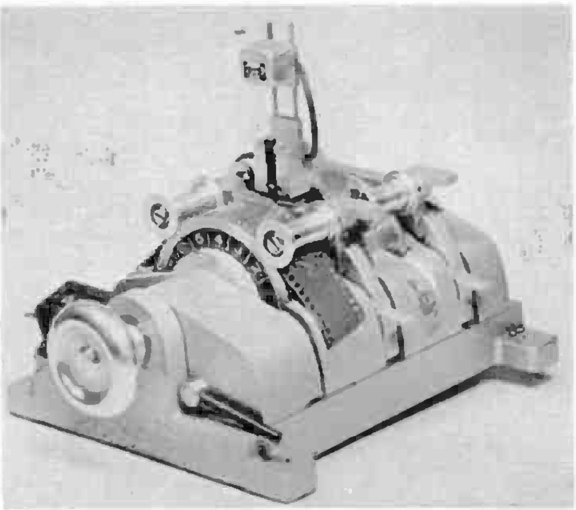


Fig. 17-213. A film footage counter equipped with a magnetic sound head. (Courtesy Hollywood Film Co.)

recorders having speeds of 7.5 and 15 inches per second are available, tapes may be rerecorded to play back at the speeds indicated below.

Original Speed	*Repro-duce*	*Re-record*	*Back Play*
3¾	7.5	15	7.5
7.5	15	7.5	3¾

If a recorder is available with a speed of 30 inches per second, the following combinations may be obtained.

Original Speed	*Repro-duce*	*Re-record*	*Back Play*
15	30	15	7.5
15	30	7.5	3¾

When the foregoing are rerecorded, equalization is introduced to compensate for the lower linear speeds used in the final reproduction.

17.216 ***What test tapes and films are recommended for testing ¼-inch tape and magnetic film recorders?***— Standard ¼-inch test tapes are available from the National Association of Broadcasters (NAB) and several different manufacturers. These tapes include an azimuth adjustment, standard output level, and a group of frequencies for making response measurements. Magnetic test films for 8-, 16-, 35- and 70-mm recording and reproducing equipment are available from the Society of Motion Picture and Television Engineers (SMPTE). Such films include 3-track balancing, 3-track azimuth, 3-track 3000-Hz flutter, multifrequencies for response measurements, and listening tests including music, dialogue, and many others. Similar test films are also available for photographic sound reproduction (optical film recording) for both studio and theater projection equipment.

17.217 ***Describe in injection frequency measurement.***—An injection frequency measurement is made of a playback circuit, with the playback head connected in the circuit as normally

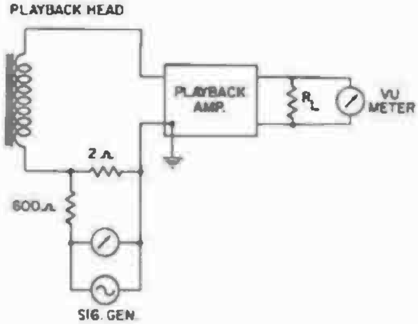


Fig. 17-217. An injection circuit for measuring the frequency response of a magnetic reproducer amplifier.

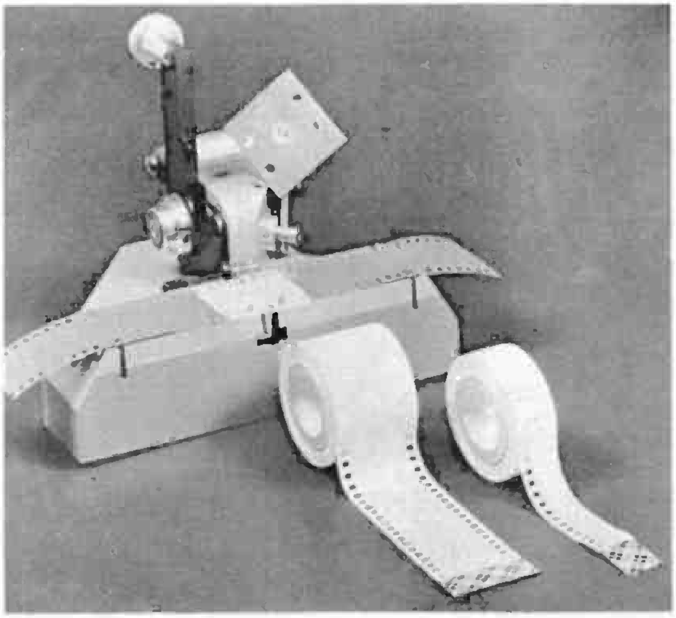


Fig. 17-214. A magnetic film splicer manufactured by the Hollywood Film Co. with "scotch" magnetic film splicing tape.

used. Such measurements provide a measure of the frequency response under impedance conditions that normally prevail with the playback head in the circuit.

The frequency response is measured by connecting a 2-ohm noninductive resistor in series with the reproducing head as shown in Fig. 17-217. A constant voltage is maintained across the 2-ohm resistor and the frequency response measured at the output of the amplifier system. Adjustment of the amplifier equalizer circuits are made to attain the desired playback frequency response.

After the amplifier frequency response has been adjusted using the injection method of measurement, a final check is made by reconnecting the reproducer head and the frequency response measured using a multifrequency tape or film. (See Question 23.52.)

17.218 In what type environment should magnetic tape and film be stored? —Both magnetic tape and film should be stored at a temperature ranging between 60 and 80 degrees Fahrenheit, with a relative humidity of 40 to 60 percent.

17.219 Describe the electronics for a transistor magnetic recorder.—Basically, magnetic recorders employing transistors in the electronics of the recorder are much the same as its forerunner, the vacuum-tube type, and except for a few minor changes, the controls remain about the same.

As an example of transistor engineering in a high-quality professional magnetic tape recorder, the schematic diagram for the record-playback circuitry of an Ampex Model AG-300 one-quarter or half-inch tape recorder is shown in Fig. 17-219A. Because there is considerable difference between the reproduce and the recording circuitry, the recording circuitry will be described first, by use of the block diagram (Fig. 17-219B) in conjunction with the schematic diagram of Fig. 17-219A.

The input signal is applied to input plug 4J6, then to the input selector switch 4S3, which selects either a bridging or lower input impedance. From the input transformer, the signal goes through record level control 2R38, to the base of emitter follower stage 1Q10. From this point, the signal path splits, one path leading through record calibrating amplifier 1Q11, where gain is adjusted by record-calibrate control 2R45, through the contacts of output selector switch 2S1, to line amplifier 1Q7, and the VU meter.

The second signal path from 1Q10 is passed to the base of amplifier 1Q12. The record equalization consists of a variable capacitor contained in the plug-in equalizers of Fig. 17-219D and is selected by means of a relay. This capacitor is connected in parallel with resistor 1R42 to provide the necessary high frequency pre-equalization. After amplification by 1Q12, the signal is connected to 1Q13 and 1Q14, which form a Darlington amplifier circuit. Transistor 1Q13 provides a low-impedance source for transistor 1Q14. From this amplifier, the signal proceeds to constant-current amplifier 1Q15 and 1Q16.

In this amplifier, transistor 1Q15 acts as an active load resistance for the collector of transistor 1Q16, providing a relatively low dc resistance and a relatively high ac resistance. In the audio range, the collector of 1Q16 works into an impedance which is sufficiently high to provide a constant-current source for the record head, yet allows full utilization of the dc operating voltage. From this stage, the signal is routed through a bias trap, consisting of choke 1C27, to the record head. Operating voltage is delivered to transistors 1Q13, 14, 15, and 16, only when the channel is in the record mode; therefore, these stages are inactive in any other mode.

The bias current and erase oscillator, 1Q17 and 1Q18, is a push-pull circuit, connected as a tuned flip-flop. Operating voltage is delivered only when the channel is in the record mode. Symmetry of the oscillator output waveform is adjusted by resistor 4R84 and capacitor 4C34 in the power supply diagram (Fig. 17-219C). Returning to Fig. 17-219B, the transformer-coupled oscillator output is delivered to record switch 2S5. When the switch is placed in the ready position, the oscillator output is routed through bias-adjustment resistor 2R68 and mixed with the audio signal. It is also connected through the erase-adjust capacitor 4C36 to the erase head and to erase coupling jack 4J12, also on the power-supply diagram.

When multichannel (four tracks) recording is used, erase coupling jacks

4J12, are employed to interconnect the oscillators and thus lock their frequencies together to prevent beat frequencies from being generated. (See Question 17.132.) When the record-selector switch is in the safe position, the oscillator transformer, record head, erase head, and coupling circuits are disconnected.

Referring to the lower portion of the block diagram and to Fig. 17-219A, the signal enters at connector 4J1 and is

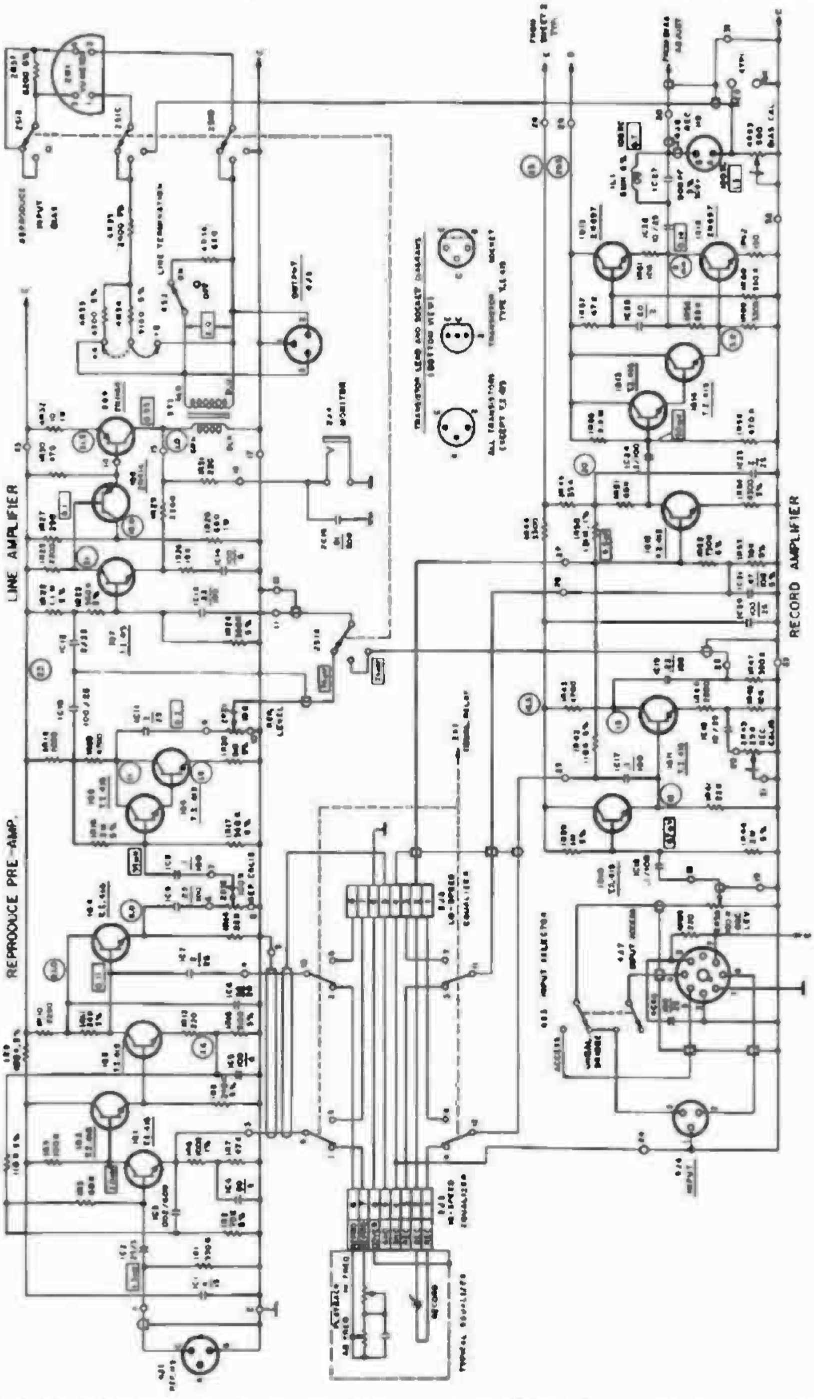


Fig. 17-219A. Schematic diagram for Ampex Model AG-300 magnetic tape recorder record-reproduce section.

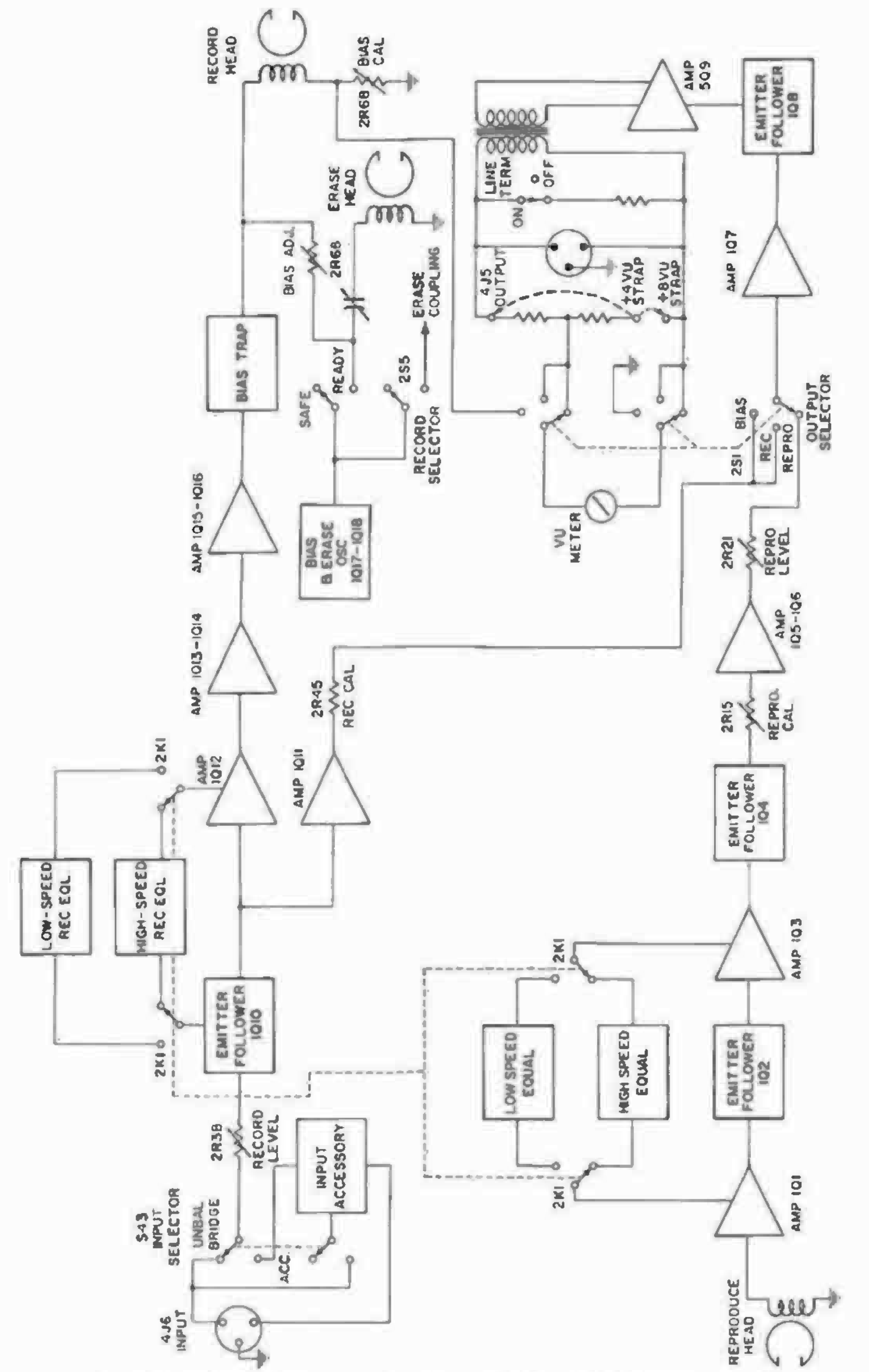


Fig. 17-2198. Block diagram for Ampex Model AG-300 tape recorder.

amplified by transistor 1Q1, then routed through emitter follower 1Q2 to amplifier 1Q3. High- and low-speed equalization is selected by relay 2K1, and is connected from the collector of 1Q3 back to the emitter of 1Q1. Direct-current feedback is provided through resistor 1R4, between these two stages.

Transistor 1Q4, an emitter follower, feeds reproduce calibrate control 2R15, to a Darlington amplifier formed by transistors 1Q5 and 1Q6. In this circuit, 1Q5 acts as a low-impedance source for 1Q6 to produce amplification of the signal with low noise. Leaving the 1Q5 and 1Q6, the signal amplitude is adjusted by reproduce level control 2R21, and is fed through the output selector switch to amplifier stage 1Q7 in the line amplifier circuit. The signal is now routed through 1Q8 to output stage 5Q9.

At the collector circuit of 5Q9 is a monitor jack. Here, headphones of 300-ohms impedance may be connected for monitoring purposes. The output signal is coupled through transformer 5T1 to

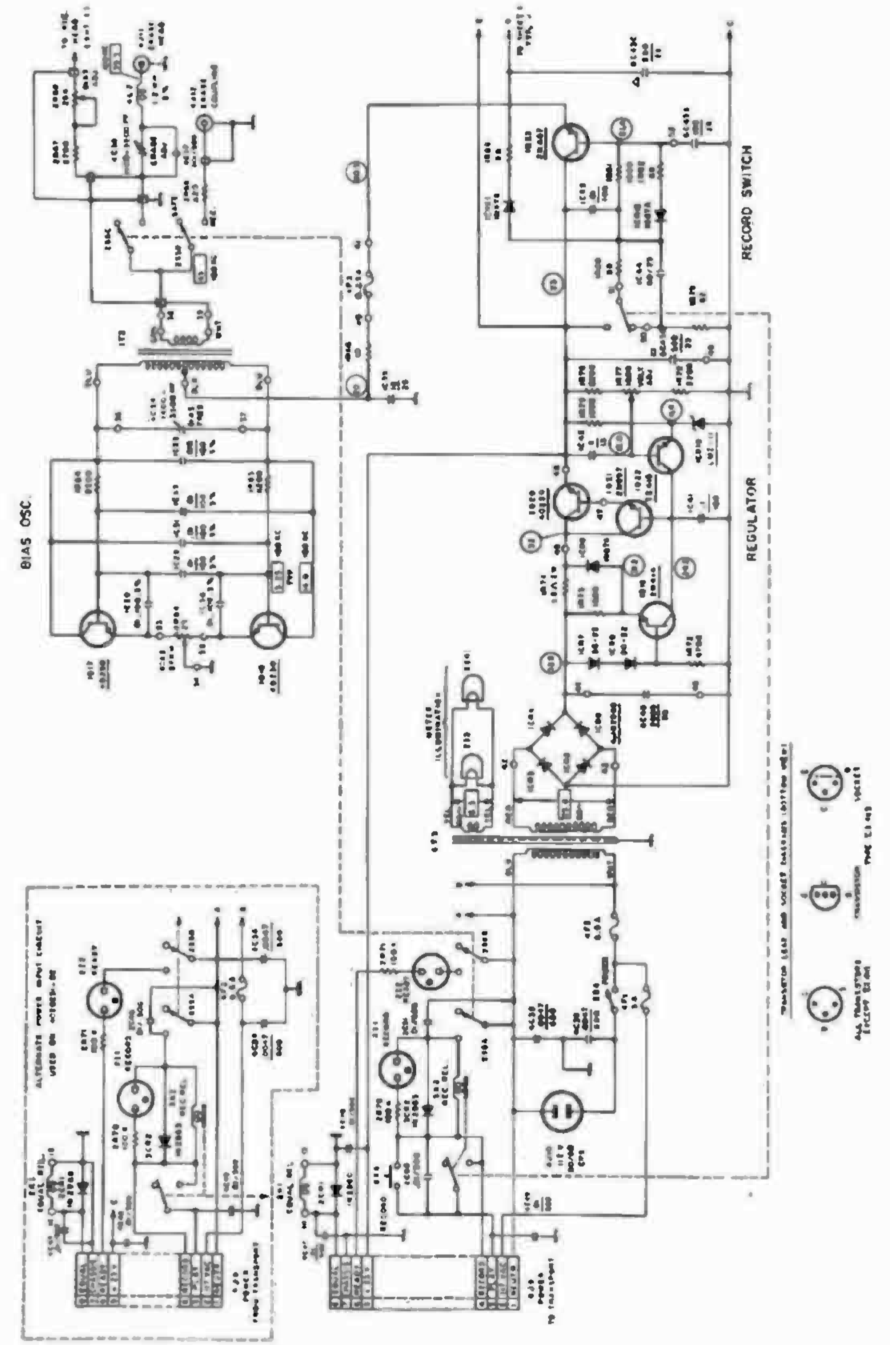


Fig. 17-219C. Power supply schematic diagram for Ampex Model AC-300 magnetic tape recorder.

line output connector 4J5. Line-termination switch 4S2 connects resistor 4R36 across the transformer secondary for tests and adjustments. Visual monitoring of the signal is provided by the VU meter. Depending on the position of the output selector switch, the meter will indicate the reproduce level, input level (record), or the bias current. The meter may be strapped to indicate zero for a plus 4- or 8-dB level.

Referring to Fig. 17-219C the schematic diagram for the power supply and bias oscillator, power from the line source is applied to terminals 1 and 2 of block 4J9 at the left, and then to the primary of power transformer 6T3, through fuse 4F2. One secondary of 6T3 is connected to the light in the VU meter; the second winding is connected across a bridge rectifier consisting of diodes 1CR3 and 1CR6, and then to a voltage-regulator circuit.

In the voltage-regulator circuit, a reference voltage is established by zener diode 1CR10. A sampling voltage is taken at 1R77, a voltage adjustment control. If the output voltage tends to vary with the load, the conductance of transistor 1Q22 changes. This in turn affects the conductance of transistors 1Q21 and 3Q20, connected in a Darlington circuit, so that the voltage is returned to normal level.

Transistor 1Q19 acts as a constant-current source. Diode 1CR9 and resistor 1R74 provide overload protection. If the current through 1R74 combined with that through 1R73 results in a voltage sufficient to break down diode 1CR9, transistor 1Q19 will be biased toward cutoff. This, in turn, underbiases the rest of the transistors in the regulator circuit. A plus 23-volts dc regulated output is delivered to the speed switch on the tape transport system, then returned to the electronics and used to energize equalization relay 2K1 in the low-speed position of that switch. It is also routed to all stages in the reproduce amplifier, the octal socket used for accessory input units, and the first three stages in the record mode.

Circled voltage values given on the schematic diagrams are taken between ground and the position indicated, measnred with a 20,000 ohm-per-volt meter. Voltages indicated in squares are ac signal voltages, taken at 500 Hz with a linear tape speed of 7½ ips, using an NAB equalizer. Signal voltages are measured with a vacuum-tube voltmeter with a 10-megohm input. The schematic diagram for the record and reprodncing equalizer circuits is given in Fig. 17-219D. This recorder with auxiliary equipment may be equipped to use ½-inch tape up to four sound tracks. Specifications for the machine are:

Frequency response:
- 15 ips plus-minus 2 dB, 50 to 18,000 Hz.
- 7½ ips plus-minus 2 dB, 40 to 10,000 Hz.

Signal-to-noise ratio:
- ¼-inch full-track 15 ips, 60 dB
- ½-inch four-track 7½ ips, 57-dB

Wow and flutter:
- 15 ips, 0.07 percent; 7½ ips, 0.07 percent.

Speed accuracy:
- 0.2 percent or 3.6 seconds for 30 minutes recording.

Even-order distortion:
- 500 Hz, 0.40 percent or less.

17.220 Describe the basic principles of a magnetic wire recorder.—Wire recorders are designed to pass a stainless steel wire approximately 4 mils in diameter over a ring-type recording head, which makes contact with one side of

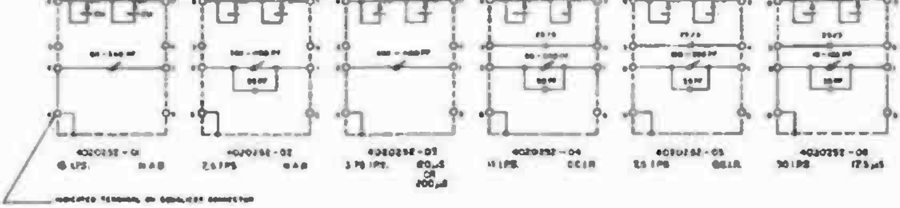


Fig. 17-219D. Schematic diagram for recording and reproducing equalizers used in Ampex Model AG-300 magnetic tape recorder.

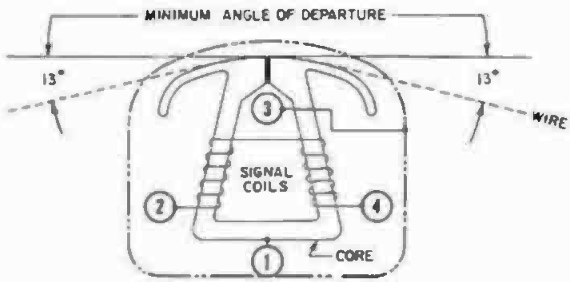


Fig. 17-220A. Cross-sectional view of a magnetic wire recording and reproducing head.

the wire (Fig. 17-220A). The construction of this head differs from that of conventional tape heads, as it has a groove for holding the wire at an angle of about 13 degrees, to provide a wraparound. The induced signal or magnetic field is confined to the wire side, or the region making contact with the head at a particular instant. When reproduced, the wire is rotated by the twisting action before it reaches the reproducing head, thus causing dropouts and variations in signal output. A typical polar plot of the magnetic field around a wire for several different frequencies is given in Fig. 17-220B.

The life of the wire is limitless, except for mechanical breakage. However, the signal-to-noise ratio is considerably less than for magnetic tape because of the small cross-sectional area of the wire, as compared to tape. Because of the magnetic properties of wire, it is generally erased by using a direct current or permanent magnet, which increases the noise, and decreases the signal-to-noise ratio.

In Fig. 17-220C are given the frequency characteristics for unequalized wire compared to magnetic tape. It will be observed that the frequency response of the tape running at 7½ ips is considerably better than wire running at 24 ips. Wire recorders have been built with fairly good frequency response, with intermodulation distortion on the order of 10 to 15 percent, with a THD of 5 to 8 percent. Wire recorders require pre- and post-equalization and a bias current similar to tape recorders. One of the principal drawbacks to wire recording is that it is not uncommon to have a 2-minute or more variation in a 10-minute program. Except for certain special devices, wire recording is considered obsolete.

17.221 What is a steel-tape magnetic recorder?—A magnetic tape recorder, designed some years ago, using a ¼-inch stainless-steel tape. The principles of the wire recorder discussed in Question 17.220 also apply to this recorder. Steel-tape recorders are now obsolete, except in special instances.

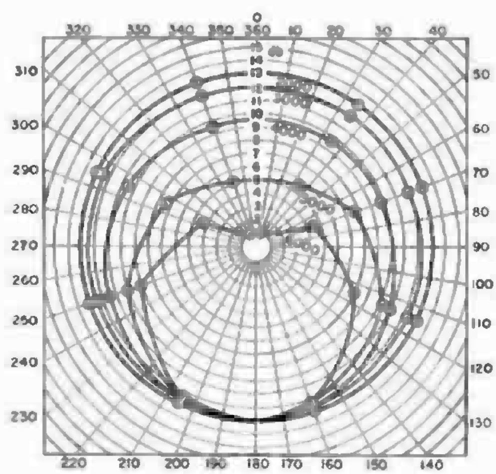


Fig. 17-220B. Variations in the output of a wire recorder with various amounts of wire twisting. (Courtesy, Minnesota Mining and Manufacturing Co.)

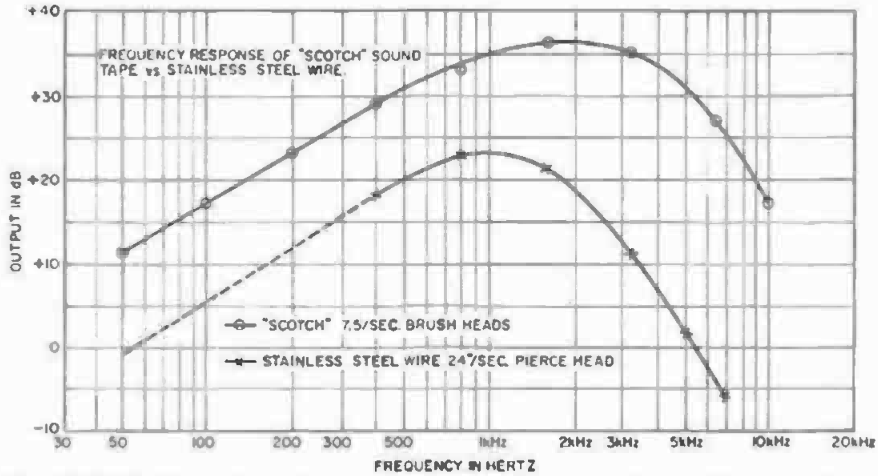


Fig. 17-220C. Comparison of unequalized magnetic tape and wire. (Courtesy, Minnesota Mining and Manufacturing Co.)

17.222 Describe the essential components and schematic diagram for a quarter-track tape recorder.—Roberts of Canada Model 770-A one-quarter inch magnetic tape recorder, manufactured by J. M. Nelson Electronics Ltd. (basic parts manufactured in Japan) is shown in Fig. 17-222A. This recorder employs the cross-field method of recording described in Question 17.39. Basically, the machine is a quarter-track two-channel stereophonic, or four-track monophonic recorder-reproducer. (See Question 17.151), completely self contained except for the microphones.

Referring to the front view in Fig. 17-222A, item A is a tape speed switch which, in combination with a removable bushing I placed over the end of the capstan motor shaft J, permits tape speeds of 1⅞, 3¾, and 7½ ips. However, it can also be adapted for 15 ips. At B

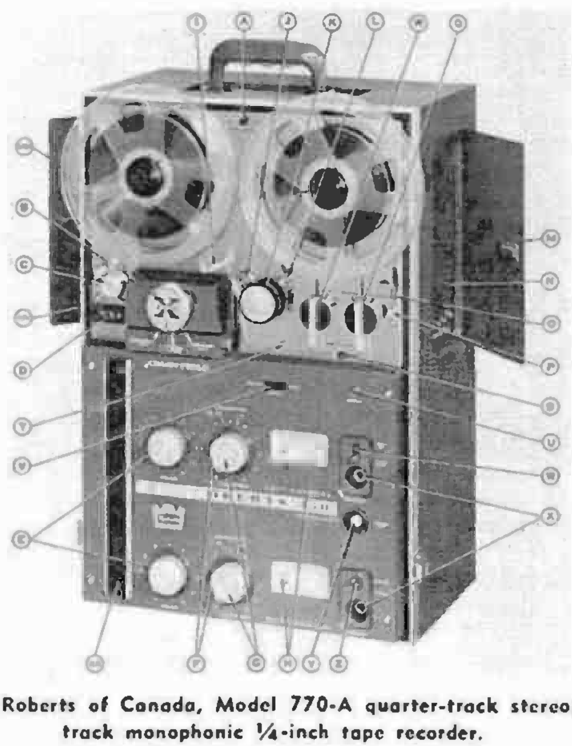


Fig. 17-222A. Roberts of Canada, Model 770-A quarter-track stereophonic, four-track monophonic ¼-inch tape recorder.

is a track-selector switch for setting the recording-reproduce circuits to either monophonic or stereophonic, and it is connected mechanically to an arrangement that moves the recording-reproduce heads to their proper track placement, as discussed in Question 17.151.

A multipurpose device serving as a spring-loaded tape guide and cleaner and flutter filter C permits the tape to

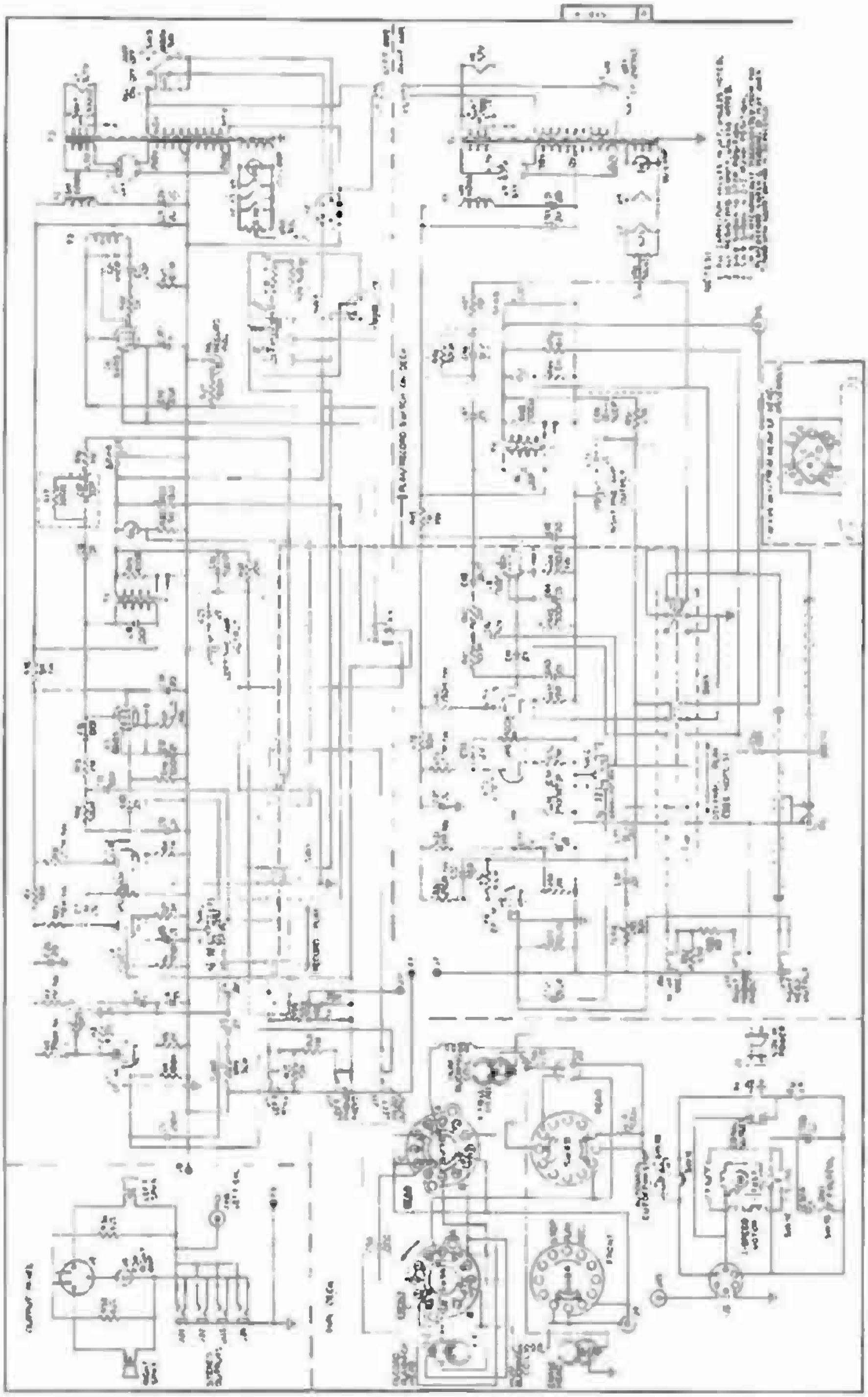


Fig. 17-222B. Schematic diagram for Roberts of Canada, Model 770-A tape recorder.

make a right angle bend from the supply reel to the transport system with a minimum of drag. This in connection with a spring-loaded arm (on the flutter filter) aids in ironing out irregularities in the tape travel which might cause flutter. The tape guide also includes a felt-pad for cleaning the tape before it arrives at the heads. Index counter D reads from zero to 9999 and may be used for supplying a reference number for later locating a given recorded selection.

Items E are gain controls for the left and right amplifiers and are used for both recording and reproduction. The outer control rings F are tone controls which are operative only in the playback mode. Knobs G are recording equalizers which are changed for the particular speed in use. Two VU meters H indicate the levels in both the record and reproduce modes and have standard ballistic characteristics as required for such meters.

I is a pin for storage of the removable capstan speed-change puck. The puck is mounted on the end of the motor shaft J, with a neoprene pinch-wheel K below, which pulls the tape over the heads, because of the pinch effect of the latter two items on the tape. A stop arm L will automatically drop, if the tape breaks or runs out, shutting off the transport mechanism. However, if switch S is in the ON position and the tape runs out, the complete system will shut off, including a convenience outlet at the rear of the machine for connection of auxiliary equipment. Two doors M and MM on the right and left sides cover two monitor loudspeakers N and NN during transport. When these doors are opened, they act as reflectors for the speakers, directing the sound forward. A pause-edit switch O may be used to stop the tape transport momentarily for editing purposes in either the record or reproduce modes. The main power switch W has three positions—off, all-on, and motors-only. In the all-on position, the capstan, fan motors, and electronics are turned on. The motors-only position is used when an output signal is taken directly from the heads for driving external equipment, and the internal electronics are not required. Push button P is the start button for both recording and playback; however when recording, the interlock button T must also be depressed before switch R can be turned to the recording mode, thus preventing an accidental erasure of previously recorded sound track. For fast-forward or rewind, switch Q is set to the desired position and provides fast rewind or forward speeds of 1200 ft iu 75 seconds.

Pilot light U indicates when the machine is in the record mode. Monitor switch V has three positions—normal, monitor, and mute. In the normal position, the internal and externally connected speakers are operative on playback. In the monitor position, the internal monitor speakers and external speakers will be activated while recording. In the mute position, the internal speakers are cut off, but jacks X are activated for use with external stereophonic speakers. Switch Z may be used to cut off the right speaker when recording monophonically.

Access is afforded to the preamplifier inputs for microphones, playback head outputs, and phono jacks by opening panel XX. Switch Y is used only when recording sound with sound. This feature permits a second sound track to be recorded in synchronization with a previously recorded track. When the two recordings are played back, they are reproduced in sync on the left and right speakers. Two power supplies are used; one each for the left and right sides. A total of nine vacuum tubes are required. The internal monitor amplifiers will each deliver 6 watts to an external load. Stereo-headphone connections are available near the right hand lower portion of the monitor speaker.

The signal-to-noise ratio, relative to 100-percent modulation is 55 dB. Wow and flutter 7½ ips, 0.12 percent; at 3¾ ips, 0.2 percent; at 1⅞ ips, 0.3 percent. Erasure is minus 60 dB. Frequency response at 7½ ips is plus-minus 2 dB, 40 to 22,000 Hz; at 3¾ ips, plus minus 2 dB, 40 to 15,000 Hz; at 1⅞ ips, 3 dB, 40 to 10,000 Hz.

The equalization meets the NAB Standards in all positions. (See Question 17.162.) The unit employs only a single motor, which functions to pull the tape and operates both the feed and rewind reels through a system of friction pucks. The motor is of the hysteresis synchronous type, turning 1800 and 3600 rpm, and employing a 24-slot wavewound rotor, with a heavy fly wheel belt-coupled to the capstan.

Internal degaussing of the magnetic heads is made possible by a simple procedure. The schematic diagram for the recorder appears in Fig. 17-222B. Basically the amplifier system consists of two sections, one for each channel, having the same number of tubes, except for the left channel, which includes a bias oscillator tube. The right amplifier is similar except for the bias oscillator. Two complete power supplies are provided. The selector switches at the left, SW8 and SW9, select the proper circuitry for either the recording or playback modes. Hum-bucking coils are connected in series with the heads to reduce hum in the output circuits to a minimum.

17.223 Define the term "looping." —Looping is a term used for indicating that a motion picture is to be post-synchronized—that is, the sound must be replaced with a new sound track in absolute synchronization with the lip-movement and action of an existing motion picture.

Post-synchronization of a picture may be accomplished in two ways. (1) A continuous loop of the original sound track is played back over headphones with the picture for the actor to listen for timing and inflections. He then records a new sound track, while watching the picture action and listening to the original sound track. (2) The same procedure is also used, except without the picture (which is preferred by many actors), as the picture may tend to distract an observer's attention from the sound track.

17.224 Describe a simplified system for looping sound tracks.—In the system shown in Fig. 17-224, only two machines are required, a recorder and a recorder-reproducer. The actor speaks his lines into the microphone, using only the original sound (guide) track. After the dialogue has been recorded, the recording patch is removed and the newly recorded sound track played back in sync with the guide track for evaluation of quality and synchronization. If the circuits of the mixer are well isolated, it is possible to leave the playback and guide-track circuits permanently patched, removing only the record circuit. If the recorder employs a single record-reproduce head, the operation is somewhat simplified.

The same setup can be employed using a ¼-inch sync-pulse tape recorder in place of the recorder-reproducer, provided a sync pulse is recorded on the new track and a resolver is used

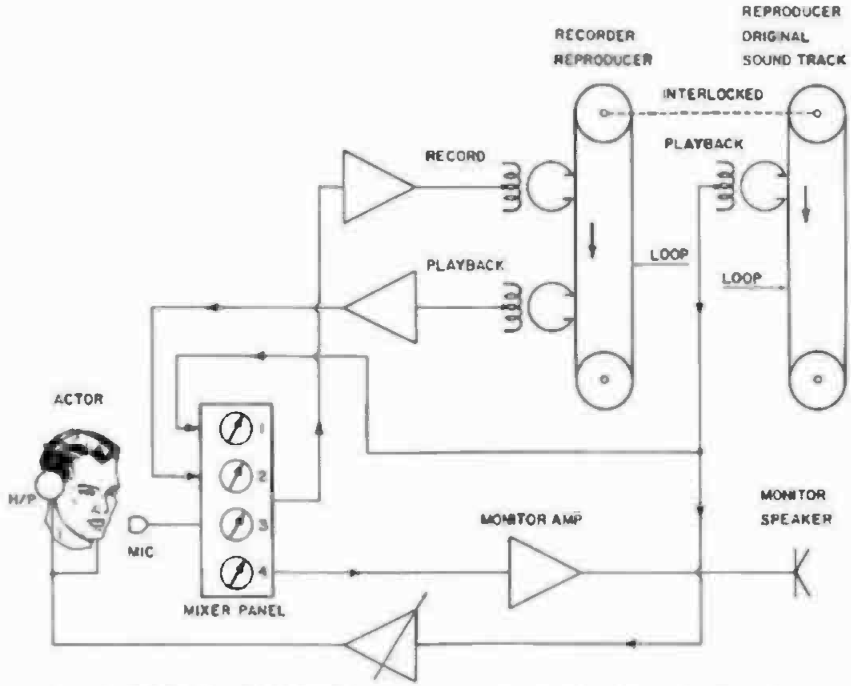


Fig. 17-224. Simplified method of looping sound tracks without picture.

when playing back the newly recorded sound track. Other systems of looping are described in Questions 17.225 and 17.226.

17.225 ***Describe a method for looping foreign dialogue.***—Looping dialogue for foreign versions is somewhat more complicated than looping the original language. In addition to the looping equipment (Questions 17.226 and 17.227) some means of indicating to the actor the volume range of the original recording is desirable but not mandatory. This is generally accomplished by projecting on the screen, with the picture, a vertical line which varies in length and represents the volume level of the original sound track. The new dialogue to be recorded is also projected in sync with the picture along the bottom of the screen to prompt the actor. As a rule, the looping of foreign dialogue is an art highly specialized within itself. Several patents have been applied for or have been issued on systems developed for this special type recording.

17.226 ***Describe the principles of a virgin magnetic looping system.***—The principal reason for looping or post-synchronization of a motion picture may be for one of several. (1) The dialogue shot on location has traffic, airplane or other unwanted noise in the background. (2) Wind machines were used on the set, blanking out the dialogue. (3) The original voice is to be replaced with that of another actor (re-voiced). (4) Certain words are indistinct. (5) Dialogue is difficult or impractical to record on location. (6) Replacement of the original dialogue with that of a foreign language is sometimes necessary.

Looping or post-synchronization is accomplished by bringing the actors to a looping stage, and while watching the picture on the screen, record new sound tracks in synchronization with the picture action.

Looping stages are generally so constructed that the interior acoustics may be controlled over a wide range of reverberation, from a dead stage to that of a live one, by the use of hinged panels on the side walls as discussed in Question 2.110. As a rule, an attempt is made to match the acoustics of the orig-

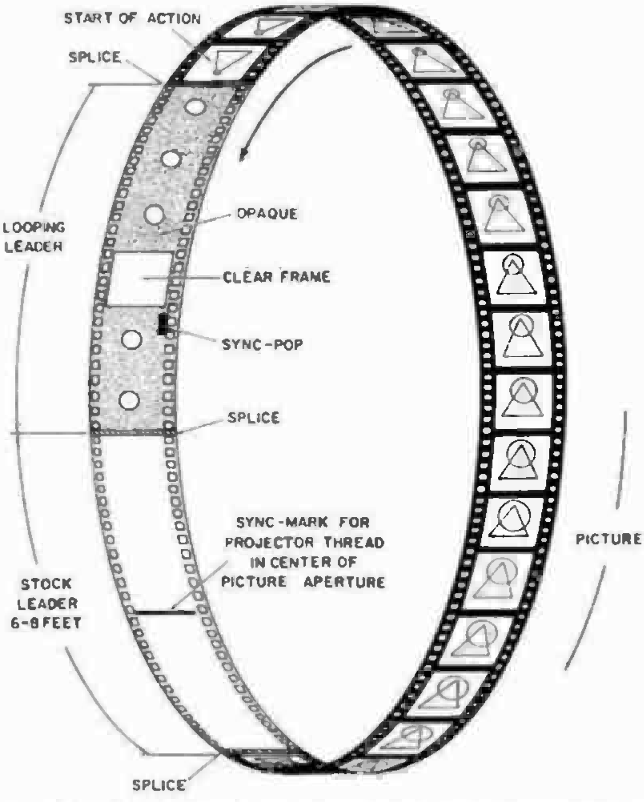


Fig. 17-226A. Picture loop for post-synchronization recording.

inal scene; however, if this is not possible, it is recorded slightly on the dull side and corrected later in the rerecording (dubbing).

In large studios where several companies are in daily production, looping may become a continuous operation, particularly where a large number of western-type pictnres are in production. The looping equipment must perform several functions. (1) Permit the actor to listen to the original sound track (guide track) for tempo and intonation, either over a loudspeaker behind the screen, or with headphones. (2) Record a new sound track in synchronization with the picture, while listening to the guide track. (3) Immediate playback with picture for checking the newly recorded sound track for synchronization and quality. (4) If the take is not satisfactory, it can be erased immediately and the system returned to the recording mode.

To facilitate the operations, a special three-position mixing panel is used, housing the microphone preamplifiers, mixer controls, equalizers, recording amplifier, and intercommunication to the projectionist and the machine room. Provision is also made for starting the machines, push button for the various modes of operation and other controls. A remote control for changing the mode of operation from record to playback is included for the dialogue director.

Three loops are required for each scene to be remade. The projector loop carries no sound, but the head-end has a special looping leader with cue marks and a sync-pop. The action (picture) to which the new sound track is to be matched is spliced to the tail end of the leader (Fig. 17-226A). The loop for the reproducing machine carries only the original dialogue (guide track) to be replaced. The recorder loop consists of a continuous piece of magnetic film. Fig. 17-226B shows that each loop for threading and synchronization purposes has a start mark relative to each other. (The three machines must be driven selsyn-interlock. Question 3.49).

It will be observed that the picture leader is opaque and has one clear frame and a sync-pop mark 20 frames in advance of the clear frame. Also, three white dots are placed behind the

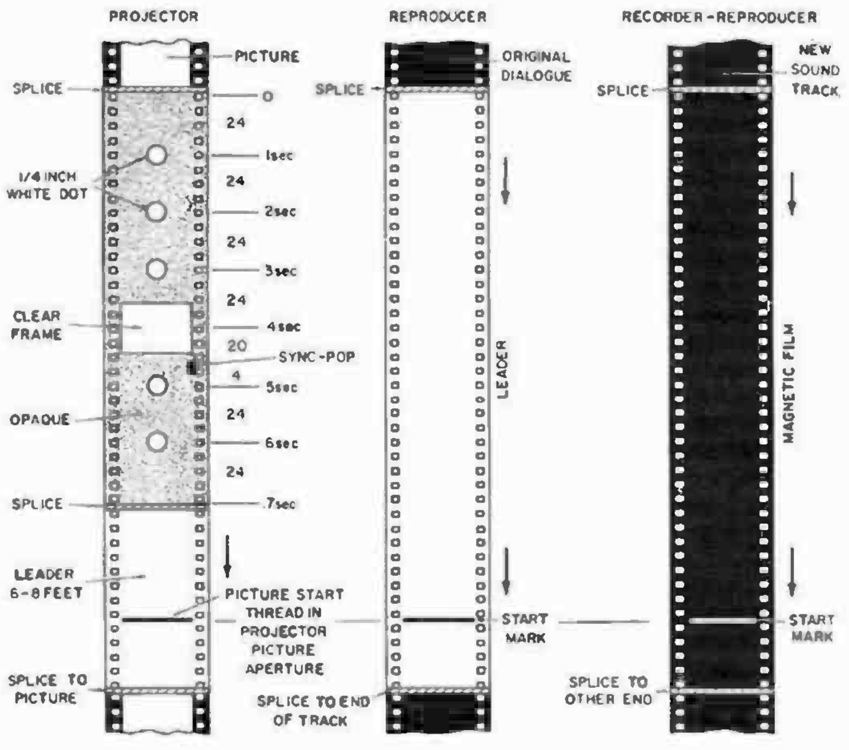


Fig. 17-226B. Loop structure for virgin loop recording system. The picture loop leader is opaque except for the white dots and clear frame.

clear frame, 24 frames apart. The dots and the clear frame flash on the screen at the rate of 1 per second, or a total of 6 seconds, to impart rhythm to the actor and cue him for the action to come. Sometimes numerals are used instead of dots. The start mark for the projection machine is placed 6 to 8 feet in advance of the first white dot. When the clear white frame passes the projector aperture, a bright flash appears on the screen and the sync pop mark is picked up by a photocell in the projector sound head and recorded on the sound track, to be used later by the editorial department to synchronize the action with the new sound track. Plain leader stock with a start mark is spliced ahead of the dialogue guide track. The recorder loop is a continuous piece of magnetic film long enough to record the dialogue, plus a few extra inches. The recorder start mark is placed near the splice, thus assuring that the splice will not be in the dialogue area when recording.

When the machines are threaded, the start mark for the picture is placed in the center of the projector picture aperture. The start mark for the guide track is placed over the reproducer head gap, and the start mark for the recorder is placed over the record head-gap. In operation, the three machines are interlocked and the start marks are rechecked. The marks on the recorder and guide track must be within 1 sprocket

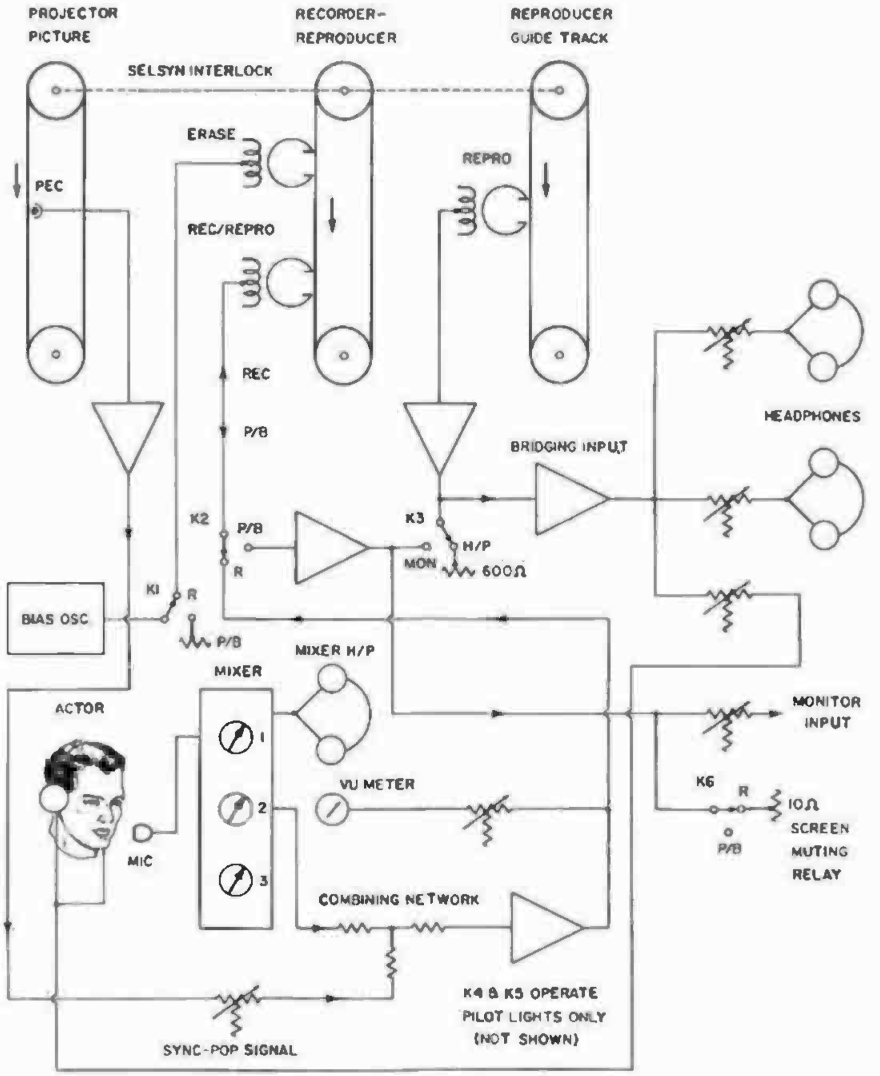


Fig. 17-226C. Simplified block diagram for automated virgin loop recording system.

hole of the picture. The machines are started. The actor listens to the guide track over the speaker behind the screen or over headphones (generally headphones are used). When he is ready, the record button is pushed, actuating several relays which set the system in the record mode. A take is made. If the actor or dialogue director feels it is not right another take is made, as the first take will be erased automatically if the system is not thrown to

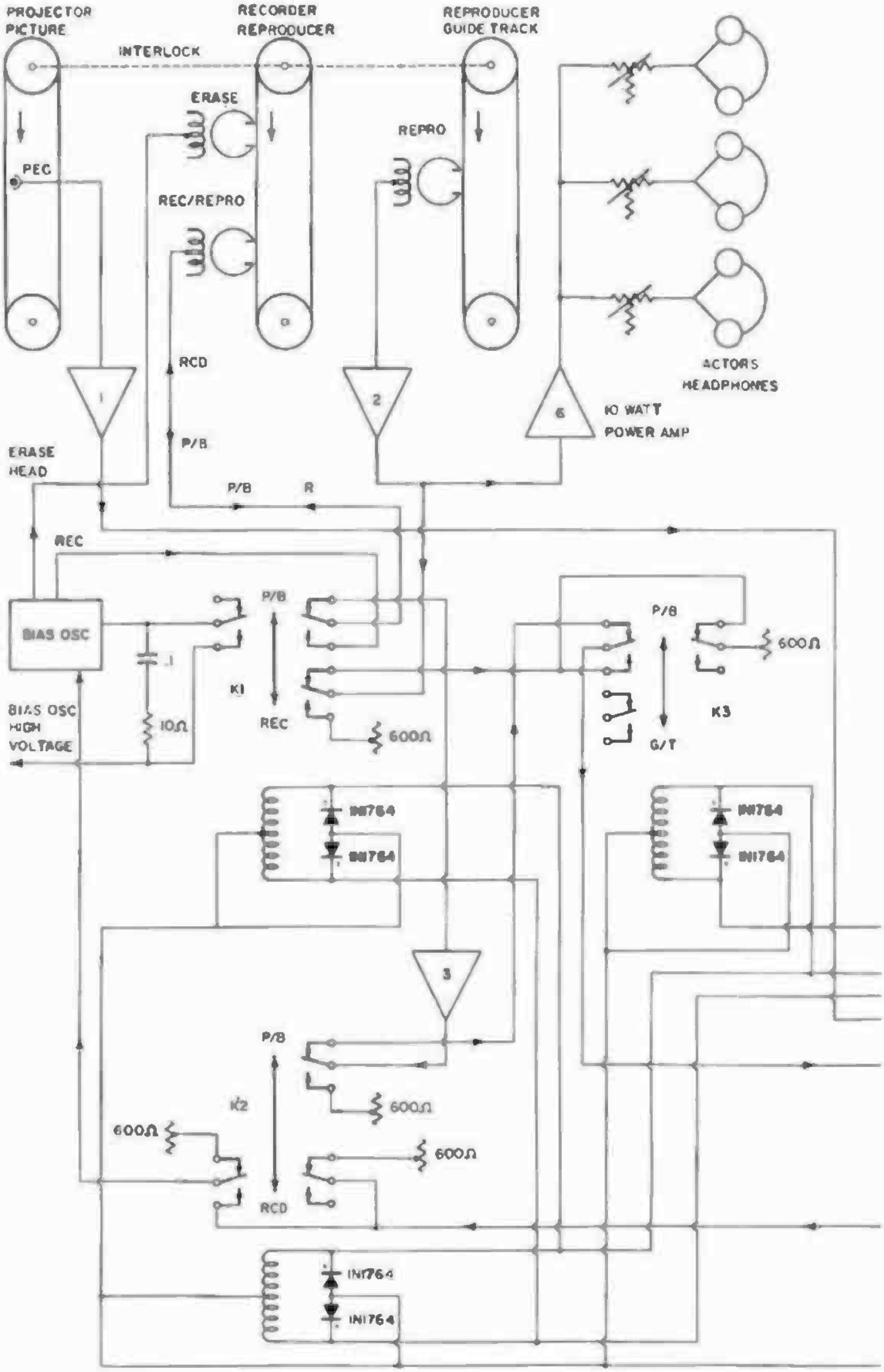
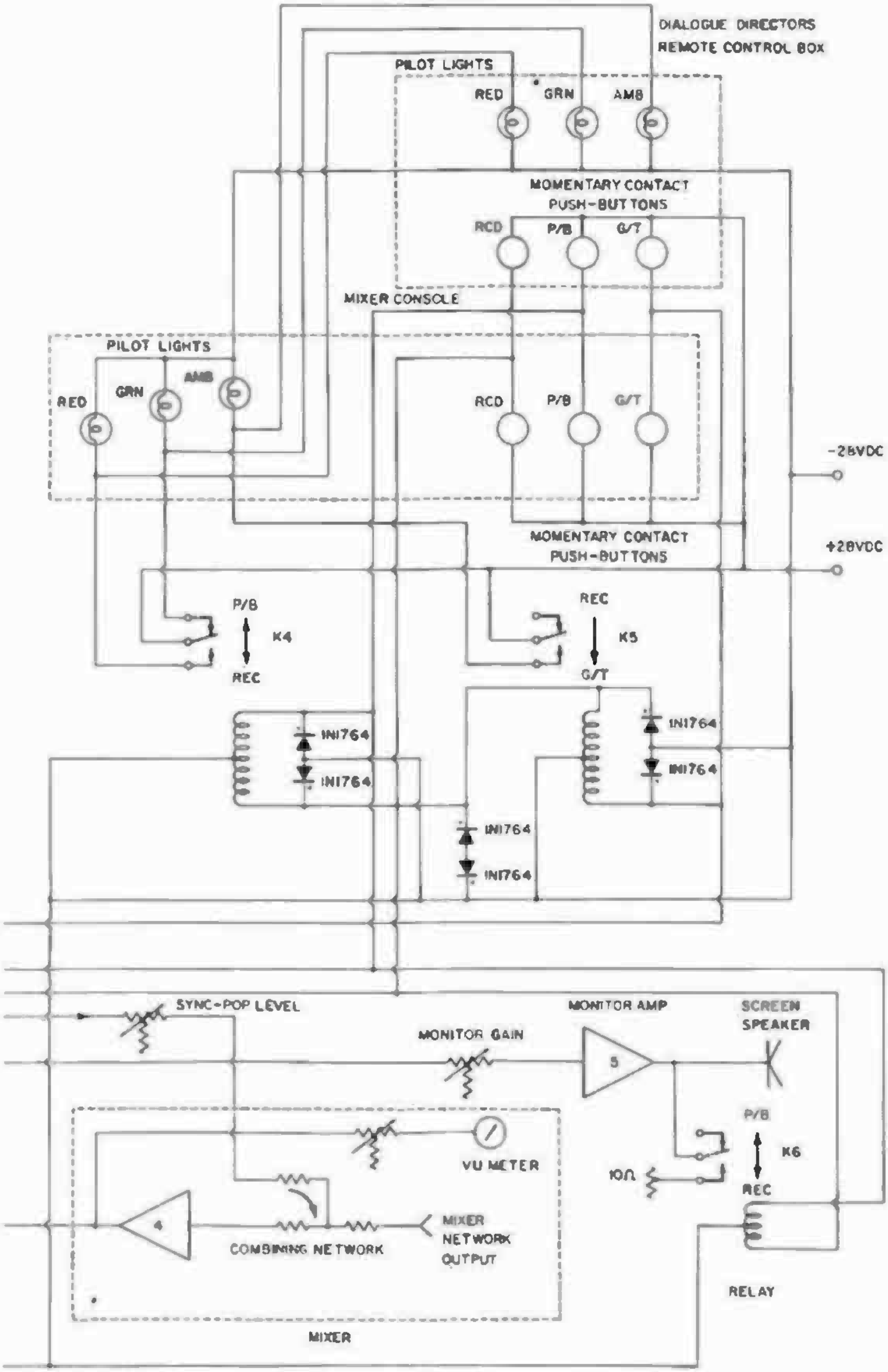


Fig. 17-227. Schematic diagram for

the playback mode within 4 seconds after recording.

As many tracks as necessary are made. When a satisfactory one is had, the system is set to the playback mode. The newly recorded sound track with picture is played back immediately over the loudspeaker behind the screen.

The pop-mark is recorded on the new sound track and will be heard when the clear frame of the picture flashes on the screen. This assures that



virgin loop magnetic recording system.

the sound and picture are in sync, and any variation in lip movement or other action is due entirely to the actor being out of sync with the action.

The special picture leader is a print made from a negative. The sync pop-mark is three or four black marks placed on the negative. The holes are made with a film punch. Once the time for the system to come up to speed has been determined, the same length of leader is always used. A block diagram of the system is given in Fig. 17-226C.

17.227 Describe the schematic diagram and method of operation for the virgin looping system of Question 17.226.—The schematic diagram for the automated virgin looping system discussed in Question 17.226 is given in Fig. 17-227. The system shown is completely push-button operated. The control circuits are so designed that they can be patched into an existing installation without disturbing the normal interconnections. The mixing panel consists of a 3-position mixer network with equalizers, on-off key, VU meter, talk-back system to the projectionist and machine room, push buttons and pilot lights for setting the system to its three modes of operation, microphone preamplifiers, lock and run switches for the selsyn-interlock system, and ready lights from the projection and machine rooms. A remote control box is also provided for the dialogue director for changing the modes of operation; however, the control of the selsyn-interlock system is generally operated by the mixer.

At the upper left in Fig. 17-277 are three machines previously mentioned in Question 17.226. At the upper right are three momentary-contact push buttons which apply a 28-Vdc pulse to six latching-type relays. (See Question 25.159.) The latching relays employ a center-tapped actuating coil connected to the common or negative side of the 28 Vdc. With a positive pulse applied to one end of the coil, the armature moves to a given position. With a positive pulse applied to the other end of the coil, the armature moves in the opposite direction. The armature always remains in the last position until another pulse is applied to the opposite end of the coil.

Depressing the record button actuates relays K1, K2, K4, and K6 downward, and K5 upward. Relays K1 and K2 connect the record-reproduce head on the recorder to the output of recording amplifier 4, and energies the bias oscillator. Relay K4 turns on the red record pilot lights, and K6 mutes the monitor speaker behind the screen, while K5 moves upward cutting off the guide-track amber light (if it had been the last operation).

Actuating the playback button energizes relays K1 and K2 upward, de-activating the bias oscillator and connecting the record-reproduce head to the input of preamplifier 3. At the same time, relay K3 moves upward, connecting the output of the preamplifier to the monitor amplifier input through the monitor gain control to monitor amplifier 5. Relay K4 moves upward connecting the playback green pilot lights, with K6 also moving upward, lifting the 10-ohm muting resistor from the monitor line.

If the newly recorded sound track is being played back over the studio monitor system, the guide track may be substituted by pressing the guide-track button. This energizes relay K3 downward, connecting the output of the reproducer to the monitor system. To return the system to the new track, the playback button is again pushed. This resets relay K3 to the playback position. Thus, the original sound track may be compared to the new sound track. Relay K5 turns the amber guide-track pilot light off and on. Headphone monitoring is always available across the output of the guide-track machine. Similar operations are performed by the dialogue director's remote-control unit.

Diode click-suppressors are installed across each relay coil to reduce clicks when switching from one mode to another. These devices were discussed in Question 24.67.

The bias switching circuit is only basic, as it will vary with different type recorders. The principal point of operation is that the bias must be completely off when in the playback position, which can be accomplished by breaking the high voltage to the oscillator circuit. The mixer may be a small roll-away type, with all operating controls, preamplifiers and recording amplifiers self-contained.

17.228 Give a basic block diagram for a magnetic tape to disc record transfer channel.—It is the present day practice to record the original sound for disc records on magnetic tape or film using three, four, or more sound tracks. After editing, the master tape is transferred to a master disc record from which stampers and pressings are later made.

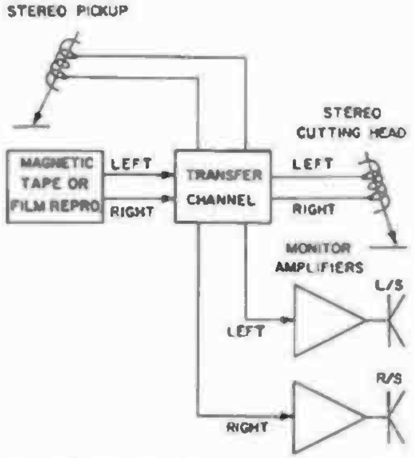


Fig. 17-228. Basic components required for a magnetic tape or film transfer channel.

The block diagram for such a channel is given in Fig. 17-228. The equipment consists of a magnetic tape or film reproducer that feeds the program material into the transfer channel containing the necessary amplifiers, equalization, filters, VU meters, monitor amplifiers and associated equipment. A stereophonic pickup and external turntable, recording lathe, and monitor loudspeakers complete the basic equipment. A detailed discussion of transfer channel design appears in Question 13.216. (See also Questions 18.331 and 18.332.)

17.229 Describe a master magnetic tape recorder employing a high- and low-level amplifier system.—The state of the art of transferring program material from a magnetic tape to a disc record has advanced to the point where the commercial disc pressing is no longer the limiting factor in the relation to signal-to-noise ratio. The limiting factor is now the source feeding the disc recorder, rather than the disc itself. Sources of noise in the reproduction of

Fig. 17-229A. Minnesota Mining and Manufacturing Co. (3M), Dynatrack professional tape recorder.

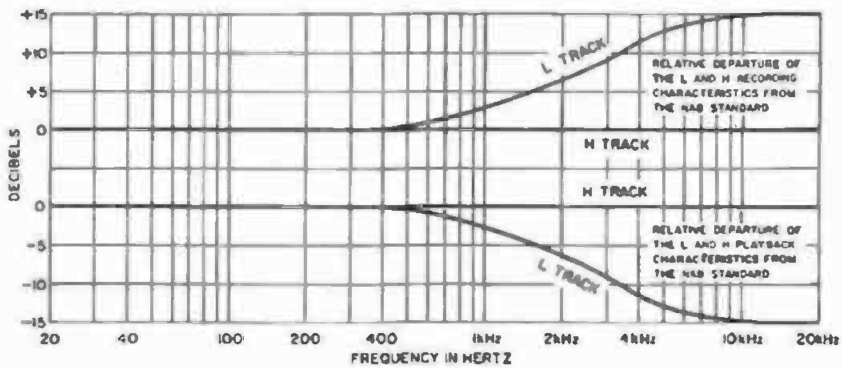


Fig. 17-229B. Recording and playback characteristics used in a 3M Dynatrack tape recording system. Curve (1), relative departure of "L" and "H" recording characteristics from the NAB standard for 15 ips. Curve (2), relative departure of playback characteristics from the NAB standard for 15 ips.

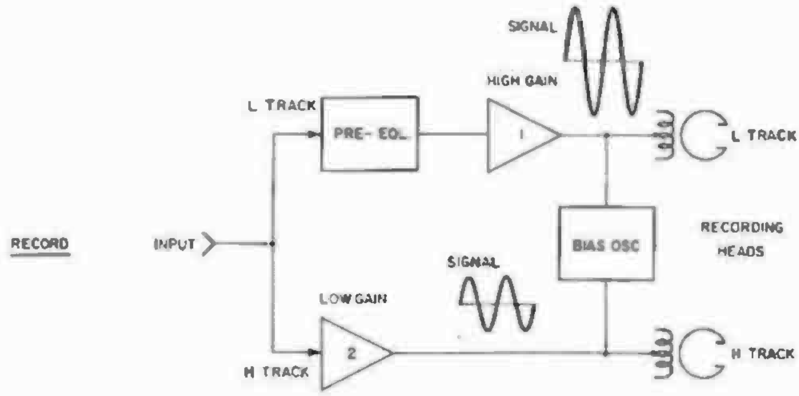


Fig. 17-229C. Basic principle of operation for 3M Dynatrack tape recorder-reproducer in record position.

magnetic tape recorders may be broken down into three categories: (1) internal noise generated in the microphones, preamplifiers and the mixers used in the original recording system, over which the transfer operation has no control; (2) noise generated within the electronics of the tape recorder-reproducer used for transfer operations; (3) noise induced by the tape during recording and reproduction. All these are factors affecting the dynamic range of the final product.

Dynamic range in magnetic tape recorders has been improved considerably over the past years due to improvements in the electronics, the manufacture of magnetic tape and film, and the production of disc record pressings. In the early tape recorders, the dynamic range was limited to about 45 dB, while today's professional equipment will measure a minimum of 55 to 60 dB and for some designs, 76 to 80 dB, using a weighted curve with reference to 100-percent modulation for a signal of not more than 3 percent total harmonic distortion. Weighted curves and networks are discussed in Questions 5.98 and 17.159.

The recorder shown in Fig. 17-229A is a development of Minnesota Mining and Manufacturing Co., under the direction of John T. Mullin, and employs a two-section recording system, using low- and high-level sound track. During playback the two tracks are automatically switched from one to the other, thereby always selecting the optimum track for the lowest distortion and greatest signal-to-noise ratio. This increase in the signal-to-noise ratio has been accomplished by the use of silicon transistors and the matching of the playback head to the preamplifier over the entire frequency range, and the use of a two-track system to be described. However, it should be pointed out, this increase in dynamic range, aside from the two tracks and transistor electronics, is achieved by the use of 3-M type 201, 202, or 203 magnetic tape.

Each channel to be recorded (one for monophonic, two for stereo, and three

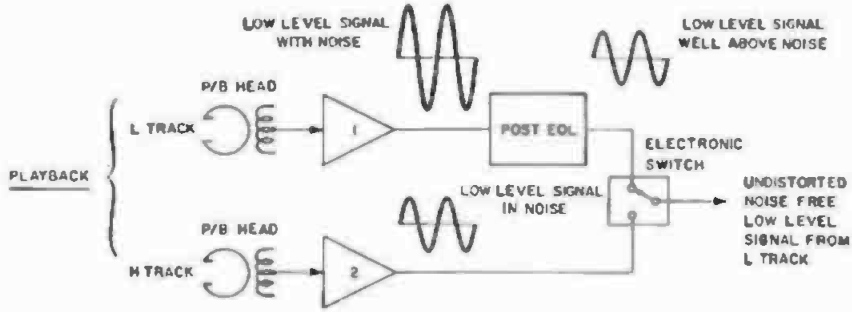


Fig. 17-229D. Basic principle of operation for 3M Dynatrack tape recorder-reproducer in playback position low-level track "L".

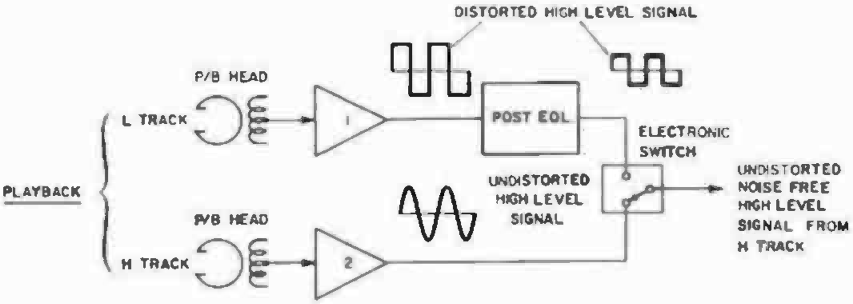


Fig. 17-229E. Basic principle of operation for 3M Dynotrack tape recorder-reproducer in playback position high-level track "H".

for multichannel stereo) has two tracks recorded simultaneously in the same head stack (Fig. 17-229C). One track is recorded at the Standard NAB recording level. The second track is modulated with the same information, but records the high frequencies at a higher level (more gain) employing a pre-equalization curve rising 15 dB from 400 Hz to 15,000 Hz (Fig. 17-229B). Because of the high-frequency equalization in this track, it will go into high-frequency overload considerably below the 100-percent modulation indicated by the VU meter. However, when reproducing low-level passages, this track will be free from noise and distortion at a level well above the tape noise. The normally recorded NAB track can handle much higher signal levels before amplitude distortion is reached. Therefore, this track is termed the high or "H" track. The first track, with its pre-equalization and extra high gain, is much better suited for recording low-level signals that might be lost in the noise. This track is termed the "L" or low-level track. Because of the higher recording level of the "L" track, it will overload more easily at the heavier modulated signal levels in the high frequencies than will the "H" track. Block diagrams of the basic principles of recording and playback are given in Figs. 17-229C to E.

When reproducing signals that might be lost in the noise, although recorded at the normal signal level, the "L" track delivers an undistorted signal well above the noise level. The basic principle of this system is that the two tracks are recorded simultaneously with the same signal at two different levels, but on playback, behave differently by going into distortion at different levels. The system automatically plays back

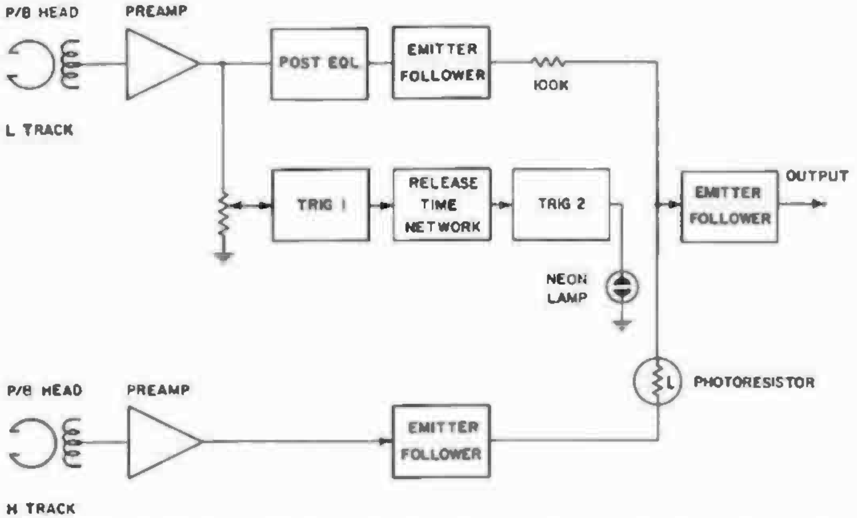


Fig. 17-229F. Basic playback circuit showing the transistor electronics and triggering units.

the optimum track at the proper signal level at a given instant, with minimum distortion and background noise at any instant of playback.

The selection of the proper sound track during playback is dependent on an electronic circuit which is fully automatic (Fig. 17-229F). The circuit is so calibrated that the signals from the "L" track are reproduced up to a point where the signal approaches 1-percent THD. At this point, the output is taken from the "H" track, which has a signal level 15 dB below the 1-percent THD-point output of the signal, which cannot continue rising until it reaches the normal, or 100-percent indication on the VU meter.

The transfer from the "L" to "H" sound track takes 200 microseconds. When the signal level on the "L" track drops below the 1-percent THD, the electronic circuit automatically transfers the signal from "H" to "L" in approximately 10 milliseconds. The change from one track to the other is faster than the human ear can detect, therefore the transfer is not noticeable during the reproduction. No switching occurs below 400 Hz, where the tape background noise is generally low.

To equalize the playback levels, a post-equalizer is connected in the playback circuit of the more heavily recorded "L" track, which compensates for the higher level at which it was recorded (Fig. 17-229E). Pilot lights on the control panel indicate which tracks are being reproduced. The characteristics of this recorder-reproducer are in accordance with the existing NAB Standards for magnetic tape recording.

To reduce flutter to a minimum, the transport system is of different design from that of the conventional tape recorder, inasmuch as the unsupported tape path is only 3½ inches in length, compared to the conventional transport system which may vary from a few inches up to 12 inches (Fig. 17-229G). Flutter is reduced by creating a difference in the effective diameters of the capstan so that a smaller-diameter capstan drives the incoming tape, and a larger-diameter capstan pulls the outgoing tape. The incoming roller A (Fig. 17-229H) is contoured to press the tape firmly into matching grooves of the smaller diameter of capstan B. The outgoing roller C is shaped to press the tape firmly against the ridges of the larger diameter capstan. The differen-

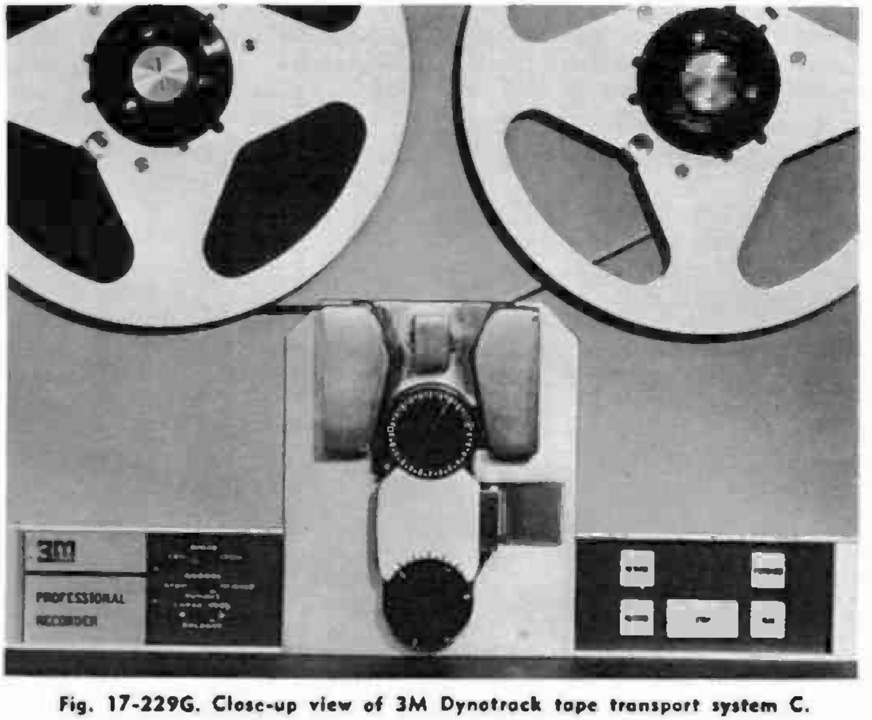


Fig. 17-229G. Close-up view of 3M Dynotrack tape transport system C.

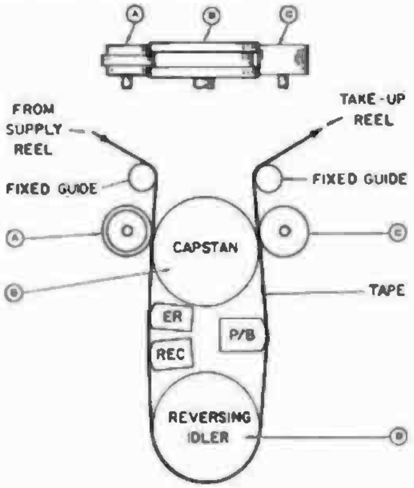


Fig. 17-229H. Tape transport system for 3M Dynatrack tape recorder.

tial capstan diameter is constantly trying to extract more tape than is being fed into the loop, and develops the necessary tape tension for the heads by means of the slight elasticity of the tape, which is kept within safe limits of the tape elasticity. A hysteresis-synchronous motor with high damping and large fly wheel on the inner end of its shaft drives the capstan by means of an inelastic tensilized polyester belt, running from a pulley on the capstan shaft. The capstan also carries a large fly wheel.

It is claimed for this type transport system that the shorter tape path reduces longitudinal oscillation and eliminates the need for tape guide posts. Flutter components in the 0.5 to 300-Hz range, scrape flutter, and flutter in the 300 to 5000-Hz range is reduced because of the shortened tape path.

As the tape is dropped onto the threading slot, it interrupts a light path to a photocell switch, which readies the controls for operation. Simultaneously, the supply and take-up torque motors are energized and apply sufficient tension to remove slack from the tape path and protect it from breakage. A second photocell sensor works in conjunction with the run-out switch to perform whatever switching function has been preselected to occur at the end of the reel. Depending on the presetting of the switching, the transport will allow normal tape run-out, rewind, stop, or start of another tape-transport system (second machine).

An interior view of the mechanical assembly, with the front panel removed, is pictured in Fig. 17-229I, with a rear interior view shown in Fig. 17-229J. The electronics are all solid-state. Specifications for the recorder are: Frequency range at 15 ips, plus-minus 1 dB 40 to 15 kHz; plus minus 2 dB 30 to 17 kHz; at 7½-ips, plus-minus 1 dB 40 to 12 kHz, plus-minus 2 dB 30 to 15 kHz. Channel separation is 50 dB at 500 Hz. Bias frequency is 120 kHz. Tape width, normal ¼ or ½ inch and 1 inch available. Signal-to-noise ratio, 80 dB or greater (employing ¼-inch tape width per channel). Flutter at 15 ips 0.5 to 300 Hz, 0.04 percent; 0.5 to 5 kHz, 0.05 percent; at 7½ ips 0.5 to 300 Hz, 0.07 percent; and 0.5 to 5 kHz, 0.09 percent. Rewind time, 2400 feet in 60 seconds. Timing accuracy, 0.10 percent.

17.230 Describe a continuous tape cartridge.—Continuous tape cartridges

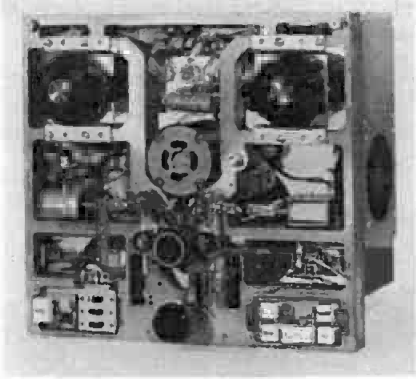


Fig. 17-229I. Interior view of 3M Dynatrack tape recorder with front panel removed.

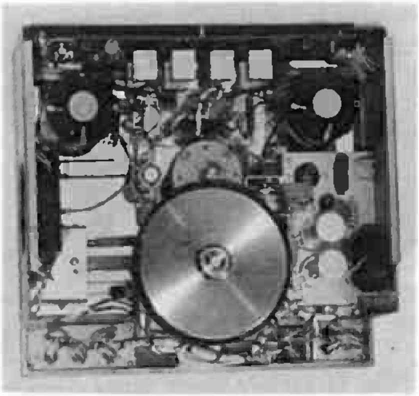


Fig. 17-229J. Rear interior view of 3M Dynatrack tape recorder.

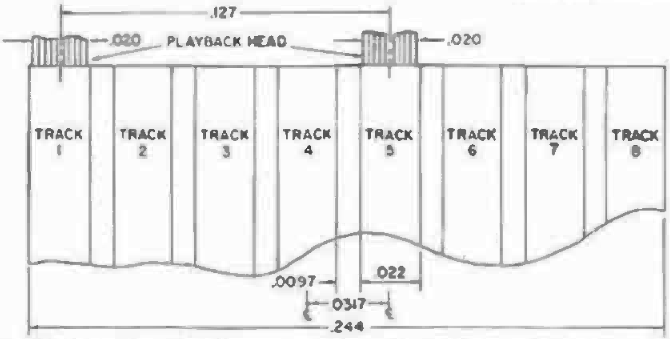


Fig. 17-230A. Sound track placement dimensions for Lear-Jet Corp. (Stereo Div.) eight-track stereophonic automobile tape reproducer.

are used in both home and commercial tape recorders employing a tape enclosed in a metal or plastic container. The cartridge is placed in a transport system which pulls the tape out of the cartridge, passes it over the reproducing head, and feeds it back into the cartridge much in the same manner as the loop box described in Question 17.208.

As a rule for the home use, the tape is designed for monophonic or stereophonic reproduction, using two or four sound tracks, as described elsewhere in this section. However, a continuous unit has been developed by Lear-Jet Corp., Stereo-Div., under the direction of William Brown, for use in automobiles. This unit uses eight sound tracks on a single ¼-inch tape, for stereophonic reproduction, at linear speeds of 3¾ ips. The dimensions for the track placement are given in 17-230A. Each track is 0.022 inch in width, with guard spacing of 0.0097 inch. The program material is recorded on a 1-mil backing imbedded with a special lubricant. Up to 400 ft of tape may be used, supplying 80 minutes of program material.

A series of rollers and guides align the tape when the cartridge is pushed into the transport system, which automatically turns on the motor, and aligns the tape with the head. In operation, the left head-gap (Fig. 17-230A) scans track 1, and the right head track 5. At the end of these tracks is placed a piece of sensing tape which causes a solenoid to rotate a cam ratchet to move the head to the right about 0.030 inch for scanning tracks 2 and 6. At the end of these tracks and the succeeding tracks, a sensing tape causes the head to move again to the right. At the end of tracks 4 and 8, the heads are returned to tracks 1 and 5. This action continues until the transport system is turned off. The flutter is approximately 0.20 percent.

Loudspeakers in an automobile are placed in the side doors to the right and left of the driver. The frequency response is the NAB Standard, for tape at 3¾ ips. An acoustic response made using a sound level meter is given in Fig. 17-230B, taken with the microphone of a sound level meter, placed at the position of the driver's head.

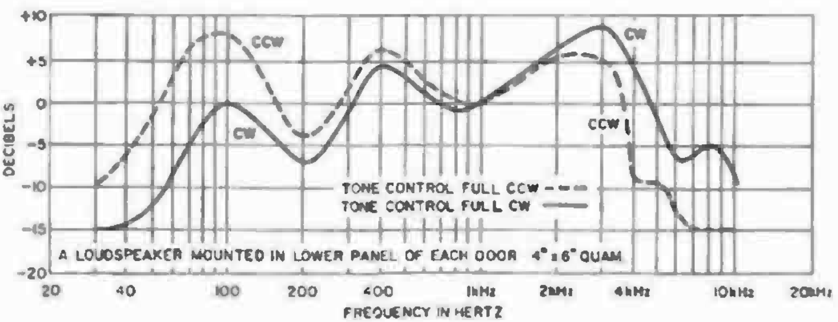


Fig. 17-230B. Acoustic sound-pressure measurements for two positions of the tone control for Lear-Jet stereo tape cartridge reproducer in an automobile. The microphone of the sound level meter was placed in the position of the drivers head.

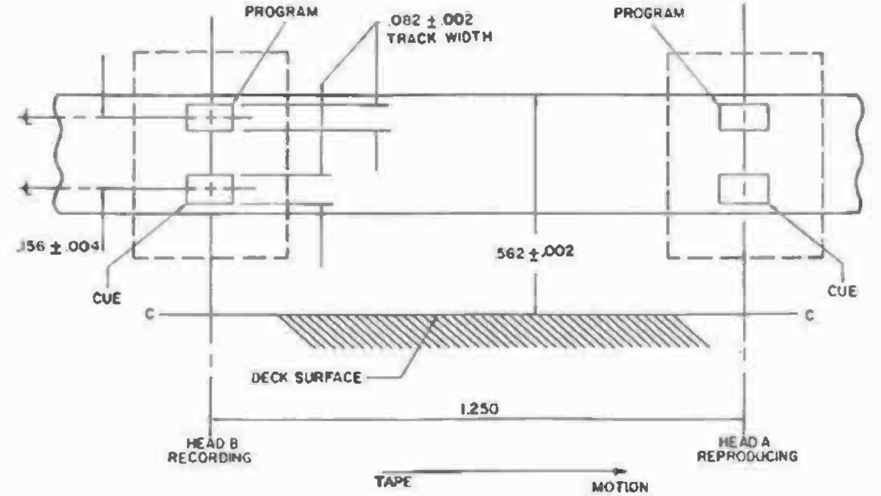


Fig. 17-230C. NAB Standard head placement for monophonic two-track head placement.

The NAB Standard, October, 1964, states that the head and sound track configuration for monophonic recording and reproduction will consist of two sound tracks, one for the program material and the other for a cue track. The cue track consists of a 1000-Hz tone, plus-minus 75 Hz, termed the primary cue tone. A secondary tone of 150 Hz, plus-minus 30 Hz, is a cue tone at the end of the program material. A third tone termed a tertiary tone, 8000 Hz, plus-minus 1000 Hz, is an auxiliary tone to be used for any other purpose desired. The length of the cue tone is to be 500 milliseconds, plus-minus 250 milliseconds.

The sound track placement dimensions are given in Fig. 17-230C. The upper track recorded by head B is the program channel, and the lower track is the cue tone channel. The upper head A is the program reproducing head, and the lower track is the reproducing cue channel.

For prerecorded stereophonic reproduction, three sound tracks are employed, as given in Fig. 17-230D. The system consists of two program tracks and a cue track. The upper track is the left program channel and the center track the right program channel, with the lower track the cue channel.

17.231 Describe the basic principles of a video-tape recorder.—Although video-tape recording is not actually a part of audio engineering, the sound engineer is sometimes involved with this type of recording; therefore, a basic knowledge of video recording is desirable. As it is beyond the scope of this work to attempt to describe in detail such equipment, the discussion will be confined to the very basic concepts.

Magnetic video-tape recording uses many of the principles of magnetic recording discussed throughout this section. Basically, a studio type video recorder employs a magnetic tape 2 inches in width, traveling at linear speeds of 15 or 7½ ips. A group of four magnetic heads, set in a wheel about 2 inches in diameter and 90 degrees apart contacts the tape and records a series of separate tracks on the tape as the tape moves in a linear direction away from the rotating heads (Fig. 17-231A).

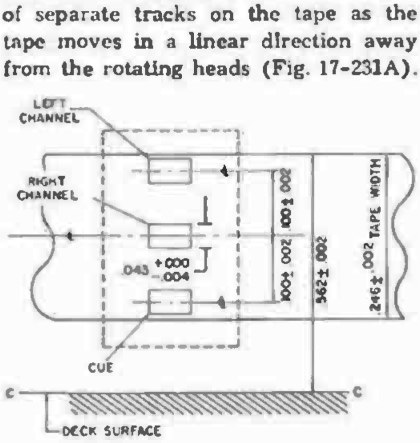


Fig. 17-230D. NAB Standard head placement for two-track prerecorded stereophonic tapes. If a second stereo head is used it is placed 1.250 inches from the first head.

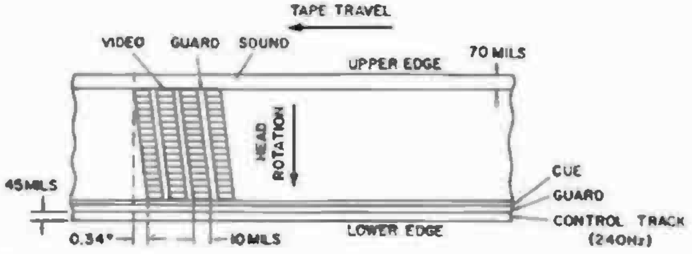


Fig. 17-231A. Track placement for magnetic video-tape recorder, using 2-inch tape.

The head assembly rotates at a speed of 240 rps, contacting the tape for about 120 degrees of arc, with tape being pulled at a linear speed of 15 ips. This takes $\frac{1}{240}$ second for 360 degrees and $\frac{1}{3}$ this time to traverse the 2-inch tape (120 degrees), or $\frac{1}{720}$ second. The head makes 960 sweeps per second across the face of the tape, during which time the tape has traveled 15 inches.

Because the tape travels 15 ips, it will move 0.02 inch or 20 mils while the head describes its arc across the face of the tape. Under these conditions, the bottom end of each track ends 20 mils later than the start, resulting in an

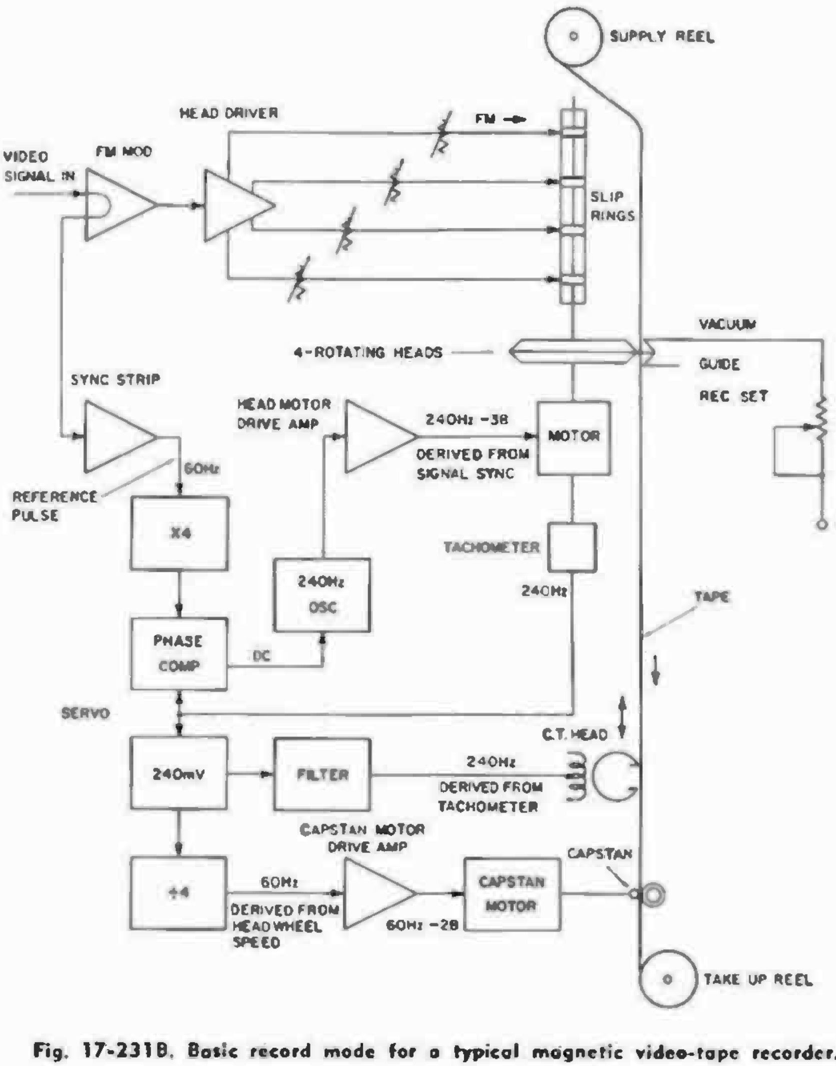


Fig. 17-231B. Basic record mode for a typical magnetic video-tape recorder.

angle of 0.34 degree. The linear speed of 15 ips and the heads rotating at 240 rps give an effective head-to-tape velocity of over 1500 inches per second. The head gaps are 0.10 mil in height. The width of the track laid down by each head is approximately 10 mils.

To achieve the wide frequency response required for video recording, a radio-frequency carrier, frequency modulated by the incoming video signal, is applied to the rotating heads. The audio signal is recorded on a 70-mil track in the conventional manner along one edge of the tape. In addition, a control-signal track is recorded on the upper edge of the tape, to assure that the heads will be in the exact position they were in the original recording.

The average signal-to-noise ratio for a 525-line picture is 37 to 40 dB; for sound, it is about 55 dB. The bandwidth required for 525 lines is 30 Hz to 4 MHz. Video-tape recorders may be designed to record both in monochrome and color. A basic diagram for a typical video recorder in the record mode is shown in Fig. 17-231B, with the reproduce mode shown in Fig. 17-231C.

Two views of an RCA TR22 studio video recorder are given in Figs. 17-231D and E. Fig. 17-231D shows the picture monitor and transport system, with a close-up of the head assembly shown in Fig. 17-231E. This machine is designed for both monochrome and color recording. Signal lights indicate the various modes of operation and functioning, with an automatic timing circuit which corrects electronically geo-

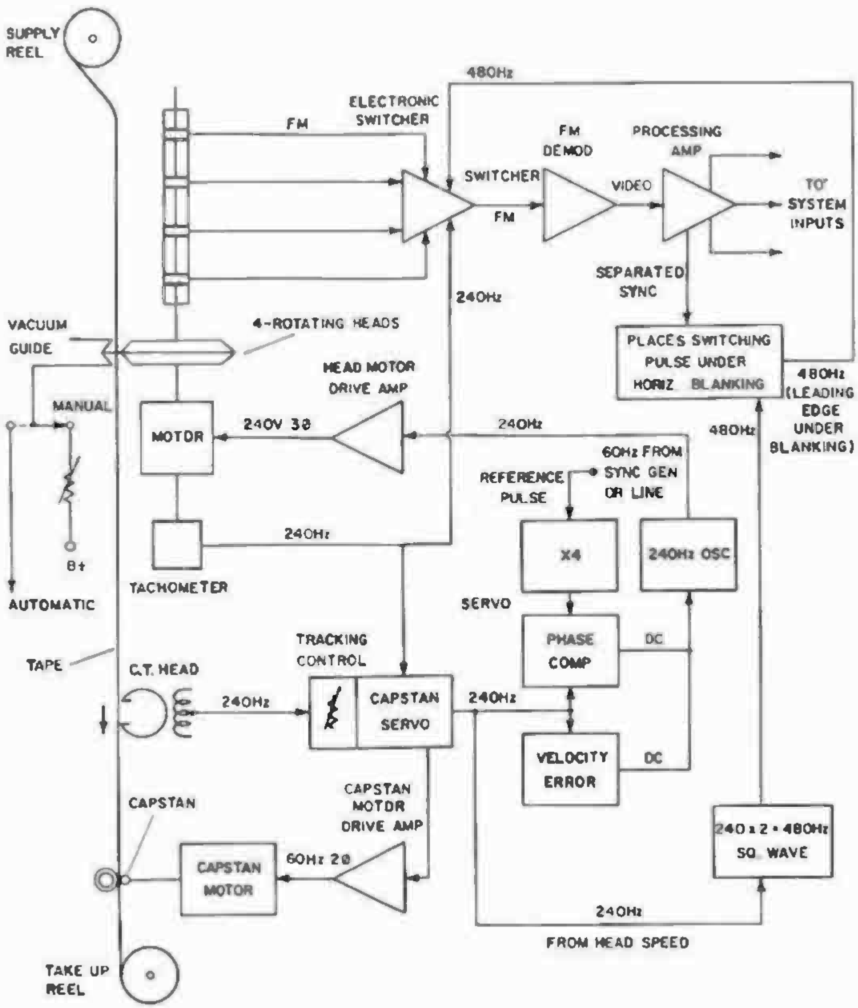


Fig. 17-231C. Basic playback mode for a typical magnetic video-tape recorder.

Fig. 17-231D. RCA Model TR22 video-tape recorder for both monochrome and color. This machine uses four rotating magnetic heads.

metric distortion (wiggly lines on the vertical or horizontal lines). Two transport speeds are provided—15 and 7½ ips. A system of electronic splicing is included, whereby segments of program material can be added to or inserted in previously recorded material without mechanically cutting the tape. Ninety-six minutes of program material may be recorded on one 7200-ft roll of 2-inch tape.

Fig. 17-231F, shows an Ampex Model VR-7000 portable video-tape recorder, using a 1-inch tape and a spiral transport system. The threading path and the transport system are shown in Fig. 17-231G. To attain a high tape-to-head ratio, a single head is used, set in a drum and rotated at 3600 rpm. The tape is wrapped around the rotating drum in an approximate helix of 3 degrees and moved at a linear speed of 9.6 ips. This combination of tape speed and rotating head results in a video recording speed of about 1000 ips. An audio sound track is recorded on the lower edge of the

Fig. 17-231E. Close-up of RCA TR22 transport system and head assembly.

Fig. 17-231F. Ampex Corp. Model VR-7000 video-tape recorder. This machine uses a 1-inch tape and a spiral transport system.

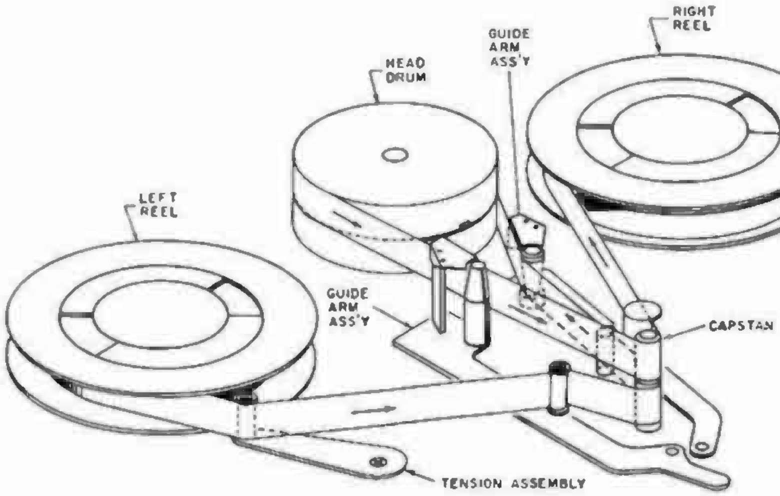


Fig. 17-231G. Tape-threading path for a spiral type transport system, as viewed from the front.

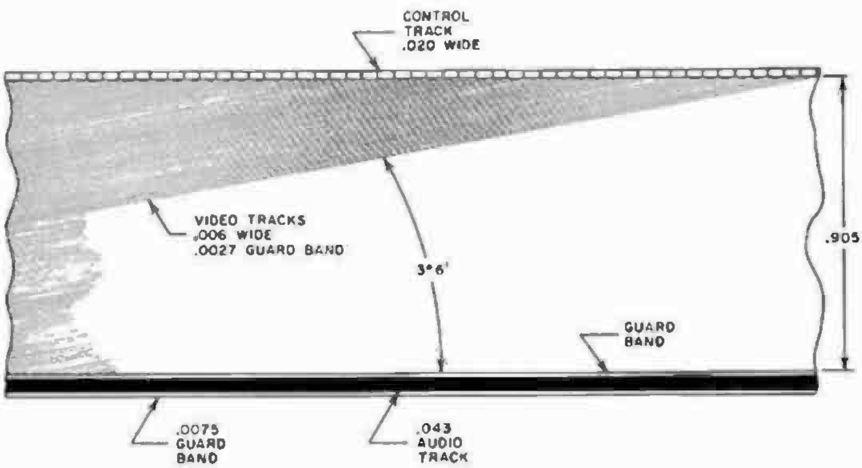


Fig. 17-231H. Scanning pattern for a video-tape recorder using one rotating head and a spiral type transport system.

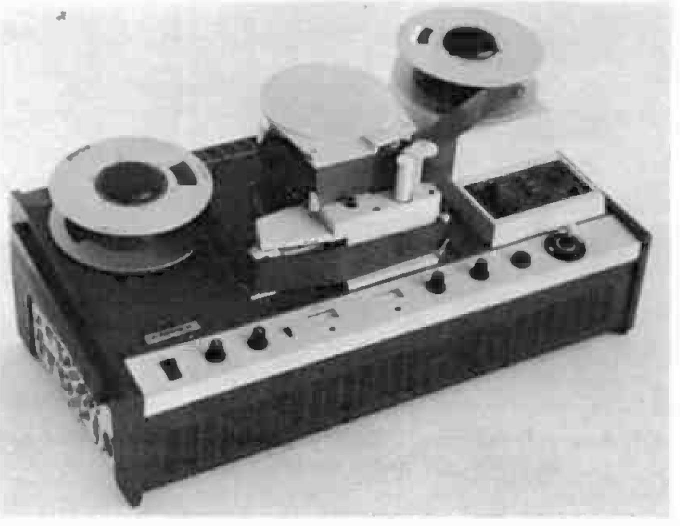


Fig. 17-231I. Ampex Model VR-660B video-tape recorder. This model uses a single magnetic head rotating at 3600 rpm.

tape, and a control track on the upper edge, to assure the proper alignment of the video head during recording and playback. The track scanning pattern is shown in Fig. 17-231H. The video bandwidth is 30 Hz to 3.5 MHz. A second portable machine, Model VR-660B, is

Fig. 17-231J. Ampex Corp. Model VR-1200 video-tape recorder. This machine uses four rotating heads and records both monochrome and color.

shown in Fig. 17-231I. This machine uses a 2-inch tape, with two rotating heads and a spiral transport system. This recorder has considerably wider bandwidth and many features approaching a larger studio model (Fig. 17-231J).

A home-type video recorder, manufactured by Sony, and marketed under the trade name of *Videocorder*, is pictured in Fig. 17-213K. This unit also includes a 9-inch television receiver, which is used for both monitoring and playback. A recording time of 60 minutes is afforded, using one-half inch tape, running at a linear speed of 7½ ips. This machine uses two rotating heads and a spiral transport system. The heads rotate at 1800 rpm with a scan length of 6½ inches. One head is used for recording, and two for playback. Provision is also made for an external camera.

17.232 What is a magnetic disc video recorder?—In construction, a magnetic video disc recorder appears somewhat like the conventional turntable for disc records. Machines of this type are employed in television broadcasting for instant playback of sporting events.

The technique used for recording video signals on a disc is one that has been used in the computer industry for some time. The recording medium consists of a 12-inch magnetically coated disc, rotating at 1800 rpm. The magnetic recording head is carried in a straight

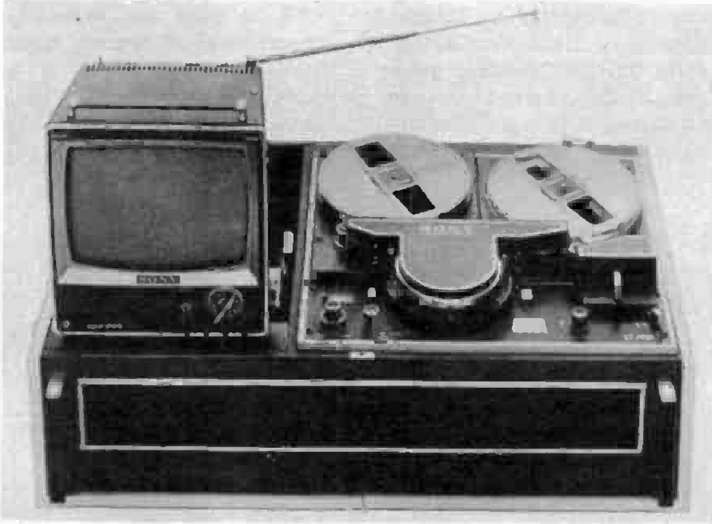


Fig. 17-231K. Sony *Videocorder* designed for home use. This machine uses a spiral tape transport system with two rotating heads.

line by a lead screw driven from the disc by a worm drive system. The worm drive has a ratio of 10 to 1; therefore, the lead screw rotates at 3 rps or 10 television frames per second. The track width is 0.004 inch with a guard band of 0.001 inch. Sufficient head travel is provided for a continuous recording of 20 seconds, or 3 inches. The direction of head travel is from outside to inside. For stop-action recording, an additional record-playback channel is used with a continuous circular track on the underside of the disc.

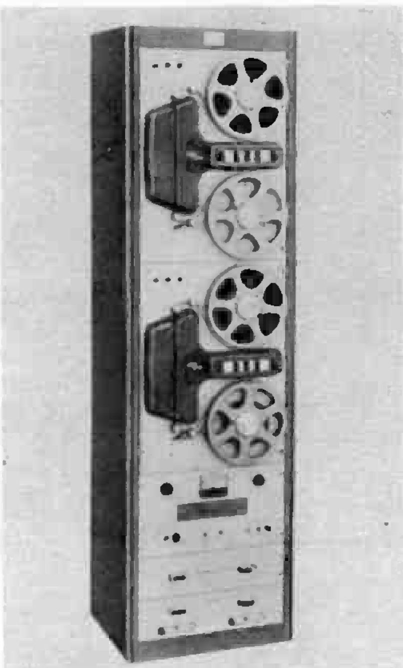


Fig. 17-233A. Magnasync - Moviola Model TR-1510 24-hour recorder-reproducer. This machine runs 1 ips using ½-inch magnetic tape.

A second machine employs two 16-inch discs rotating at 60 rps, and recording 60 fields per second (two fields equal one frame). The individual fields are recorded on tracks 7.5 mils in width, with a 2.5-mil guard band, while the recording head remains stationary. Four head assemblies are required for recording on both the upper and lower surfaces of the discs. The recording sequence is such that when head A completes a single field, head B begins recording the next field, and head A moves forward. When head B has completed its field, head C records, then head D, then back again to head A. Thus, each head records successive fields in sequence. Heads A and C record the odd numbered fields and B and D the even numbered. The heads are moved into position by a stepping motor attached to the head by a steel band. A rather elaborate control system enables the operator to record 30 seconds of play, and replay within 4 seconds at normal speed or variable slow motion.

17.233 Describe a multichannel slow speed communications recorder-reproducer.—The Model TR-1510, 10-channel record-reproduce magnetic-tape system, manufactured by Magnasync-Moviola Corp., is shown in Fig. 17-233A, with its block diagram given in Fig. 17-233B. This system has been especially designed to provide monitoring, recording, and reproduction on a 24-hour basis and is widely used in communication centers, aircraft bases, and in certain types of broadcasts, where a wide-frequency band is not a requirement. Recording, reproducing, and monitor amplifiers are solid-state

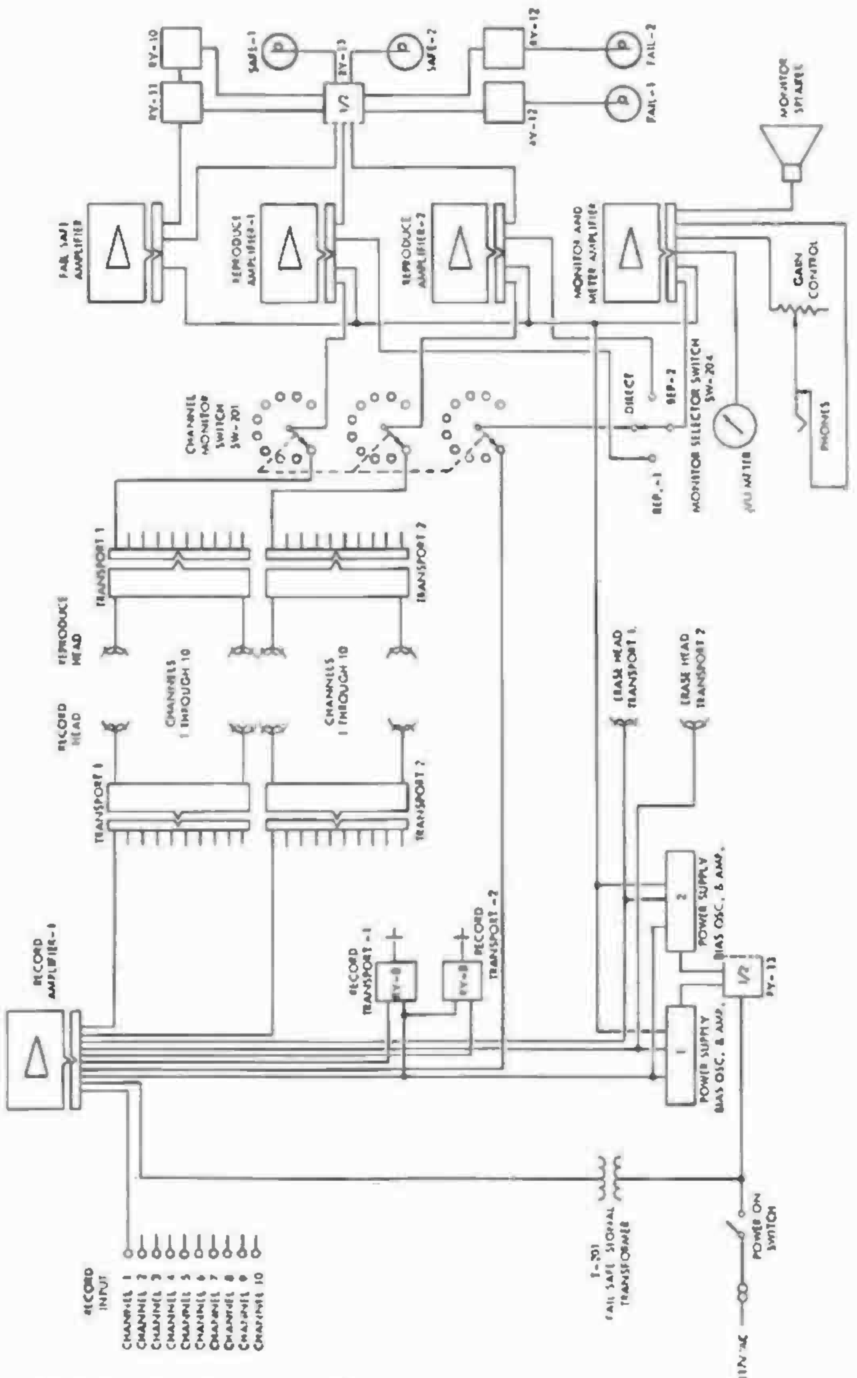


Fig. 17-233B. Block diagram for Magnasync-Moviola Corp. Model TR-1510 recorder-reproducer.

plug-in design, as are the power supplies and magnetic heads. Monitoring may be either by headphone or loudspeaker, and any one of the 10 channels (or a combination of two) is available. A rotary switch permits monitoring of the incoming signal, or the recorded signal approximately 2 seconds later.

An 0.5-mil base, 4800-foot reel of ½-inch magnetic tape, running at 1 inch per sec, will permit 16 hours of continuous record or playback time. A unique feature assures continuous operation in that it contains two identical transports. A control tone (below voice level) is recorded simultaneously with the program material. If this tone is not picked up by the reproduce head, it automatically switches on the second standby transport and its power. This function also occurs when the reel nears the end of the tape.

Specifications are: Input impedance 5000 ohms designed to bridge a 600-ohm line. Input level is —15 to plus 20dBm (continuously variable record level control on each channel). Dynamic range is 30 dB minimum. Cross talk is —25 dB minimum. Signal-to-noise ratio is 25 dB minimum. Frequency response is ± 3 dB from 300 to 2500 Hz for record-reproduce cycle. THD is 4-percent rms maximum. Bias frequency is 38 kHz. Flutter is 1-percent rms maximum. Start and stop time delay is maximum 1 sec from stop to operating speed, and 1 second operation to stop.

17.234 Describe a simple device for marking good takes while looping.—A simple marking device may be constructed by using a 500-Hz high-frequency battery-operated buzzer, and a 600/600 repeat coil. A 10,000-ohm resistor is connected in series with the output of the repeat coil to provide a high-impedance output. The output may be connected across the bridging bus or any other convenient position in the system. When the system is up to speed, a single tone is recorded ahead of each take. If the take is good, two tones are recorded at the end of the take for editorial purposes in locating selected takes.

17.235 Describe a method of increasing the signal-to-noise ratio by separating the audio spectrum into bands, in combination with compression.—Because of the masking effect in human hearing, low-level sounds are masked partially or completely by high-level sounds in the same frequency range. Thus, subjective signal-to-noise ratios in magnetic recording are better than would be indicated by an analysis of the playback waveforms. Tape noise depends on, and increases with, the instantaneous amplitude of the signal and an effect termed "modulation noise." Based on the masking effect, electronic devices have been developed for increasing the signal-to-noise ratio. One such system developed by R. M. Dolby, of the Dolby Laboratories, England, will be discussed.

By utilizing the masking principle together with signal compression and expansion, the Dolby audio noise reduction system achieves noise reduction by boosting the low-level signal components during recording, whenever possible, by compression. This is followed by complementary attenuation during playback, using expansion. The masking effect is introduced whenever the signal level is so high that compression and expansion are impracticable. Since masking is less effective with noise frequencies somewhat removed from the signal frequency, it is necessary to deal with the various portions of the spectrum independently. Such a system will then yield a lower apparent constant-nnise level, with the hush-hush and swish of the conventional compressor eliminated.

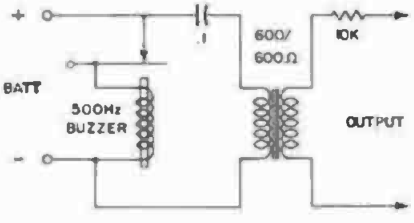


Fig. 17-234. Marking device for indicating a good take while looping.

The Dolby system splits the spectrum into four bands and compresses and expands each of these in an essentially independent manner. Separate bands are provided in the region of hum and rumble, midfrequency, medium-high frequency, and high frequencies. Thus, a high-level signal in one band cannot prevent noise reduction in a band in which the signal level may be low. The system produces a recording equalization characteristic which continuously conforms itself to the incoming signal in such a manner as to optimize the signal-to-noise ratio during playback.

The principal use of this instrument has been in magnetic recording to remove objectionable tape hiss; however, it is adaptable to any type recording system. The system is designed to have exactly 10-dB noise reduction. Thus, when used with a system of, say 60 dB signal-to-noise ratio, the ratio on playback is increased to 70 dB. As reproducing levels are sometimes held down because of the tape hiss, this method of recording permits the playback level to be increased without objectionable background hiss. Another advantage is that the original program material may be recorded at a lower level because of the additional gain of 10 dB signal-to-noise ratio on playback.

Referring to the basic block diagram (Fig. 17-235A), in the first two bands the noise reduction is 10 dB, while in the upper frequencies the noise reduction is 10 to 15 dB. Compression is achieved by raising the low-level signals and not by attenuating the high-level signals. The graph in Fig. 17-235B gives the characteristics for a conventional compressor where an input signal of 60 dB is compressed to 40 dB. Fig. 17-235C shows the Dolby system where the high-frequency signals are not compressed, and pass through the device unaltered in any manner.

Matching of the playback characteristics to that of the recording is accomplished by connecting identical networks with those used in the recording circuits in the negative-feedback loop of the amplifier system. In this manner, steady-state and transient effects are

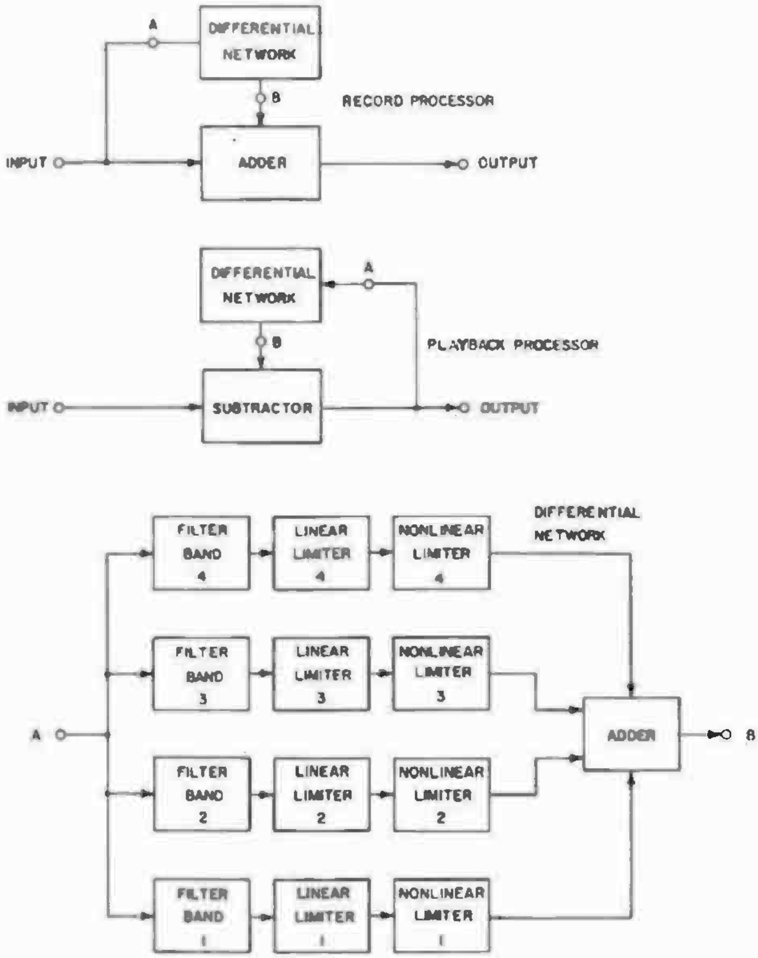


Fig. 17-235A. Block diagram for Dolby Laboratories (England) Model A301 audio background noise suppressor.

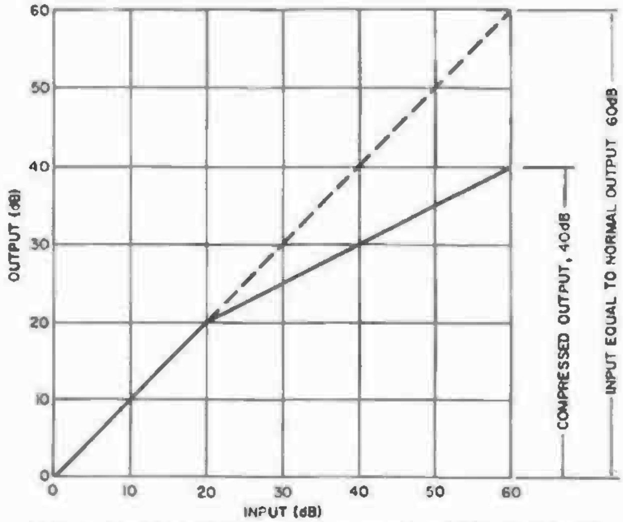


Fig. 17-235B. Conventional compressor characteristics.

automatically taken into account and the output signal becomes identical to the input signal. The use of separate bands is principally responsible for the elimination of hush-hush effects of compression, along with a special control signal, rectifying, and filtering techniques. By this method, low distortion with a 1-millisecond rise time is achieved. The compressor operation is fast enough that it is not heard by the ear during playback, even when the signal is keyed from peak amplitude to off.

Because of the compression characteristic, interchangeability of tapes and noise reduction units is permissible. An error of 3 dB in the matching of the record and playback level does not cause a perceptible difference in the frequency response or dynamics. It is claimed for this system that program material may be rerecorded repeatedly with no accumulated degeneration of the signal beyond that induced by the amplifier systems, which are on the order of 0.1 percent THD at their peak operating levels. The system may be used for both monophonic and stereophonic recording.

The instrument requires 99 silicon transistors and 163 diode rectifiers and is mechanically constructed using modular techniques. Its external appearance is pictured in Fig. 17-235D. Specifications are: Frequency range 30 to 20,000

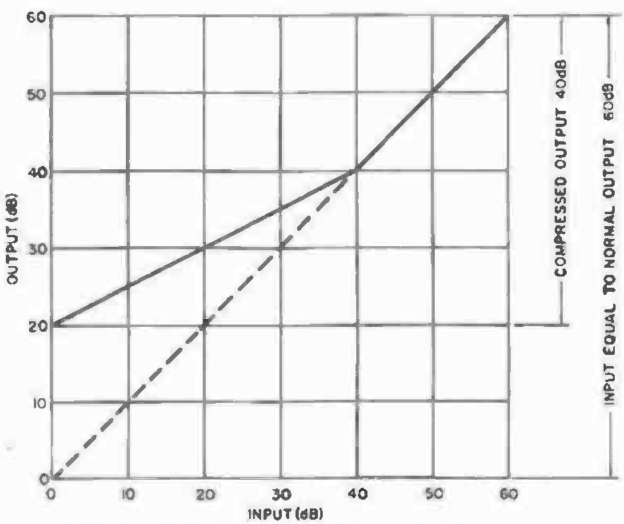


Fig. 17-235C. Dolby audio noise reduction system.

Hz plus-minus 1 dB, with a THD of 0.2 percent at an output level of plus 18 dBm. Noise reduction is 10 dB from 20 Hz, rising to 15 dB at 15,000 Hz. Unweighted internal noise is better than 80 dB below operating peak. Input impedance is 10,000 ohms bridging or 600 ohms matching. The reader is referred to the reference.

17.236 Describe the basic characteristics of chromium dioxide magnetic tape.—Such magnetic tape coating is a development of Photo Products Department of E. I. du Pont de Nemours Co., and marketed under the trade name *Crylon*. Chromium dioxide is an invented synthesis derived from chromium trioxide. It was developed by Dr. P. A. Arthur, Jr., of du Pont's central research department several years ago for the purpose of replacing the conventional iron oxide generally employed. The coating is manufactured by decomposing chromium trioxide in the presence of water at a temperature of 900°F under a pressure of 30,000 psi. (US patent 2,956,955.)

The coating is synthesized in the form of acicular, single-domain particles which can be varied from 4 to 400 microinches in length, having an aspect ratio of 10:1. Its coercivity can be varied from 25 to 700 oersteds, with a saturation flux density of 6100 gauss and a Curie point of 126°C. At the present time, the use of this tape has been confined to video recording and computer usage, although it can also be used for audio-frequency recording. Generally speaking, such tape has coercivities ranging from 360 to 530 oersteds with a residual flux density up to 1600 gauss, with low noise and print-through characteristics.

The coating thickness is on the order of 80 to 250 microinches, with about 200 microinches being employed for audio-frequency recording. Such tape has extremely better signal-to-noise ratio and response to short wavelengths at lower speeds than that usually employed for a given type of recording. (See Questions 17.28 to 17.30.)

17.237 What type magnetic test films are available?—Many of the magnetic test films are similar to those used for testing optical sound reproducers. Such films may be obtained from the SMPTE. (See Question 19.61.)

17.238 What is a pick-up magnetic film recorder?—It is a 16-mm or 35-mm magnetic-film recorder used in motion picture studios for rerecording. The term "pick-up recorder" implies that the master recorder may be stopped or started at will any time during the rerecording sequence. Insertions and deletions may be made as might be necessary for the final composite master sound track. The master recorder and reproducers along with the projector must be interlocked for absolute synchronization.

Included in the system is a control panel which permits the operator to erase, record, and reproduce in synchronization backward or forward. In this instance, the recorder is equipped

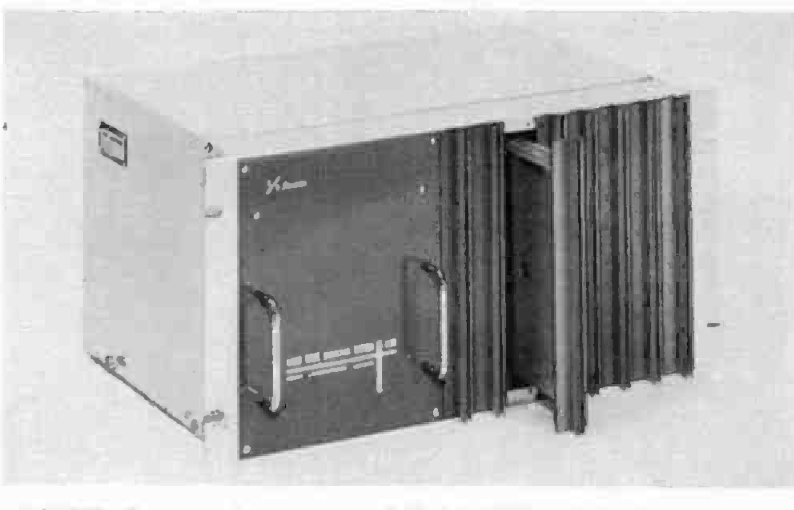


Fig. 17-235D. Front panel appearance of Model A301 audio background noise suppressor manufactured by Dolby Laboratories, England.

with an erasure and combination record-reproduce head.

Such recorders must be capable of running in reverse synchronization. In a well-designed system, no clicks or pops are induced in the sound track when reversing, stopping, or starting.

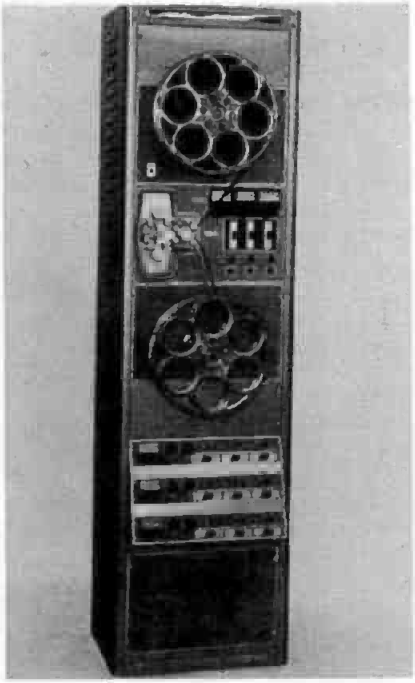


Fig. 17-238. Amego Model PIC-3 pick-up 16/35-mm recorder-reproducer.

The injection of the erase or recording bias current has no effect in the picking-up process. Such machines are quite convenient for recording post-synchronization foreign dialogue, as they are equipped with remote control for use by the dialogue director. Pictured in Fig. 17-238 is a Model PIC-3 pick-up recorder designed for studio use by Amega. (See Questions 17.224 to 17.239.)

17.239 ***Describe the basic principles of electronic video recording (EVR).***—It is a prerecorded television motion picture program system, developed by Dr. Peter Goldmark of CBS Laboratories, that employs motion picture film in a unusual manner. The film to be reproduced is contained within a cartridge similar to magnetic tape, in conjunction with an electronic reproducing system. The device is connected directly to the antenna terminals of a standard television receiver. The film is 8 millimeters in width, with two picture frames side-by-side (containing separate program material), each with its own magnetic sound track. No sprocket holes are used in the film base, the film being pulled by a pinch roller and capstan at a linear speed of 6 ips. Playing time is 26 minutes per picture strip. Controls are provided for starting and stopping, stop motion for viewing a single frame, and for selecting picture A or B.

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Section 18

Optical Film Recording

As early as 1881, an article appeared in the *Scientific American* describing the experiment in which Dr. Alexander Graham Bell transmitted sound over a modulated beam of light. By using a selenium cell, he converted the light to electrical energy. This achievement stimulated research, which in later years, was to result in photographic sound recording for motion pictures. However, it was not until after World War I that a concentrated effort was made to make optical film recording a reality, and many organizations and researchers became involved. Among them were, Thomas A. Edison, Lee DeForest, Theodore Case, E. C. Wente, E. W. Kellogg, J. Crabtree, G. L. Dimmick, E. I. Sponable, Frayne and Scoville, and many others.

While magnetic recording predominates in motion picture production recording and rerecording, the optical sound track is still widely used for release printing. With the development of new type sound recording emulsions and methods of processing film, great strides have been made in the sound quality obtained from optical sound prints. This section discusses many different types of optical sound recording equipment, with their associated components such as, galvanometers, light valves, film loss equalization, noise reduction, compression and limiting, density control, cross-modulation testing, the appearance of different types of sound track both old and new, frequency response, intermodulation testing, Standards of the industry, and other pertinent information pertaining to the art of photographic film recording.

Several early methods of recording and their equipment have been included for reference purposes and historical interest. Standards given are those in present usage; however, several are at present under review and may be changed in the near future.

18.1 What is a variable-area recording?—A photographic sound track in which the exposure is held constant while the area of the exposure is varied. This type sound track is used by both RCA and Westrex, although it is accomplished iu a different manner. The Westrex Corp. uses a light valve, while RCA uses a galvanometer. A typical variable-area sound track is shown at (a) in Fig. 18-29.

18.2 What is a variable-density recording?—A photographic sound track in which the area of exposure is held constant and the density of exposure is varied. This type sound track is used by Westrex. A typical variable-density optical sound track is shown at (1) in Fig. 18-29.

18.3 What is a striation?—A term used to identify the modulations of a variable-density sound track.

18.4 What is an Aeo Light recording?—A variable-density sound track made by means of a glow tube and used in the original Fox-Case Movietone camera in the early days of sound motion pictures. This system of recording is now obsolete. The outline of a typical Aeo Light tube is shown in Figs. 18-4A and B. (See Question 18.204.)

18.5 Describe a single system recorder?—A motion picture film recording system in which the picture and sound track are both exposed simultaneously on the same film base. Such systems are often used in newsreel cameras. The chief disadvantage of this

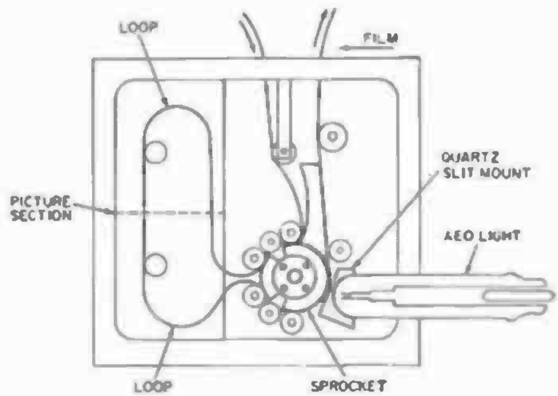


Fig. 18-4A. Bell and Howell single system camera (1928) fitted with Case Aeo light, for Fox Movietone Newsreel recording. This was one of the original methods of recording sound on photographic film. (Courtesy, Journal of the SMPTE)

system is that the sound track is developed with the picture and little latitude is available for proper control of the sound-track density. Newsreel cameras use both the variable-area and variable-density systems of recording. Modern newsreel cameras employ magnetic recording equipment. The sound track is recorded on a magnetic stripe, using either prestriped or post-striped film. (See Question 17.181.)

18.6 What is a double recording system?—A motion picture film recording system in which the sound track is recorded on a separate film in a sound camera or recorder. The camera and sound recorder are driven by a synchronous motor system. This method of operation allows a greater latitude in editing and cutting the picture. A special sound-positive recording emulsion is used on the film in the sound recorder and is developed in a special developing solution separate from the picture. Although the sound recording stock is called sound positive, it exposes as a negative; that is, when developed the image is black. (See Question 18.135.)

18.7 What type light is used for exposing optical sound track?—Both white and ultraviolet light are used. Originally, recorders used only white light. However, it was found that by using an ultraviolet filter in the optical train, the definition of the sound track was greatly improved.

18.8 How is ultraviolet light used with photographic film recording?—An ultraviolet (UV) filter is placed in the recording optical train just before the

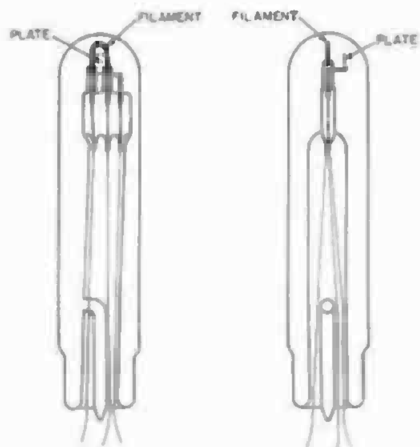


Fig. 18-4B. A Case Aeo Light recording lamp (1928) used in the Fox-Case recording system. (Courtesy, Journal of the SMPTE)

objective lens system. Before the advent of the modern fine-grain emulsions, ultraviolet light was used to obtain greater resolution of the photographic image with less background noise from the film. Ultraviolet light exposes the lower layer of the emulsion and reduces halation caused by white light. Many recording activities still use ultraviolet light. For ultraviolet recording, an additional 1-ampere of exposure lamp current is required.

18.9 Describe a sound head.—A sound head is the section of a motion picture projection machine that houses the phototube and magnetic sound reproducing components. Before the advent of magnetic recording, this name also applied to studio reproducing equipment. Generally, the machines consisted of a motion-picture projector transport system and a sound head similar to a projection machine, but without the lamphouse and picture head. Such machines were also referred to as dummies, meaning that they were dummy projection machines that contained only the sound head portion. As a result of the advent of magnetic recording, studio reproducing equipment is now installed in steel cabinets and is referred to as reproducers.

The modern studio reproducer consists of a transport system, magnetic head, and if it is of the mag-optical type, a phototube and exciter lamp are included for reproducing photographic sound tracks. In addition, a motor-driving system and automatic rewinds are included. When the machines are used for rerecording purposes, they are fitted with selsyn interlock motors which are driven from a selsyn interlock distributor system, described in Question 3.49. Typical studio rerecording equipment is shown in Figs. 17-178A to E, and 17-208E.

18.10 What is a film phonograph?—It was a name originated by RCA for their original photographic film recording machines that were also used for reproducers. Now obsolete.

18.11 What is a recording galvanometer?—It is a light modulator used by RCA and other companies for recording photographic sound track.

18.12 Describe the construction of a recording galvanometer.—A front view of a variable-area recording galvanometer—developed by G. L. Dimmick of RCA, and used in their photographic film recorders—is shown in Fig. 18-12A. The movement consists of an iron-nickel alloy armature A, clamped between two nonmagnetic shims B and BB, and held in place by pole pieces C and CC. Two Alinco magnets D and DD

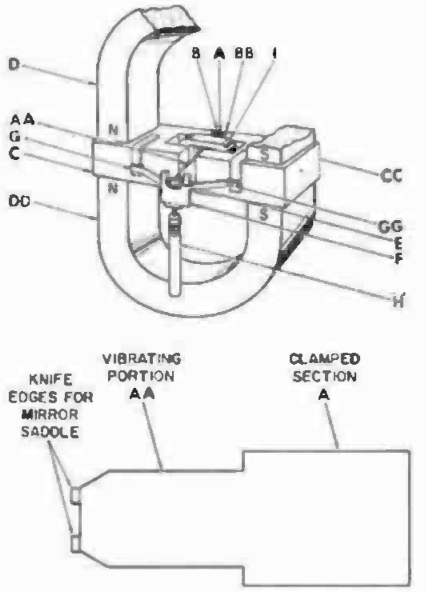


Fig. 18-12A. Recording galvanometer Model MI-10751 manufactured by RCA and used in their photographic film recorders. (Courtesy, Radio Corp. of America, and Journal of the SMPTE)

supply the magnetic field. A small gap for the upper end of the armature AA to vibrate in is left between the upper pole pieces. The top edge of the armature is ground to a knifelike edge, as shown in the insert in Fig. 18-12A. The knife edge mates with a mirror saddle E, which supports a 0.10-inch square front-surfaced mirror F. The saddle is

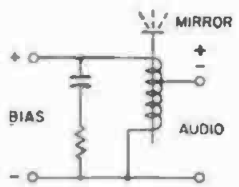


Fig. 18-12B. Internal connections for a single-coil type galvanometer.

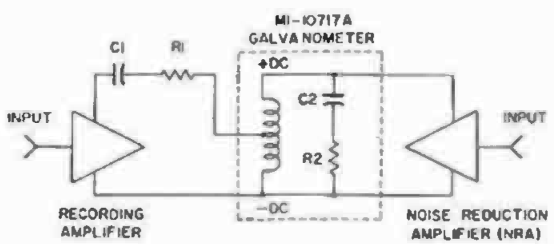


Fig. 18-12C. Connections for a recording and noise-reduction amplifier (NRA) to a single-coil galvanometer.

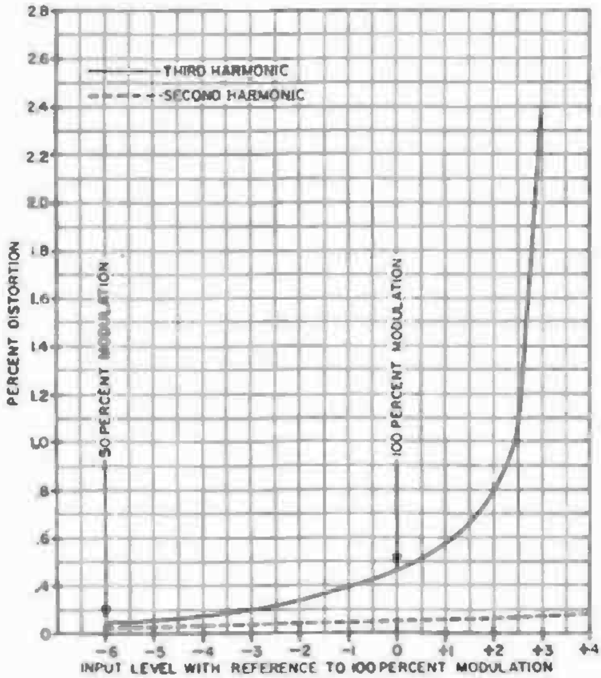


Fig. 18-12D. Second and third harmonic distortion for RCA recording galvanometers.

held in place by a phosphor-bronze bridge assembly G and GG. A tungsten-loaded neoprene line H is attached to the saddle plate, with the other end free. As the mirror is rotated by the audio signal, the neoprene line is deflected in torsion. At frequencies between 5000 and 10,000 Hz, a considerable amount of the energy is transmitted through the line and when reaching the end, is reflected back toward the mirror, but is dissipated before reaching the mirror. Neoprene rather than conventional rubber is used for the damping line because it has a higher coefficient of damping and is less affected by temperature changes. It is loaded with finely powdered tungsten which is heavy and inert.

The modulating coil I consists of a single winding, with a tap for the audio signal. The noise-reduction current is applied to the complete coil. The design is such that the use of an impedance-matching transformer is not required between the output of the recording amplifier and the galvanometer. A resistor in the amplifier builds out the impedance. A 10-μF capacitor is connected in series with the modulating coil to prevent the dc bias current from the noise reduction amplifier (NRA) from entering the output circuit of the recording amplifier. The circuit for the modulating coil in Fig. 18-12B shows the tap for the audio signal. A capacitor and resistor in series are connected across the whole winding to partially neutralize the inductance of the modulating coil in the frequency range of

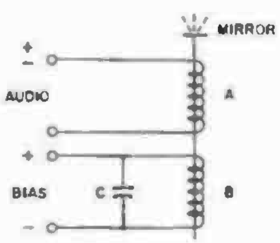


Fig. 18-12E. Internal connections for a two-coil galvanometer.

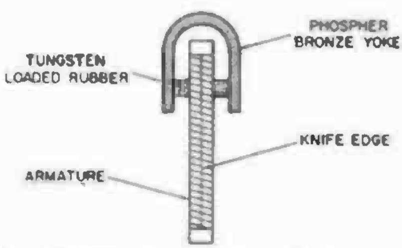


Fig. 18-12F. Top end view of the armature in an RCA recording galvanometer showing the damping assembly attached to the side of the armature.

3000 Hz. The basic circuit for driving the galvanometer and applying noise-reduction bias current is given in Fig. 18-12C.

The operation of a galvanometer is now explained. With the armature at rest and accurately centered in the magnetic field, the mirror is held at a right angle to the armature by the saddle and bridge assembly. Application of an audio-frequency current to the modulating coil causes the armature to vibrate between the pole pieces in accordance with the audio-frequency currents, because of the reaction of the armature to the fixed magnetic field. Since the armature is clamped at its lower end and can only vibrate at the upper end, the knife edge transmits a rotary motion to the mirror saddle, which causes the mirror to be deflected in a direction opposite to the armature motion. With a sine wave applied to the coil, the armature will have equal displacement from side to side. Thus, light falling on the mirror surface as it vibrates is displaced in a like manner and is proportional to the amplitude of the applied signal. By means of the mechanical and electrical damping, a fairly uniform frequency characteristic is obtained.

During periods of low or no modulation, the noise-reduction current through the coil deflects the armature to one side, causing the mirror to be deflected to a previously determined point. This permits a small amount of exposure light to reach the film and creates the bias lines discussed in Question 18.79 and illustrated in Fig. 18-290. A galvanometer biased by a noise-reduction amplifier is termed a biased galvanometer.

A typical frequency-response characteristic is given in Fig. 18-13. It will be noted that the response is within plus or minus 1 dB with reference to 1000 Hz, ranging from 30 to 9000 Hz, and down approximately 5 dB at 10,000 Hz. It should be remembered that motion picture photographic recording rarely exceeds 8000 Hz. Typical harmonic distortion characteristics are given in Fig. 18-12D, plotted percent distortion versus second and third harmonic distortion, for 100-percent modulation of the galvanometer. It will be noted that the third harmonic (the most important one) has less than 0.5-percent distortion. Even-order distortion (second harmonic) is held down by the mechanical balance of the magnetic structure and damping, and also by the use of push-pull circuitry in the recording amplifier. Second-harmonic distortion for any frequency between 100 and 6000 Hz never exceeds 0.6 percent; harmonic distortion never exceeds 0.2 percent. With the total harmonic distortion (THD) of not more than 2.5 percent, the average driving power required for 100-percent modulation is approximately plus 24 to 28 dBm, depending on the type of galvanometer used.

In the two-coil galvanometer (Fig. 18-12E), the general construction is similar to the single-coil type, except that separate coils (A and B) are used for the audio current and bias current. In addition, mechanical damping is applied to the upper portion of the armature by means of a U-shaped phosphor-bronze clamp and two tungsten-loaded rubber dampers, centered one to each side of the armature (Fig. 18-12F).

At the lower frequencies, the whole damping assembly vibrates with the armature and is ineffective. As the higher frequencies are approached, because of the inertia of the clamp, the damping assembly tends to stand still; however, the armature continues to vibrate the clamp, compressing the tungsten-loaded damping pads and reducing the resonant peak at 9000 Hz. Without the damping device, the resonant peak is about 12 dB, and with damping it is reduced roughly 3 dB. To further reduce the peak and smooth out the characteristic, a capacitor of 0.028 to 0.038 μF is connected *across* the bias-current coil (Fig. 18-12E), to neutralize the inductive reactance of the modulation coil.

Each galvanometer is individually tested and a value of capacitance is selected that will result in the smoothest response to frequency. The frequency characteristic for the two-coil galvanometer is similar to that of the single-coil type.

Different type sound tracks may be recorded using the same galvanometer. All that is required is to change the image mask in the optical train to produce the desired image. (See Question 18.77.) Galvanometers used in the first variable-area recording systems were called vibrators, and were designed

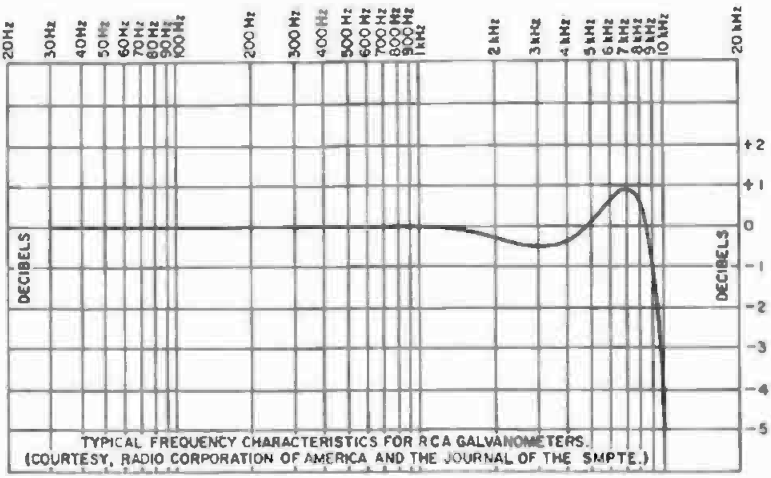


Fig. 18-13. Typical frequency characteristics for RCA galvanometers. (Courtesy, Radio Corporation of America, and Journal of the SMPTE)

similar to the Dudell oscillograph shown in Fig. 22-75.

18.13 What are the frequency characteristics for the galvanometers in Question 18.12?—Typical frequency characteristics for both single and two-cell types are given in Fig. 18-13. The frequency characteristic is quite uniform, and there is only a plus or minus variation of 1 dB between 30 and 9000 Hz. Since optical film recording rarely exceeds 8000 Hz, this is quite satisfactory.

18.14 Describe a light valve.—It is a light modulator, invented by E. C. Wente of the Bell Telephone Laboratories in 1922, used for recording sound on photographic film. Snch devices may be used for recording both variable-area and variable-density sound tracks. Light valves are used by Westrex in their photographic film recorders.

18.15 Describe the construction of a light valve.—The construction of an early model Western Electric (now Westrex Corp.) light valve is shown in Fig. 18-15A. The light-modulating mechanism consists of two Duralumin ribbons A, approximately 6 mils wide by 1 mil thick. The loop end is wrapped around a grooved button B, and tension is applied by spring C. The free ends of the ribbons are attached to a small windlass D, which is used to apply tension to the ribbons and enable them to be tuned to a frequency above the normal recording range. The center portions of the ribbons are held at a fixed distance of 1 mil by a bridge E, forming a slit 1 mil in height.

The whole assembly is mounted at right angles to and in a strong magnetic field supplied by an electromagnet. Behind and in line with the slit formed by the spacing of the ribbons is mounted an exciter lamp and optical system for exposing the sound track on the moving film. Althongh light valves are generally employed for recording variable-density sound tracks, they may also be used for recording variable-area sound tracks with special considerations. An

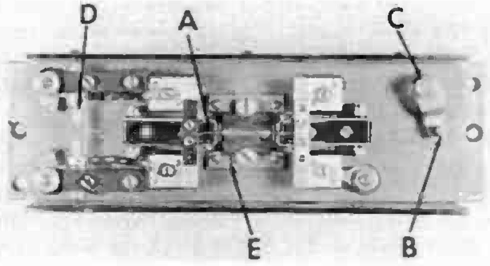


Fig. 18-15A. An early model (1930) Western Electric light valve for recording variable-density sound tracks.

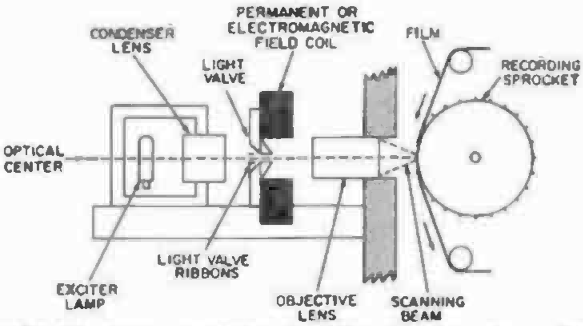


Fig. 18-15B. The basic principles of a film recorder using a light valve.

elementary light valve and optical system schematic diagram for variable-density sound recording is shown in Fig. 18-15B.

The basic concept of a light valve is that of a device having an air gap in a magnetic circuit in which one or more ribbons are mounted under tension. A signal (audio frequency) applied to the ribbons will move the ribbons in accordance with the theory of a conductor carrying current in a magnetic field. The movement of the ribbons causes the light-slit formed by the separation between adjacent ribbons to vary in height (zero to 1 mil) thereby permitting more or less light from the exciter light to fall on the moving film. Thus, a light valve may be considered to be a light-modulating device.

The ribbons in the light valve are so polarized with respect to the external magnetic field, that their reaction in the magnetic field causes them to open and close, depending on the instantaneous polarity of the applied audio-frequency signal voltage. Thus, in this manner the amount of light falling on the film is controlled, causing the density of exposure to be varied. Hence, the term "variable density."

The appearance of the recorded sound track is similar to that shown at (1) in Fig. 18-29.

Modern designed light valves employ ribbons operating on the foregoing principle; however, they now use permanent magnets rather than the electromagnet used in early model recorders as shown in Fig. 18-15B.

A Westrex Corp. two-ribbon permanent-magnet type light valve is shown in Fig. 18-15C. It records a single variable-density should track similar to that seen in (1) in Fig. 18-29.

The light valves may be used for recording either 16-mm or 35-mm sound tracks, the only difference being the dimensions of the masks which provide the proper slit width.

The Westrex Corp. light valve has been designed to have a low circuit resistance and an air gap of high flux density. This combination produces a circuit "Q" of 2.0, with a resulting resonant peak of about plus 5 dB at the resonant frequency with reference to 1000 Hz. A low circuit resistance is obtained by shunting each ribbon with a resistor approximately equal to the ribbon resistance.

The high flux density is obtained by designing the light valve so that it may be magnetized as a complete assembly and by the use of special materials. When magnetized, a light valve is ready for use with three major sections strongly bound together by the force of the magnetic section. While in this condition, the unit cannot be adjusted until it is removed from service and demagnetized. Magnetizing the assembly as a whole reduces the losses which occur when a portion of the magnetic circuit is broken. The special materials used are: Alnico V for the permanent magnets, and Permendur for the pole pieces. Alnico V permits a high magnetomotive force to be secured for the magnetic circuit, while the Permendur pole pieces permit high flux saturation. When demagnetized, the light valve separates into three major pieces: the magnetic section (Alnico V), the cap pole piece, and base pole-piece assemblies.

The details of the construction may be seen by referring to Fig. 18-15C. The magnet section A is cut away on the inside to provide proper space for the two pole piece assemblies. The cap pole piece M contains a glass cover P and a mask Q. Two mask lengths are available. These are: 0.212 inch, for 35-mm film recording; and 0.176 inch, for 16-mm film recording.

The base pole piece N is most clearly showu in the view without the cap pole piece. In this view may be seen the ribbon clamp carriages J used to clamp, space, and stretch the ribbons C, ribbon shunts F, and the other pole pieces. Attention is called to the cover glass S in the base pole piece section, which is shown as having an appreciable thick-

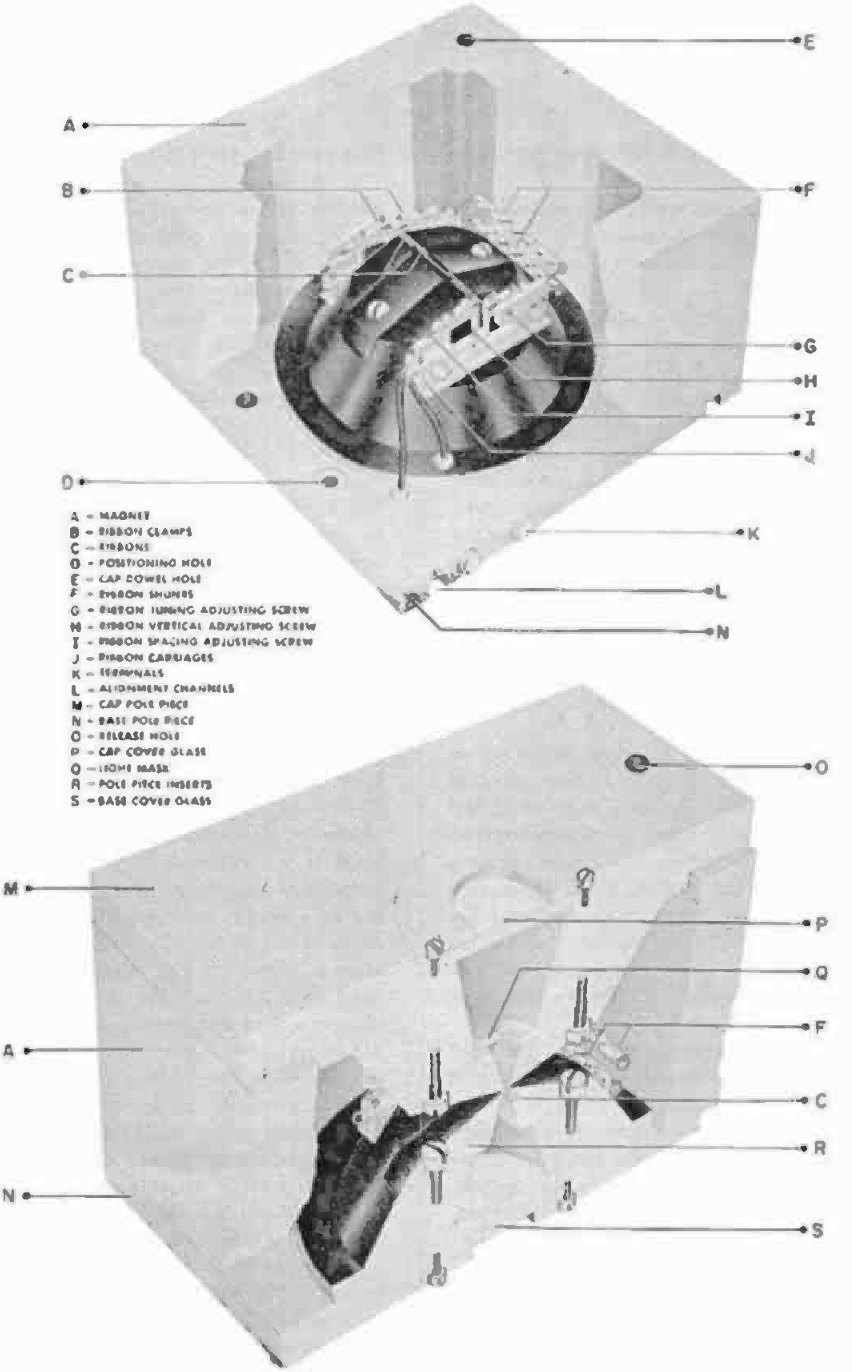


Fig 18-15C. The Westrex Corp. two-ribbon light valve used for recording single variable-density sound tracks.

ness. This provides the light valve with the same glass path as a light valve used for recording push-pull sound tracks (to be described later in this question), thereby eliminating the need for refocusing each time the light valves are interchanged in a recorder.

Light valves ready for use contain two short sections of Duralumin ribbons approximately 0.50 mil thick by 6 mils wide, mounted in the base pole-piece assembly by means of four clamping pieces B. Spacing screws I are used to properly center and space the ribbons. Tension is applied to each ribbon by means of a tuning screw G until resonance is obtained at 8500 Hz. The resulting slit formed by the parallel edges of the ribbons spaced 1 mil apart provides a light slit which is limited in length by the mask Q. As previously mentioned, the mask is available in two lengths, 0.212 inch for 35-mm sound tracks and 0.176 inch for 16-mm sound

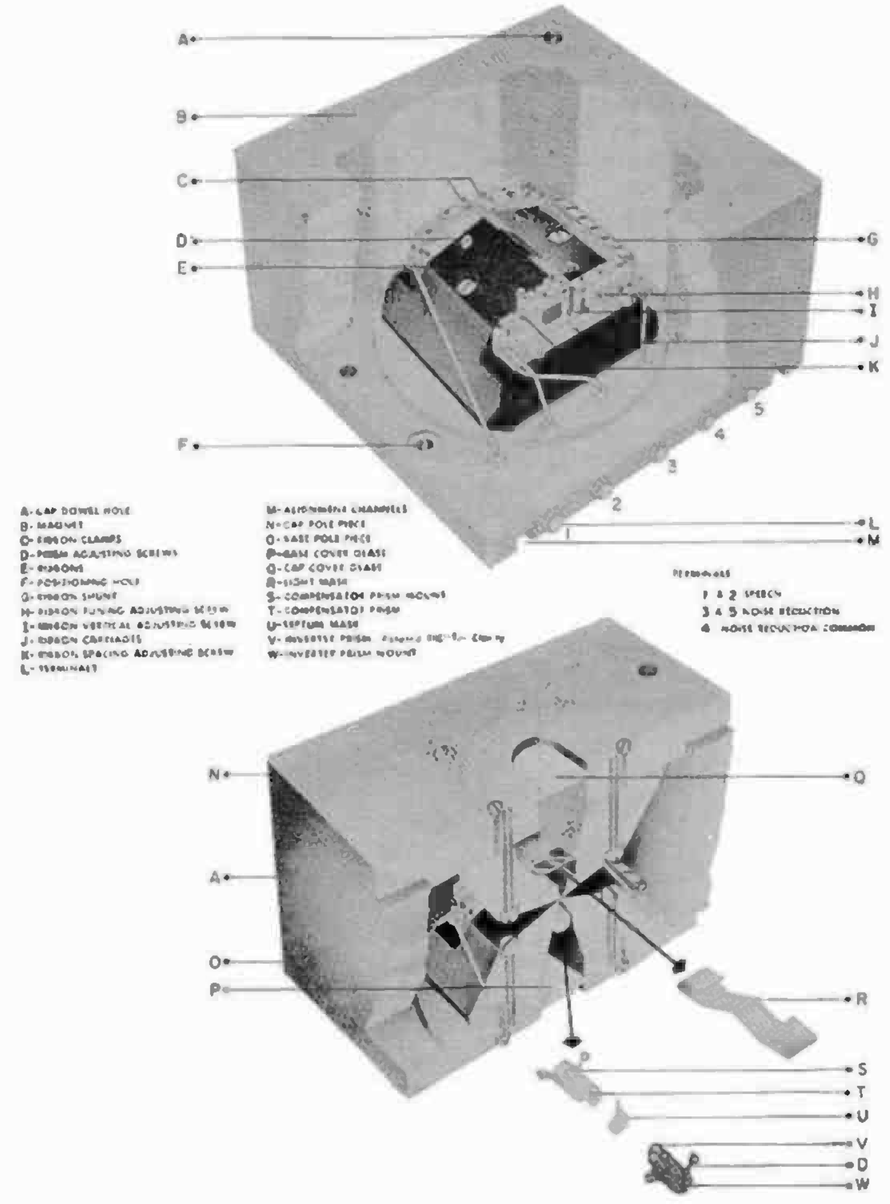


Fig. 18-15D. The Westrex Corp. 3-ribbon light valve used for recording push-pull variable-density sound tracks.

tracks. Item O is one of two threaded release holes on each end which are used to separate the tight fitting pieces after the light valve has been demagnetized.

When assembling the light valve, the three major pieces are aligned by the snug fit of the end pieces aided by single dowel pins in each piece which fit into holes in the magnetic section.

Because of the rise in frequency response at the resonant frequency of the ribbons, a special equalizer is used to attenuate the rise and provide a flat frequency response. The recording circuit connections called a Simplex circuit, are shown in Fig. 18-19.

The term "mask" used throughout the foregoing discussion should not be confused with the term "mask" in Question 18.77, used with variable-area film recording galvanometers, as they serve two different purposes.

Specifications for a typical two-ribbon light valve are: Resistance (including the shunt resistors), 0.55 ohms; power required for 100-percent modulation at 1000 Hz using the Simplex circuit (Fig. 18-19), plus 9 dBm; noise-reduction current, light valve 135 mils, into Simplex circuit 270 mils; ribbon spacing, 0.0010" plus or minus 0.00010"; ribbon tuning frequency, undamped 8500 Hz; resonant rise, 5 dB at 8000 Hz. (The term "undamped" means the frequency of resonance before the complete assembly is magnetized.)

A Westrex three-ribbon light valve using permanent magnets designed for recording push-pull, variable-density sound tracks is shown in Fig. 18-15D. The audio signal is impressed on the center ribbon. The two outer ribbons are connected in series to move toward the center ribbon in response to the noise-reduction current. Valving action takes place between each of the two edges of the center ribbon and its adjacent noise-reduction ribbon.

Certain inherent advantages are obtained by the use of a single ribbon for the signal currents (audio signal) as compared to previous multiribbon types. These advantages are: (1) increased stability of resonant tuning achieved by eliminating the bias current (noise reduction) with its resulting heating effect on the signal current ribbon. (2) reduced resistance for the signal current, since a Simplex circuit is not required for the noise reduction amplifier circuit. A low resistance results in a low resonant peak amplitude. (3) the outer ribbons may be tuned to a lower frequency than the signal ribbon, thereby reducing not only the required amount of bias current, but also the corresponding heating effect.

The materials used are the same as for the two-ribbon light valve. The frequency response rises to 5 dB at the frequency of resonance, referred to 1000 Hz. A low circuit resistance is obtained by shunting the single ribbon with a resistor approximately equal to the ribbon resistance.

A push-pull type sound track is obtained from two well-defined light slits

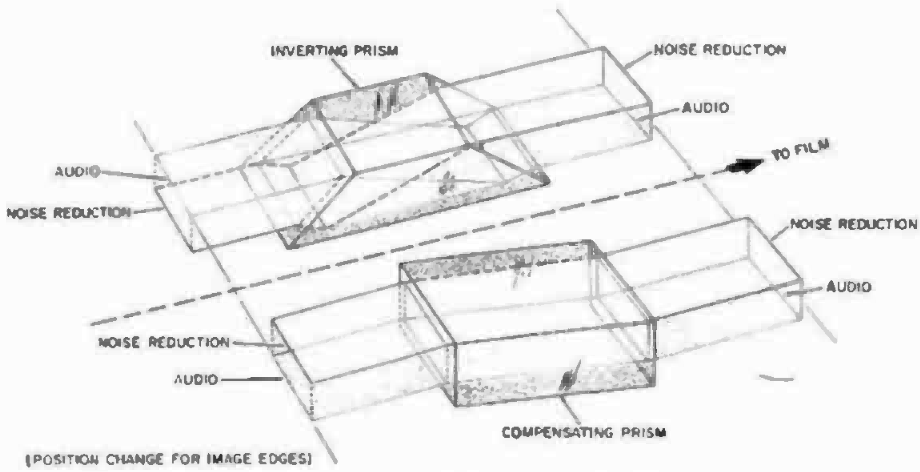


Fig. 18-15E. Schematic diagram for optical prisms used with Westrex Corp. push-pull three-ribbon variable-density light valve.

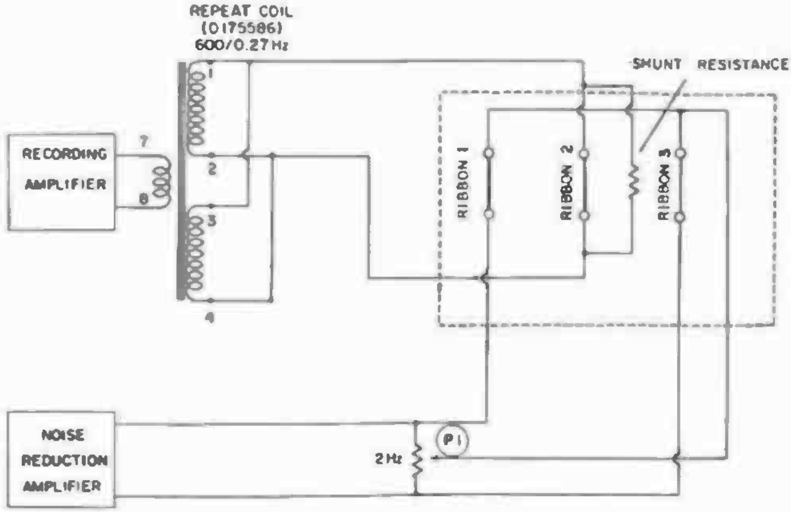


Fig. 18-15F. Connections for Westrex Corp. RA-1238 push-pull three-ribbon variable-density light valve.

which operate in opposite phase; that is, as one slit closes, the other opens. This action tends to cancel even-harmonic distortion resulting from the film velocity effect (see Question 18.22), in-phase noise reduction frequencies, etc. As an additional advantage, a 3-dB increase in the signal-to-noise ratio is obtained from a double-width push-pull sound track as compared to a single-width sound track.

The mechanical design of a three-ribbon light valve results in two apertures which are physically offset, with the bias and signal edges for each aperture occurring on opposite sides. An inverting prism is used to secure the optical inversion of one aperture image, thus bringing the bias and signal edges to common sides for the two images.

The other aperture image is then corrected for focus by means of a compensating prism which increases its light path to that required for the inverting prism. Alignment of the images is obtained by tilting the compensating prism. The functioning of the two prisms is shown in the optical schematic of Fig. 18-15E.

Referring to the cutaway drawing in Fig. 18-15D, the major pieces are the magnet piece B, the cap pole piece N, and the base pole piece O. The three major pieces are aligned by the close fitting of the parts and the single dowel pin in each end piece. The cap pole

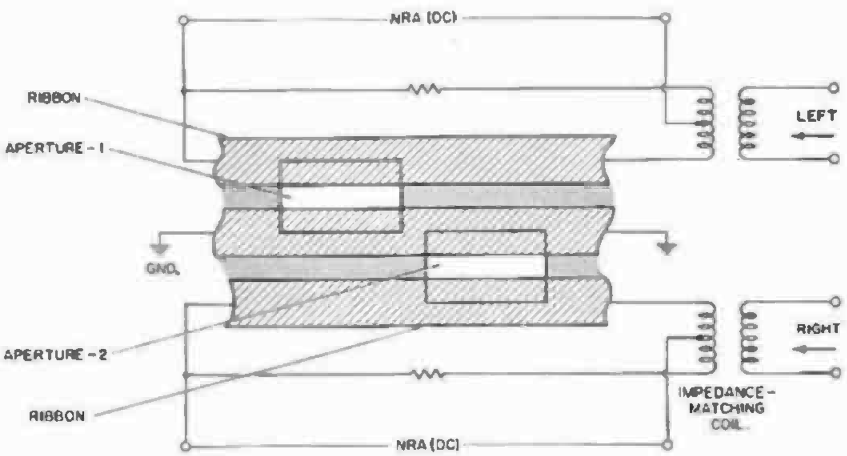


Fig. 18-15G. Westrex three-ribbon light valve for stereophonic variable-density photographic sound track recording.

piece N contains the glass cover Q to keep out dirt, and the mask R which provides the limiting edges to the light aperture formed by the ribbons.

Each mask aperture measures 0.125 inch × 0.209 inch with a separation of approximately 0.022 inch between the two ends. The permanent magnet B is cut away on the inside to provide the necessary clearance for the two pole pieces and ribbon clamp carriages J. A ribbon clamp carriage is required for each end of a ribbon. This device provides clamps C for the ribbons E, adjustment screws for spacing K, tension H, and elevation I. Only one tension screw and block are provided per ribbon. Terminals L provide connectious to the ribbons as follows; the center ribbon (audio signal) is connected to terminals 1 and 2. Terminals 3 and 5 (the noise reduction current) are connected to the noise reduction ribbons, with the middle terminal 4 serving as a common connection to the junction of the two ribbons. The shunt resistor for the center ribbon is item G.

Directly below the ribbons on the base section may be seen the prism mounting consisting of two cylinders, each containing a prism. Enlarged views of both cylinders are shown directly below the sectional view. Cylinder W contains an inverting prism V which transposes the image edges for one aperture. The compensating prism T in the other cylinder S provides the same glass path for the second image so that the two images are in common focus. Alignment of the edges is secured by rotating the compensating lens to obtain a slight displacement of the image.

Adjustment and locking screws for the cylinders are shown as item D. A septum mask U is placed between the two mountings. One of the two alignment holes in the base pole piece which is used to mount the light valve on a modulator unit is shown as item F.

The ribbons consist of three pieces of Duralumin measuring approximately 0.5 mil × 6.5 mils. These ribbons are located with respect to the alignment holes and spaced under tension to form two apertures 1 mil in width. The tension is the greatest for the center ribbon (audio signal), which is tuned to resonate at approximately 8500 Hz, while a lesser amount of tension is applied to the outer ribbons (noise reduction) to secure a resonant frequency of approximately 5500 Hz. When the complete assembly is magnetized the resonant frequencies are shifted downward to about 8200 and 5300 Hz respectively.

The three ribbons are adjusted in elevation so that the signal ribbon and the outer ribbons are in different planes, separated by approximately 1.5 mils. This separation prevents the ribbons from colliding when the signal level exceeds that required for 100-percent modulation.

Fig. 18-15F shows the method used to connect the light valve for recording. The 2-ohm potentiometer P1 is used to balance out the small differences in resistance between the noise reduction ribbons. This is accomplished by means of a photocell monitoring device in the recorder. The recorded sound track should appear as shown at (m) in Fig. 18-29.

Typical specifications for a three-ribbon valve are: impedance of signal ribbon (with shunting resistor), 0.3 ohm; impedance of noise-reduction ribbons, 0.6 ohm; power required for 100-percent modulation at 1000 Hz using the circuit of Fig. 18-15F, plus 12 dBm; noise-reduction closure current, 52 mA; ribbon spacing, 0.0010 inch.

The light valve is mounted in a modulator unit which houses the necessary optics that operate in conjunction with the light valve. The term modulator (Westrex) should not be confused with the term light modulator, as this term refers to a recording light valve, galvanometer, or any device used to modulate the exposure lamp beam of light. A modulator unit holds the light valve and is a part of the recording system.

Fig. 18-15G shows a three-ribbon light valve modified to produce two independent variable-density sound tracks for stereophonic recording. The center ribbon in this valve is grounded and is used only as a mask. The top and bottom ribbons carry both the audio signal and noise-reduction bias current. The two apertures are brought into line by the use of optical compensators mounted in the valve assembly, resulting in two sound tracks which are colinear.

For variable-area recording, a special light valve containing four ribbons and capable of recording two separate bilateral sound tracks which are also

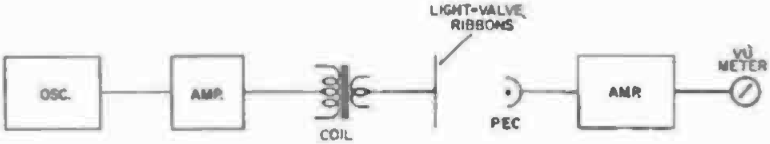


Fig. 18-16. The circuit used for tuning light valve ribbons.

colinear, is used. For variable-density recording, to prevent shifting of the stereophonic pattern because of unequal channel gain, it is important that the two sound tracks are of nearly the same value. Variations as small as 1 dB will sometimes cause an apparent shift in the sound and the action as shown on the screen. Therefore, density variations must be less than 0.05 between the two tracks.

For variable-density recording the two tracks appear together in the usual track position on the negative, and are each 48 mils wide and spaced 4 mils apart. When recorded as variable-area bilateral, each track is 36 mils wide, and spaced 4 mils apart. The trade name for this system is *PhotoStereo*. Since magnetic recording can offer so many other advantages, the practice of recording optical stereophonic sound tracks is not used to any great extent.

18.16 What is meant by tuning a light valve?—It is the process of resonating the light valve ribbons to a frequency above the normal recording frequency range. The tuning is done in a light-valve fixture consisting of an exciter lamp, optical system, photocell, and an amplifier with a VU meter at the output. (See Fig. 18-16.) In the early models of light valves, the peak at the resonant frequency varied from 12 to 20 dB above the reference frequency of 1000 Hz. In present day designs, this peak has been reduced to about 6 dB. In practice, an equalizer with an inverse frequency characteristic to the light valve is connected in the recording channel to smooth out the response in the region of the resonant frequency.

18.17 What is the clash point of a light valve?—It is a reference of 100-percent modulation of the valve, obtained by sending into the valve a signal that will cause the ribbons to just strike each other. This is the point of zero transmission of the valve. If the ribbons are permitted to clash during recording, strong harmonics will be generated, causing distortion.

18.18 Is the impedance of a light valve constant?—No. To prevent variation of the light-valve impedance from being reflected back to the driving amplifier, it is common practice to insert an attenuator of 3- to 6-dB loss between the light valve and its matching transformer and the output of the driving amplifier. At the lower frequencies, the valve impedance is practically a constant resistance, equal in value to the dc resistance of the light-valve ribbons. At the higher frequencies, the impedance of the ribbons rises. The pad isolates this impedance change from the amplifier, thus reducing the harmonic distortion.

18.19 How is the low impedance of a light valve matched to the output of the driving amplifier?—By an impedance-matching transformer. Generally, the input of the transformer is 600 ohms, with a secondary of 0.5 ohm to several ohms, depending on the design of the light valve. The impedance below 1000 Hz is approximately the dc resistance of the ribbons. An adjustable attenuator is provided for adjusting the 100-percent modulation point of the light valve for a given input signal. The driving circuit of a typical light valve is shown in Fig. 18-19.

18.20 What is a monoplanar light valve?—One in which the ribbons move in the same plane. It is also called a coplanar valve. This type of light valve may be used for the recording of variable-area sound tracks.

18.21 What is a biplanar light valve?—One in which the ribbons move in separate planes. Light valves which have more than two ribbons are of this type. Light valves of this type are generally used for the recording of variable-density sound tracks.

18.22 What is the velocity effect in a light modulator?—In variable-area sound recording, the exposure light is focused on a slit and the illumination

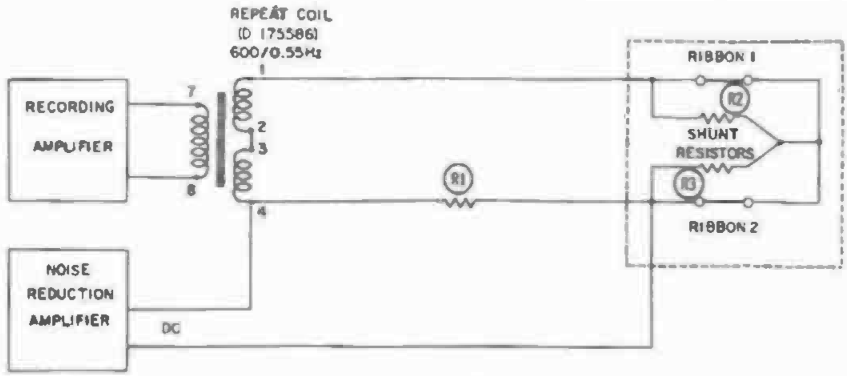


Fig. 18-19. The Simplex circuit used with a Westrex Corp. light valve for photographic recording.

at the slit is then focused onto the film. With the film moving past the recording light beam at a given rate of speed, a wave shape which corresponds to the electrical signal applied to the light modulator is traced onto the film. When the frequency of the signal is low, the waveform traced on the film will be long in length. If, however, the frequency is a very high one, the length of the waveform is very short. Under close examination, it will be found that the recording of the low frequency is true in its shape and that the density of the exposed area is relatively constant. This is not true in the case of the high-frequency recording. The shape of the exposed area does not correspond to the shape of the image that was traced on the film. The reason for this is the difference in velocity with which the light beam moves across the slit. The amount of exposure that is applied to the film is governed by several factors: (1) the time required by a particular point on the film to travel past the light beam; (2) the intensity of the exposure light; and (3) the time it takes the exposure light to move from its rest position on the slit to the limit of its excursion and to return to its rest position again.

With the rate of film travel and the intensity of the exposure light held constant, the factor that determines the amount of exposure on the film is the velocity at which the exposure light travels across the slit. Thus, it may be seen that the frequency of the recording signal affects the exposure of the film.

When a high frequency is recorded, the image on the film will have a greying out towards the peak of the waveform due to the diminishing exposure at that point. This is caused by the very short time period that the exposure light rests at the peak of its excursion.

The results of this phenomenon are two-fold: (1) As the frequency of the recorded signal is increased, the output level with respect to a reference frequency of the same amplitude will decrease. This is known as film loss. This loss is usually measured at 9000 Hz using a reference frequency of 1000 Hz. (2) The higher the frequency recorded, the greater will be the distortion of the waveform caused by the velocity effect. Film-loss measurements are discussed under cross-modulation measurements in Question 18.232.

"Ribbon velocity effect" is the term used to describe the distortion created by the light-valve ribbons in variable-density light valves, due to their direction of travel with respect to the film motion. If the velocity of the ribbons is comparable to the film velocity, a spurious variation in exposure results, producing distortion.

18.23 What are the essential differences of a variable-area light valve designed for negative recording and one for direct-positive recording?—In a light-valve setup for negative recording, the light from the exposure lamp passes through the light-valve ribbons and is then focused on the film. In a light-valve setup for direct-positive recording, the light is reflected from the ribbons and then focused on the film. Direct-positive valves are also capable of making negative recordings.

The ribbons in a direct-positive light valve are polished on one side to reflect the light. Negative-valve ribbons are not polished. Polishing the ribbons appears to have an annealing effect, resulting is a somewhat higher ribbon failure in direct-positive light valves.

18.24 What is the relationship in percent modulation of a light modulator to its deflection in decibels?—If a condition of 100-percent deflection (modulation) of the light modulator can be assumed, the percent modulation and modulation in decibels can be read directly from the scale in Fig. 18-24A. For deflections other than those shown on the scale, the table in Fig. 18-24B can be used. (See Fig. 25-133.)

Percent Deflection	*Decibels Down From 100-Percent Modulation*
80	2 dB
50	6 dB
25	12 dB
10	20 dB

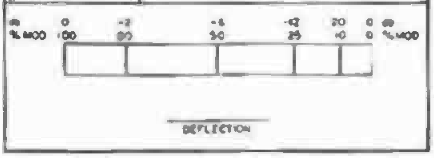


Fig. 18-24A. Percent light modulator deflection versus decibels.

dB	%	dB	%
0	100	26	5.012
1.0	89.13	27	4.467
1.5	84.14	28	3.981
2.0	79.43	29	3.548
2.5	74.99	30	3.162
3.0	70.80	31	2.818
3.5	66.83	32	2.512
4.0	63.10	33	2.239
4.5	59.57	34	1.995
5.0	56.23	35	1.778
5.5	53.09	36	1.585
6.0	50.12	37	1.413
6.5	47.32	38	1.259
7.0	44.67	39	1.122
7.5	42.17	40	1.000
8.0	39.81	41	.8913
8.5	37.58	42	.7943
9.0	35.48	43	.7080
9.5	33.50	44	.6310
10	31.62	45	.5623
11	28.18	46	.5012
12	25.12	47	.4467
13	22.39	48	.3981
14	19.95	49	.3548
15	17.78	50	.3162
16	15.85	51	.2818
17	14.12	52	.2512
18	12.59	53	.2239
19	11.22	54	.1995
20	10.00	55	.1778
21	8.913	56	.1585
22	7.943	57	.1413
23	7.080	58	.1259
24	6.310	59	.1122
25	5.623	60	.1000

Fig. 18-24B. Percent modulation versus decibels.

18.25 What is a penumbra intensity recorder?—A system of recording variable-density sound tracks by means of a galvanometer which was developed by RCA several years ago hut is not used commercially.

18.26 Describe the essential components of a photographic film recorder.—In many respects, the transport system and other mechanical details of a photographic (optical) recorder are quite similar to a magnetic film recorder. The essential differences are the optical recording system and its associated equipment. The two film recorders that will be discussed are typical of those used in the motion picture industry. The first to be discussed is the Westrex Model 1581A recorder. This recorder is available for both 16- and 35-mm film recording. The following description will apply to either type, except for the width of the transport components. The same light valves are used for both sizes of film.

Fig. 18-26A shows that the film path is symmetrical and permits the film to travel either forward or backward for negative or direct-positive recording. The film travels downward from magazine A through a light trap in the magazine, over idler roller B, engages the left side of film sprocket C, then goes under the filter roller D, over impedance drum E. The film now proceeds under a second impedance drum EE under filter roller DD, engages the right side of film sprocket C, then goes over idler roller DD, and back into the right-side of the magazine A.

Before making the second engagement with film sprocket C, the film loop length is adjusted until the flanges of the filter rollers D and DD bisect or just cover two black reference dots on the panel behind the rollers. As an aid in threading, a loop lamp H on the control panel will go out when the correct length of film is threaded. Changing the film loop plus or minus one sprocket hole will cause the lamp to light. At F and FF are two film strippers, one of which serves to operate a film buckle switch. Knob G is used when necessary to position the two filter-arm flanges D and DD, with respect to the two dots when the machine is in operation. The mechanical filter system is the Davis tight-loop system, discussed in Question 18.28.

The light valve I is mounted in a modulator unit J with the objective lens in the film compartment (not shown). At K is a focusing and adjustment control for the objective lens. In the housing is also a rotating mirror, a part of the light-valve monitoring system, with the monitoring screen at M. A footage counter appears at N.

This machine can be used for recording negative or direct-positive sound track, using either "A" or "B" wound film. With the installation of a suitable light valve in the modulation unit, the following can be recorded: double-width push-pull negative; standard-width bilateral negative; standard-width double bilateral negative; standard bilateral direct-positive; and double-width push-pull direct-positive sound track. Any one of these light valves can be installed without requiring any adjustment of the optics except for the introduction of the appropriate mask width for standard or double-width sound track. It is recommended that the variable-area double bilateral sound track be used in preference to the bilateral sound track. Such sound tracks are illustrated in Figs. 18-29B and 18-299A.

Direct monitoring from the light valve image is possible with the use of a system of mirrors which reflects a part of the modulated light beam to an eight-sided rotating prism. The prism reflects an image of the light valve ribbons on the monitor screen M, which results in a variable area image similar to that being recorded on the film. A diagram of the recording and monitoring optical system is given in Fig. 18-26B.

An exposure meter is used to monitor the exposure lamp (exciter lamp). The exposure meter optical path provides a means for comparing the light from the exposure lamp that passes through the light valve with light from a reference lamp. By rotating a mirror and aperture assembly to either of two positions, the light from either source is directed to a photoconductive cell which is connected to an external exposure meter. The mirror aperture assembly in Fig. 18-26B is shown in the position that permits the light from the reference ex-

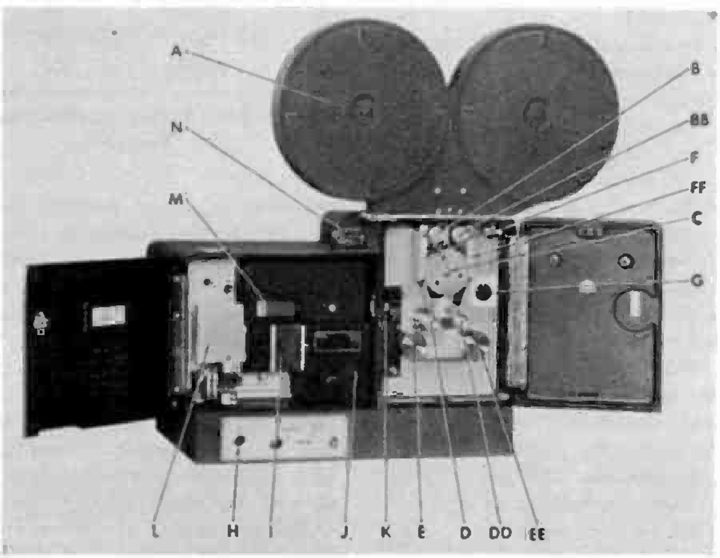


Fig. 18-26A. Front view of Westrex Corp. Model 1581A photographic film recorder.

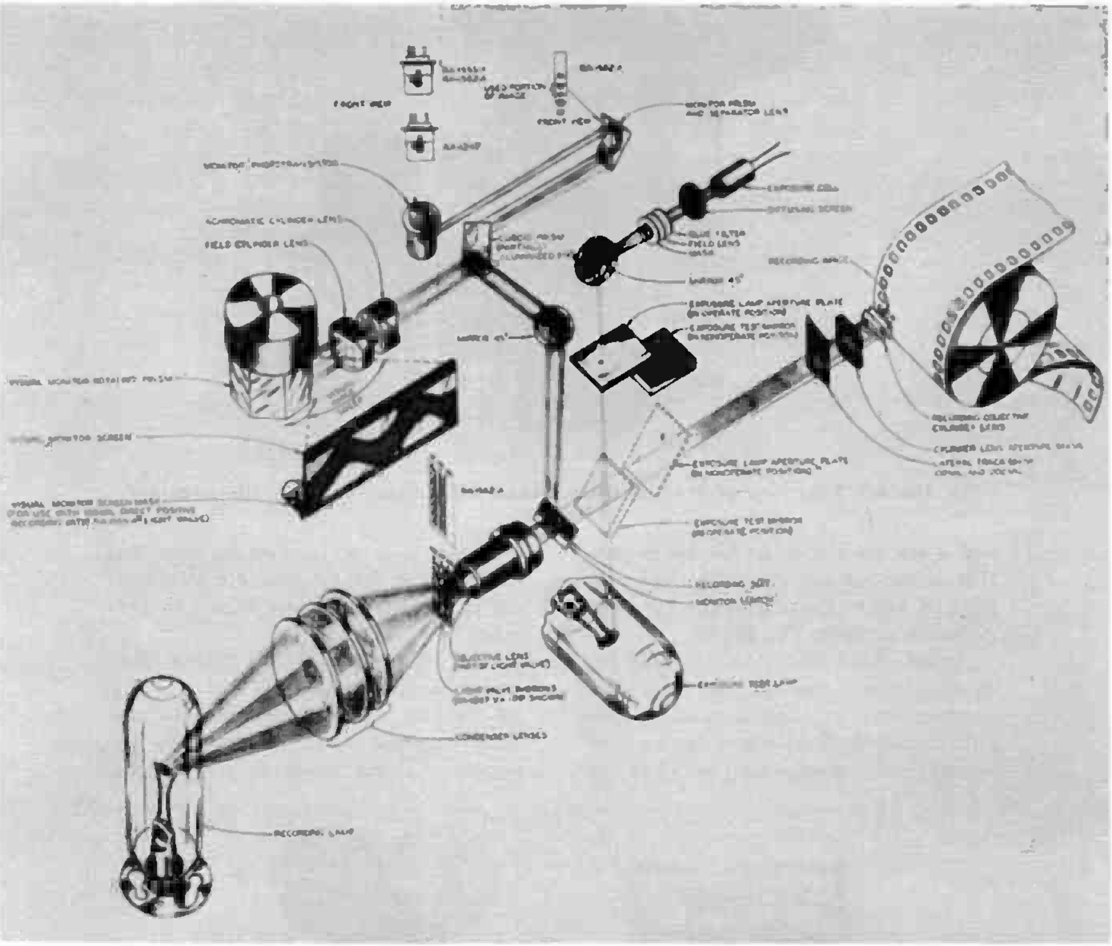


Fig. 18-26B. Diagram for the optical system used in the Westrex Corp. Models 1581A and C photographic film recorders.

posure test lamp to pass through the aperture to be reflected by the 45-degree mirror, pass through the mask, field lens, blue-filter assembly, diffusing screen, and to the photoconductive cell. With the mirror rotated to the alternate position, the light passing through the light valve is reflected upward into the 45-degree angle mirror through a field-mask, blue-filter, diffusing screen, to the exposure cell. (Blue filter refers to an ultraviolet filter discussed in Questions 18.7 and 18.8.)

An interior rear view of the recorder, showing the film take-up and motor-drive mechanism, is given in Fig. 18-26C. The plugs at the bottom are for external noise reduction amplifier (NRA), audio, and power connections.

An overall view of a Series 900 recording system with its control equipment is shown in Fig. 18-26D. Contained within the cabinet, starting at the upper left are noise reduction amplifier and controls, exposure lamp control, VU meter and attenuator, and the compressor amplifier. In the second row is a photometer for measuring the exposure lamp intensity for both variable-area and variable-density recordings,

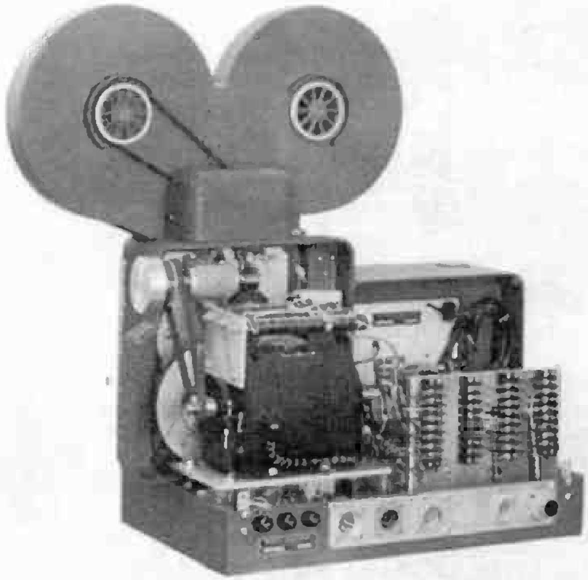


Fig. 18-26C. Rear view of Westrex Corp. Model 1581A photographic film recorder.

and a test oscillator. At the lower portion of the cabinet are the power supplies. A block diagram of the complete system is given in Fig. 18-26E.

As an added feature, a magnetic recording head can be installed in the film compartment in such a manner that either method of recording can be used without any mechanical or electrical changes. It is of interest to note that recorders of this type have a total flutter of considerably less than 0.10 percent.

The RCA Model PR-31 35-mm photographic film recorder is shown in Fig. 18-26F. Its essential components are: film magazine A, for holding the film supply; locking screw B, which holds

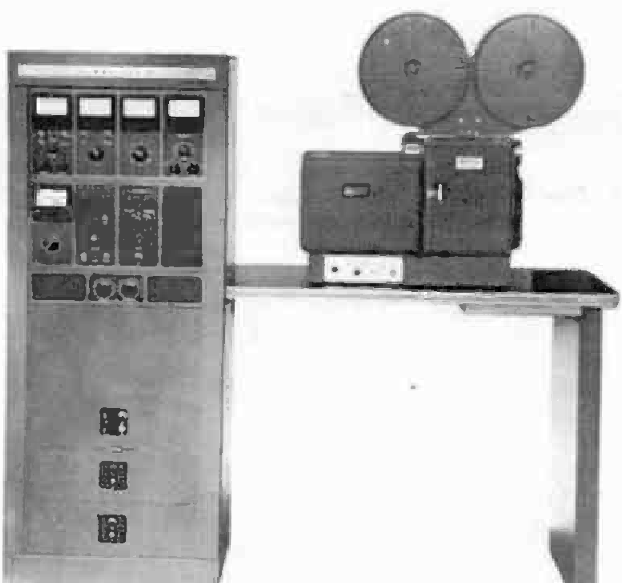


Fig. 18-26D. Westrex Corp. Series 900 photographic film recorder with its compressor, noise-reduction amplifiers, and associated equipment.

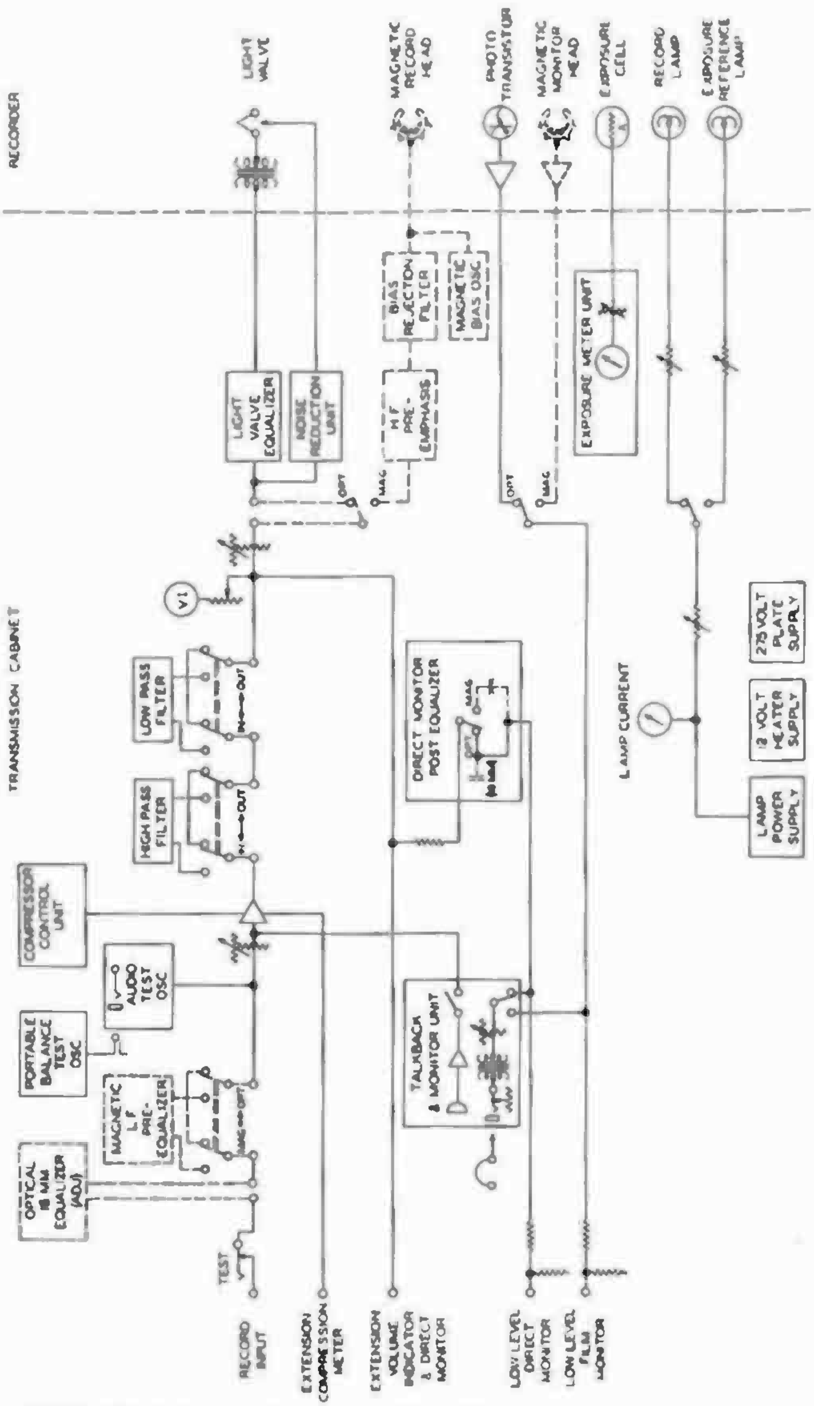


Fig. 18-26E. Block diagram for Westrex Series 900 photographic film recorder and control cabinet.

the magazine to the recorder case; supply magazine spindle and plastic film core C; light trap D, through which the film passes to the recorder transport system; optical system E for slating mechanism; digits and letters EE for photographic slating mechanism; bearing heater switch F; galvanometer monitoring card G; exciter-lamp housing H; galvanometer tilting screw I;

galvanometer II; recording optical system J; photometer lens K; optical-system housing with objective lens L; objective-lens focusing ring LL; photometer meter M; motor-system starting switches N and NN; film-marker button O; galvanometer switch P; magnetic-drive system rheostat Q; field-current meter QQ; exciter-lamp hold switch R; master dc switch RR; exciter-lamp current S; exciter-lamp rheostat SS; film punch T; pull-down pull-up sprocket U; idlers V and VV, for holding film against sprocket; tight-loop idlers W and WW; recording drum X; film Y; take-up magazine AA; and take-up magazine spindle and core CC.

The film drive consists of a gear-driven 32-tooth sprocket with double-pad rollers, magnetically driven sound drum (impedance drum), and two undamped sprung rollers. The sprung rollers are provided with position stops, which are held open by spring action. The design is such that the same length of film is used each time the recorder is threaded. Threading is done with one spring roller in its normal position as held by its spring, and the other spring roller is held against a stop opposing the spring action. This provides a tight-loop path. Damping is provided by a magnetic drive to the impedance drum. The current through the field coil increases the damping, but makes little difference in the performance of the recorder. The impedance drum employs a bronze-sleeve bearing with a heater and thermostat set for a 75-degree operation. This eliminates the necessity of the bearing warm-up time. The current through the magnetic-drive field-coil is controlled by rheostat (Q), which may be adjusted while the machine is in motion. This transport system and magnetic drive is used in all RCA 5-mm photographic film recorders. The total flutter is on the order of 0.05 percent. It should be mentioned that optical film recorders use sprockets, which are designed to operate with either negative or positive film-base perforations. These

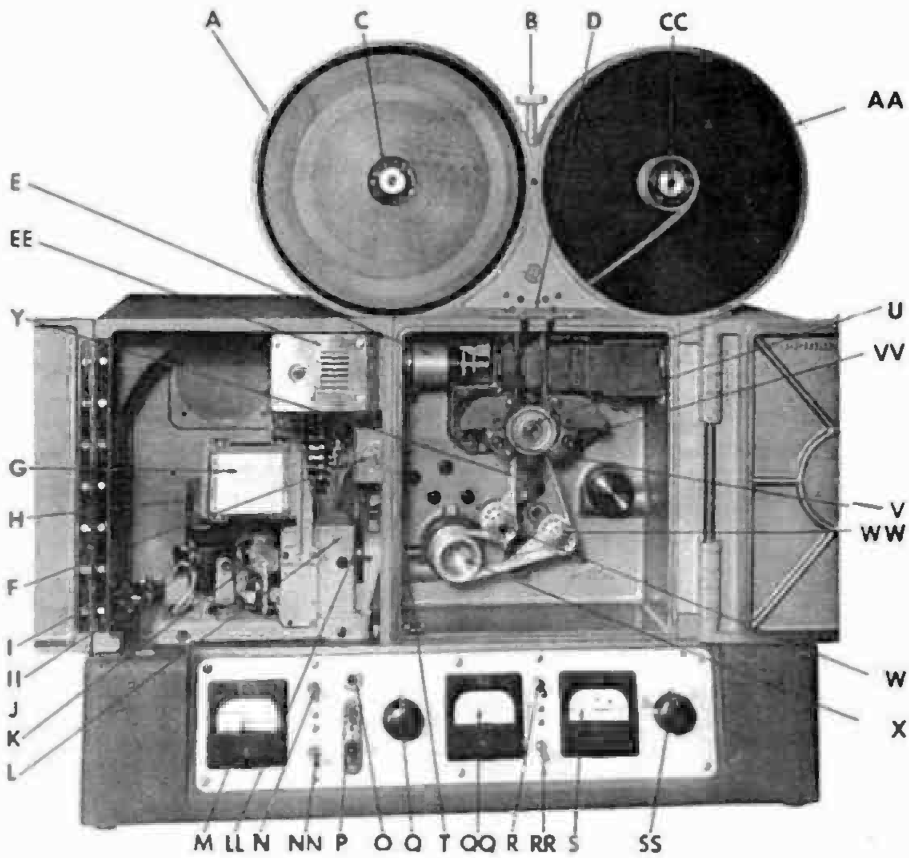


Fig. 18-26F. The RCA MI-10745, PR-31, 35-mm photographic film recorder.

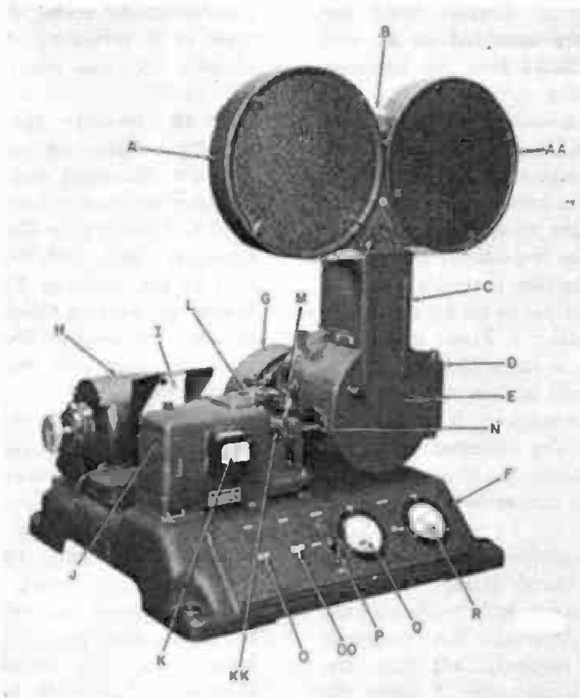


Fig. 18-26G. The RCA PR-23 photographic film recorder manufactured in 1936.

sprockets are also used for magnetic recorders.

Fig. 18-26G shows an RCA PR-23 photographic film recorder, manufactured about 1937. It is interesting because of its film transport system. This machine did not use a tight-loop; instead, it used a free-loop system. It attained its steadiness of drive by the use of a magnetic-drive system coupled to the impedance drum, similar to that used in the PR-31 recorder. The long neck below the film magazine held a film punch and numbering mechanism.

The principal components of the PR-23 recorder are: film magazine A, locking screw for film magazine B, magazine supporting neck C (although not shown, a slating and film-punching device may be installed in this portion of the recorder), housing door D, compartment containing the film transport mechanism E, recorder base F, magnetic-drive G, motor H, galvanometer monitor screen I, galvanometer and optical-system housing J, photometer meter K, photometer operating lever KK, exciter-lamp monitoring mirror L (image of filament is thrown on the galvanometer monitor card for constant observation), optical-system focusing ring M, objective lens N, motor-system starting switches O and OO, galvanometer key switch P, field-current ammeter Q, and exciter-lamp current ammeter R.

A diagram of the PR-23 transport system is given in Fig. 18-26H. The film A leaves a supply magazine above and is pulled downward by a 32-tooth sprocket B. Two small rollers C and CC

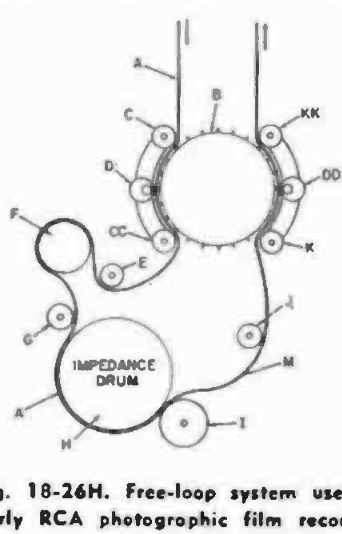


Fig. 18-26H. Free-loop system used in early RCA photographic film recorders (1933).

hold the film in contact with the sprocket and are mounted on an arm which may be lifted from the sprocket by the action of a cam D.

Leaving the sprocket, the film passes under roller E making a small loop between the bottom of roller CC and roller E. The film now passes over roller F making a tight wrap, held by roller G, then over the impedance or recording drum H. The film is tightly clamped to the surface of the drum by a rubber-surfaced pad roller I. From this point, the film forms a large loop M, passes over roller J, and is then again held in contact with the sprocket B by the rollers K and KK, also mounted on an eccentric, DD. Finally, the film is taken up by the take-up magazine at the upper right.

When the machine is first threaded, the roller J is used to form the loop M. If the loop is correctly made and the field current through the magnetic drive system properly adjusted, the film will not touch roller J when the machine is running. This system is called a free-loop drive. Tight-loop drives are shown in Figs. 18-26A and B.

18.27 What is an impedance or sound drum? What is its purpose?—A metal drum used in the film transport system over which the film passes. Its purpose is to smooth out irregularities in the motion of the film caused by the action of the sprockets, thus reducing flutter.

The drum is free wheeling and revolves by the friction of the film on its surface pulling it around. On the opposite end of the drum shaft is mounted a heavy flywheel. When the drum attains its normal speed, the flywheel opposes any change in speed transmitted to the film on its surface. In a magnetic recorder, the drum is constructed from a nonmagnetic metal. A cross-sectional view of a recording drum used in a photographic film recorder is shown in Fig. 18-27.

18.28 Describe the basic principle of a Davis tight-loop transport system.—The film transport and filter system to be described was a development in 1945 of C. C. Davis of the Electrical Research Products Corp. (ERPI), a former subsidy of the Western Electric Co., now known as Westrex Corp. Such transport systems are used in Westrex photographic and magnetic recorders and reproducers.

Fig. 18-28A shows that the film path and damping section consists of an upper and lower sprocket, an impedance drum with a heavy flywheel on the opposite end, two pivoted arms with idler rollers which form a tight loop around the impedance drum, a spring connected between the two arms, and an oil-filled dash-pot connected to the lower arm. The double-tension loop supplies a relatively large amount of film tension, which is quite suitable for flywheel starting and driving, and at the same time supplies the correct amount of loop compliance. The loops are not tensioned independently, but by the differential action through the spring between the two arms. The reactance opposing the flywheel motion is that resulting indirectly from changes in the working angles through which the spring tension is applied to the loops.

A second method of using the tight-loop system for magnetic recorders is shown in Fig. 18-26B. Here the film passes over a 16-tooth sprocket, an idler roller mounted on a pivoted arm, over a second idler roller to an impedance or sound drum, in which is located a magnetic recording head. The film now

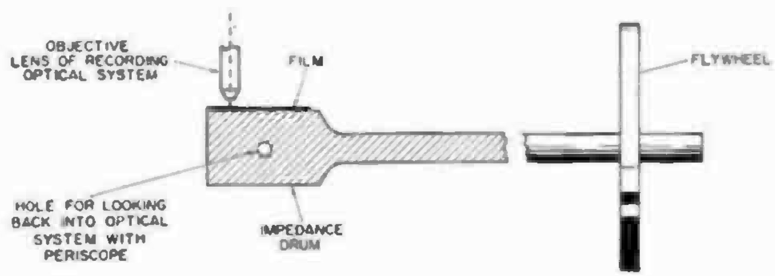


Fig. 18-27. Impedance or recording drum, flywheel, and objective lens of a photographic film recorder.

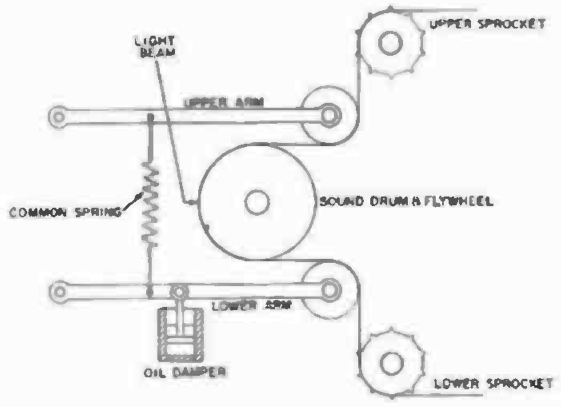


Fig. 18-28A. Schematic diagram for Davis tight-loop film transport system used in photographic and magnetic film recorders and reproducers.

passes over a second idler to a second pivoted arm and idler roller, then to a second 16-tooth sprocket. The filter elements consist of the two arms with a spring between, and an oil-filled dashpot connected to the left hand arm. The filter action is the same as for the system in Fig. 18-28A. A magnetic monitor head may be used to replace the lower roller on the right hand side.

Fig. 18-28C shows a third method of applying a tight-loop system to a photographic film recorder, similar to that shown in Fig. 18-28A. Uniform film motion at the translation point, with relative freedom from flutter, is achieved by the introduction of mechanical filtering in the film path between the two

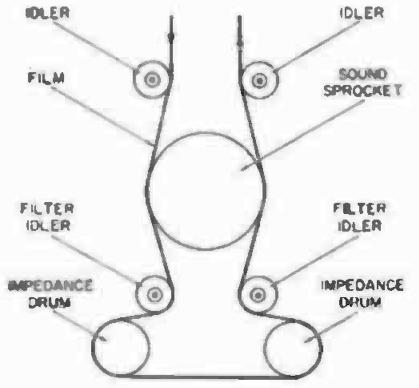


Fig. 18-28C. Third version of Davis tight-loop transport system, used in photographic film recorders.

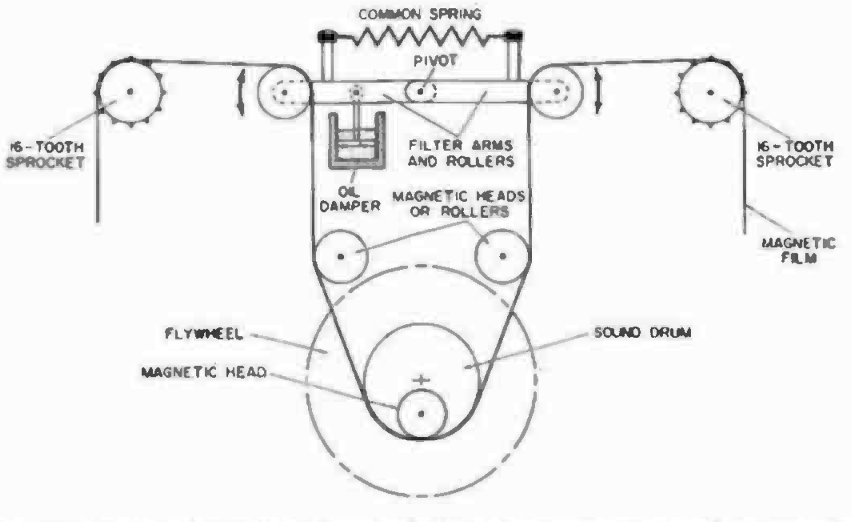


Fig. 18-28B. Second version of Davis tight-loop transport system, used in magnetic film recorder.

engagements of the film with the film sprocket. The filter action consists primarily of the inertia of the impedance drum assemblies and the damped compliance of the filter-arm assembly. At high rates of flutter, the compliance of the film becomes an effective element in the filter mechanism. The filter-arm assembly provides two rollers, mounted on pivoted arms and arranged with one roller in each of the film paths between an impedance drum and the film sprocket. Two coiled springs are employed in the filter. One spring is con-

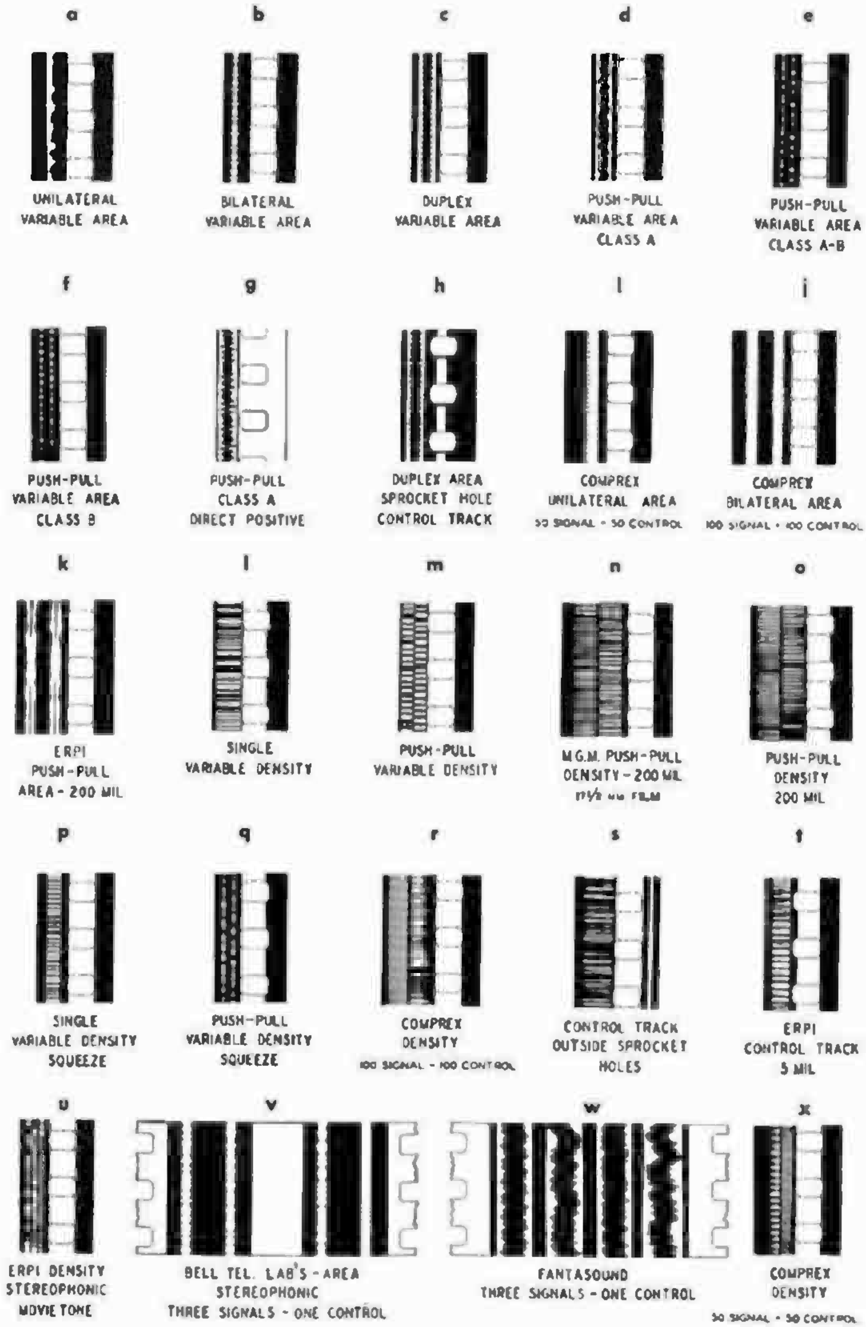


Fig. 18-29. Photographic sound track images used both past and present. Several of the above images are now obsolete, but are included for reference purposes.

nected between the filter arms to provide the compliance, as well as the film tension in the filtered film loop. The other spring is adjustable and provides symmetrical positioning of the arms. Damping is applied to the right-hand filter roller (viewed from the front) in the form of a fluid dash-pot filled with a silicone fluid of 200-centistokes viscosity. The film tension is adjusted by control G (Fig. 18-26A).

18.29 Describe the various types of sound track both past and present used for photographic film recording.—During the development of photographic sound recording, many different types of sound-track images were created for both variable-density and variable-area recording. Among the variable-density recordings were the standard, push-pull, squeeze, and comprex. For variable-area recordings, there were unilateral, bilateral, and several variations. For both types of recordings, a number of push-pull configurations operating class-A, -B and -AB were developed. Stereophonic sound tracks were also recorded, using up to four tracks. Twenty-four different sound tracks are shown in Fig. 18-29. For present-day recording, both the standard variable-density and variable-area bilateral sound tracks are used, with the variable-area type prevailing.

18.30 Describe a unilateral variable-area sound track.—It is a sound track consisting of a single group of modulations along one edge of the sound track area (Fig. 18-29). This was the design of the original sound track used with the first RCA photographic film recorder, Model R-3, in 1926.

18.31 What is a bilateral variable-area sound track?—A sound track with symmetrical modulations along both edges of the sound track area, as shown at (b) in Fig. 18-29.

18.32 What is a class-B, push-pull, variable-area sound track? — A sound track containing two groups of modulations as shown at (f) in Fig. 18-29, and in Fig. 18-298. Each chain of modulations is 180 degrees out of phase with the other. The modulations of each track are connected by a thin line called a bias line.

18.33 What is the appearance of a variable-density sound track?—It consists of a group of thin lines running at 90 degrees to the motion of the film as shown at (l) in Fig. 18-29. Their degree of exposure varies with the percentage of modulation.

18.34 What is a push-pull variable-density sound track? — Two variable-density sound tracks recorded alongside each other as shown at (m) in Fig. 18-29. The modulation of the two sound tracks are 180 degrees out of phase with respect to each other.

18.35 What is a dulateral sound track?—Two variable-area, unilateral sound tracks recorded in phase alongside each other, as shown in Fig. 18-35.

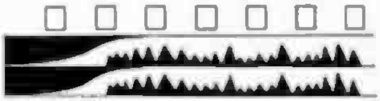


Fig. 18-35. The dulateral variable-area sound track. The two groups of modulations are in phase.

18.36 What is a squeeze track?—A single variable-density sound track in which the width is reduced during periods of low or no modulation to reduce the background noise of the film. Reducing the exposed area 50 percent reduces the background noise of the film 6 dB or 50 percent. A typical squeeze track is shown at (p) and (q) in Fig. 18-29.

18.37 What is the purpose of recording multiple variable-area sound tracks? — In multiple variable-area sound-track recording, six or more identical bilateral sound tracks are recorded side by side (Fig. 18-37). Such sound tracks are generally used with 16-mm photographic sound recording. The advantages claimed for this method of recording are that because the tracks are quite narrow, distortion caused by azimuth deviation and uneven slit illu-

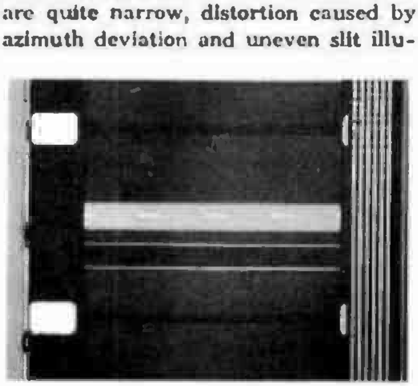


Fig. 18-37. The Maurer 16-mm multiple variable-area photographic sound track.

mination is considerably reduced, and the signal-to-noise ratio is increased. This type sound track is used by Maurer, and the German Tobias-Klangfilm Co. A method of recording 13 sound tracks, developed by Siemens-Halske, also of Germany, has been described in the literature. The image of this sound track should not be confused with a stereophonic or push-pull sound track, as all six of the tracks carry the same information.

18.38 What is a control track?—A sound track consisting of single or multiple frequencies for the purpose of controlling amplifiers or loudspeakers in a multiple-channel reproducing system. The control frequencies are passed through bandpass filters to a control rectifier-amplifier and changed to direct current. This dc is used to control the opening and closing of the amplifiers driving the loudspeakers. Two such sound tracks are illustrated at (v) and (w) in Fig. 18-29.

18.39 What is minus dialogue track?—A prerecorded music and sound effects track for dubbing foreign language motion pictures. The dialogue is carried on a separate sound track. Thus, any language may be used without having to rerecord the music and sound effects a second time.

This system was used principally with photographic sound recording; however, since the advent of magnetic recording, it is common practice to record the dialogue, music, and sound effects on separate sound tracks using a three-track recorder.

18.40 If raw stock film is stored in a high temperature what effect does it have on the emulsion?—It causes the emulsion to develop a fog which is independent of development.

18.41 What is a direct-positive recording? — A sound track which has been so exposed that it requires processing only once. The image is positive and may be used as any normal positive print. Direct-positive recording may be accomplished by changing certain components in the recorder optical system to produce positive images rather than negative. In the RCA variable-area system of recording, this is done by shifting the image from the optical mask to expose the film in such a manner that it processes as a positive; that is, the image is reversed. Where the film would normally be exposed and develop black, it is not exposed; therefore, it develops white or positive. In the Westrex Corp. recorders, direct-positive recording is accomplished by changing the light valve to a direct-positive light valve. Direct-positive recording is used only with variable-area recording systems. (See Questions 18.23, 18.26, and 18.304.)

18.42 What are the advantages of direct-positive recording? — Actually, there is no advantage soundwise. Direct-positive recording is used only when time is a factor and only one print is required. Considerable time is saved, as the track is recorded directly on the print and requires processing only once. It should be understood that the sound of a direct-positive recording may not be equal to a negative and print, unless a cross-modulation compensator is used. Such sound tracks are developed to a gamma of approximately 2.5 to 3.0, with the density ranging from 1.5 to 1.9. Direct-positive sound tracks are played back in the normal manner as for any positive sound track. The proper exposure is determined by a series of cross-modulation tests, as discussed in Questions 18.232 and 18.246, and a sibilant test, as described in Question 18.283. Direct-positive recording is used only with variable-area sound track.

18.43 What is the azimuth setting of a film recorder?—An adjustment of the optical system which provides a means of setting the optical slit at exactly right angles to the direction of travel of the film. (See Question 18.47.)

18.44 What is the effect of incorrect azimuth in a variable-density film recorder?—Loss of high frequency response and level.

18.45 What is the frequency used for adjusting the azimuth of a 35-mm film recorder?—9000 Hz.

18.46 What is the frequency used for adjusting the azimuth of 16-mm film recorders?—In the earlier model 16-mm film recorders, 4000 Hz was used to adjust the azimuth because the speed of 16-mm film (36 feet per minute) is 40 percent slower than 35-mm film running at 90 feet per minute. When 4000 Hz is recorded at 36 feet per minute, it corresponds to a frequency of 10,000 Hz at a speed of 90 feet per minute ($4000 \times 2.5 = 10{,}000$ Hz). Because of the better optics and resolution of present-day re-

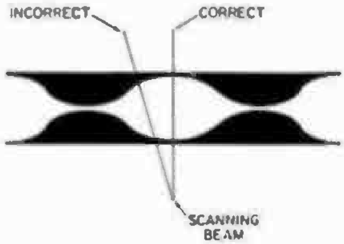


Fig. 18-47. The effect of incorrect azimuth adjustment on a variable-area sound track, when being reproduced.

cording emulsion, 16-mm recorders now use 6000 to 7000 Hz for azimuth adjustment.

18.47 What are the tolerances for 16-mm and 35-mm azimuth adjustment? —Typical specifications for the slit rotation are: The error must not exceed plus or minus 0.0001 inch for both 16- and 35-mm systems. The effect of incorrect azimuth adjustment in a recorder is the two sides of the sound track are not symmetrical. The effect of incorrect azimuth of a variable-area sound track when reproduced is shown in Fig. 18-47. (See Question 18.53.)

18.48 What is the effect of incorrect azimuth in a variable-area film recorder? — The distortion is increased with a loss of high frequencies. When reproduced, the two sides of the modulation are not scanned in the same plane as shown in Fig. 18-47.

18.49 At what speed is 35-mm film recorded and projected?—90 feet per minute, 18 inches per second, or 24 frames per second.

18.50 Describe the differences between single and super 8-mm prints.— For single 8-mm prints, an 8- or 16-mm negative is exposed, processed, and a direct print is made from the 8-mm negative. Or an optical reduction is made from a 16-mm negative, then printed in the usual manner. If the picture is to carry a sound track, the print material can be pre- or post-striped magnetically. The sound is recorded at a speed of 24 fps, using conventional techniques within the limits of USASI (ASA) Standard given in Question 17.170.

If the sound is recorded at a speed of 24 fps, the product is termed Super 8-mm film. For silent films, the frame rate is 18 fps. (See Question 19.26.)

For Super 8-mm film, a 16-mm negative is used and may be processed by the laboratory in several different ways. The use of 16-mm original negative increases the picture resolution, resulting in a superior quality picture.

To describe only one method of laboratory practice: Four exposures using an internegative are made on 35-mm print stock that has been prestriped and perforated for 8 mm. The print is exposed, and the sound is recorded and monitored while running at a printer speed of 200 fpm.

After processing, the 35-mm print containing four 8-mm exposures is slit. This results in four 8-mm prints, each containing a single row of perforations and a sound track. Such a method is one way of reducing processing time and cost, and is quite important where several hundred prints of a given subject are to be printed and sound recorded. A proposed SMPTE Standard USASI (ASA) Ph22-164 is given in Fig. 18-50.

18.51 What is split film?—35-mm film which has been split in half and is called 17.5-mm film. Split film has only one row of sprocket holes and runs at a speed of 90 feet per minute in optical film recorders. Magnetic film recorders also use split film (17.5 mm) and run at a linear speed of 45 or 90 feet per minute. The use of split optical film is now obsolete.

18.52 What is a focus test?—A test of the recording optical system in a film recorder to determine the sharpness of the image on the film.

18.53 How is a focus test made?— If the optical system objective lens is not provided with a calibrated focus ring, a paper strip is prepared, as shown in Fig. 18-53, and wrapped around the end of the objective lens. The scanning beam is focused on the film by means of a magnifying glass. When it appears to be in focus, the paper is fastened to the revolving lens barrel with the zero calibration opposite a reference mark.

Start the test by rotating the lens barrel to number ten (10) negative of the calibration mark. Make a recording using either 9000 or 7000 Hz (see Question 18.45 to 18.48) for each calibration mark on each side of the zero calibration. The locking screws in the lens barrel must be tightened for each focus recording to assure a positive set. Process the film and observe the image under a microscope. The sharpest image

Proposed USA Standard Dimensions for

Magnetic Sound Record on Super 8 Motion-Picture Film Perforated 1R-1667

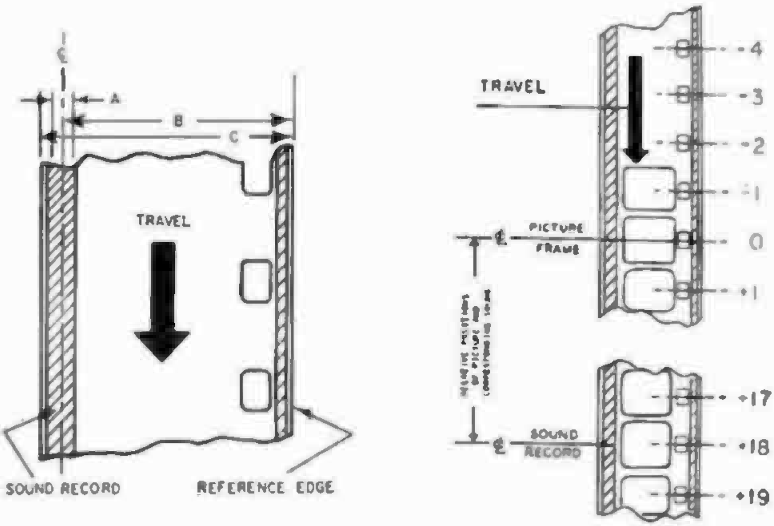
PH22.164

1. Scope

1.1 This standard specifies the lateral location and dimensions of the magnetic sound record on super 8 motion-picture film.

Dimensions	Inches		Millimeters	
A*	0.019	min	0.48	min
B	0.298 ± 0.001		7.58 ± 0.03	
C	0.314	nom	7.98	nom

* See Note.



Film As Seen Looking Toward the Lens

1.2 This standard also specifies the picture-sound separation of super 8 motion-picture film with a magnetic sound record.

2. Dimensions

The dimensions shall be as given in the figure and table.

3. Picture-Sound Separation

The magnetic sound record on the film shall precede the center of the corresponding picture by a distance of 18 frames ± 1 frame.

4. Magnetic Striping

The magnetic striping shall be as specified in Proposed USA Standard Dimensions of Magnetic Striping of Super 8 Motion-Picture Film Perforated 1R-1667, PH22.161.

NOTE: Dimension A applies to records produced in equipment using the same head for recording and reproducing. In commercially-produced prints intended for use on a variety of reproducers, it is recommended that a recording head be used capable of producing a 0.025-in. minimum width record having the same centerline.

Fig. 18-50. Proposed SMPTE Standard for magnetic stripe sound track placement, on Super 8-mm film.

represents the correct setting. (See Fig. 18-301.)

If a microscope is not available, reproduce the negative and measure the signal amplitude. The one with the highest output is the one with the sharpest focus. If a negative is reproduced with the sound track and emulsion in position, the test will reproduce in reverse to the way in which it was recorded. This is especially true with 35-mm film. When the test position for

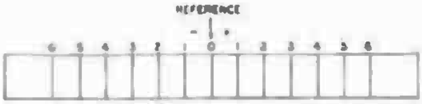


Fig. 18-53. Paper scale for adjusting the focus of an optical film recorder.

the sharpest focus is found, the lens barrel is returned to the corresponding calibration in the direction the test was made. This removes any play in the threads. The locking screws are then set.

18.54 What solution is used for cleaning recorder optical systems?—The coated surfaces of modern lenses, although hard, are microscopically thin and require but little abrasive action to cut through their coating and affect their antireflective qualities. A piece of pure soap the size of a pea, dissolved in a pint of distilled water, will work very well for removing scum and other dirt. Apply with a camel hair brush. Wipe clean with a soft, lint-free cloth. Do not use laundry soap powders. Solvents such as carbon tetrachloride and similar liquids should never be used to clean lenses. Isopropyl alcohol may be used if extreme care is taken to prevent it from entering the lens system, as it may dissolve the cement used to hold the optical elements together. The recommended cleaner is petroleum ether applied with a cotton swab.

18.55 What is noise reduction in a film recording system? — An electronic system of reducing the exposure to the film during periods of low or no modulation, thus reducing the inherent noise of the processed film when reproduced.

18.56 Where is the noise-reduction bias current applied to a light modulator?—In the original RCA recorders, the noise-reduction current was applied to a coil in the recording galvanometer. Later, the method of noise reduction was changed to use noise-reduction shutters placed in the optical train. After certain changes in the galvanometer design, the noise reduction was again applied to the galvanometer, and this is the system presently used. In the Westrex recorders, noise-reduction current is applied directly to the light-valve ribbon.

18.57 What is ground noise?—It is the residual noise in a recording system as the result of the nonhomogeneity of the recording media. This applies equally

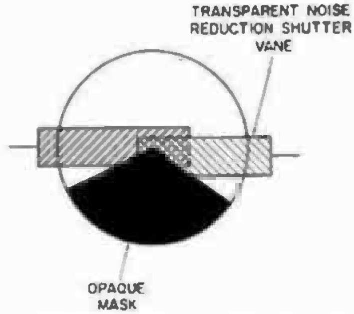


Fig. 18-58A. Shutter system for double noise reduction variable-area direct-positive recording.

well to both optical and magnetic recording systems.

Ground noise is proportional to the exposed area of the positive sound track. The noise-reduction system used when the negative sound track is exposed reduces the exposed area to an amount just large enough for the sound track, without clipping the peaks of the modulations. (See Question 18.58.)

18.58 Describe a double-noise reduction system.—It is a system proposed by J. G. Streiffert of Eastman Kodak Co., for variable-area direct-positive recording. The system involves the use of noise-reduction shutters, with polarizing devices in the optical train, an opaque mask, and transparent noise reduction shutters, to produce an elliptically polarized light beam. The basic plan of the double noise-reduction system appears in Fig. 18-58A, with a typical sound track shown in Fig. 15-58B. Although considerable experimental work was done on this system, it has not been used commercially.

18.59 What are noise-reduction shutters?—The first RCA optical film recording systems used no noise reduction. Later, to reduce background noise, a single noise-reduction shutter was added in the optical train to reduce the exposure to the film during periods of low or no modulation. The resulting sound track had the appearance of (a) in Fig. 18-29. Later, the shutter was removed and the noise-reduction current applied to a coil in the galvanometer as-

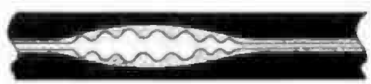


Fig. 18-58B. Appearance of direct-positive sound track.

sembly, which biased the galvanometer to one side during periods of low modulation, thus reducing the exposure at the film as the modulation varied. In later recorders, a return was made to a method employing the use of two shutters in the optical train (Fig. 18-78). The sound track had the appearance of Fig. 18-29 (b) and (c). Certain changes were then made to the galvanometer, permitting noise-reduction current to be again applied to a coil in the galvanometer. (This method is used in all present-day RCA film recorders.)

When the noise-reduction current is applied to shutters at high levels of modulation, they are fully opened to prevent clipping of the galvanometer light beam. At intermediate levels, the shutters follow the modulations of the galvanometer with sufficient margin to prevent clipping. At periods of low no modulation, the shutters close down to a very small opening, leaving a thin line of exposure to the film, called bias lines. In this manner, the background noise of the film caused by its granular nature is reduced as it is unexposed.

The action of the shutters on a biased galvanometer is operated from a noise reduction amplifier (NRA). The NRA is actuated by the same audio signal that operates the recording amplifier driving the galvanometer. Noise-reduction shutters are used only with the RCA variable-area recording system. A block diagram showing how the NRA is driven from a bridging bus and applies the bias current to a pair of shutters is given in Fig. 18-59.

With no signal at the bridging bus, the noise-reduction amplifier applies a fixed dc bias to the actuating coils of the shutters, closing them down and leaving only a small area of exposure for the bias lines (Fig. 18-290).

When a signal appears at the bridging bus, it actuates both the recording and noise-reduction amplifiers, simultaneously. The signal passing through the noise-reduction amplifier is rectified and changed to direct current and used to cancel the fixed dc bias flowing through the noise reduction shutter coils. The amount of cancellation depends on the percentage of modulation or amplitude of the signal at the bridging bus. The signal is passed through the recording amplifier without change and deflects the galvanometer mirror in proportion to the amplitude of the signal. The resultant sound track appears as shown in Fig. 18-61.

The first few modulations are clipped by the noise-reduction shutter because of the delay in its operating time. The amount of modulation clipped depends on the attack time. Typical operating times for noise-reduction shutters are shown in Fig. 18-61. A typical optical system layout and the position of the shutters behind the first condenser lens is shown in Fig. 18-338A.

In the variable-area biased galvanometer system, shutters are not used. The noise-reduction bias current is applied to a coil in the recording galvanometer which pulls the galvanometer mirror to one side during periods of low or no modulation, leaving only a small

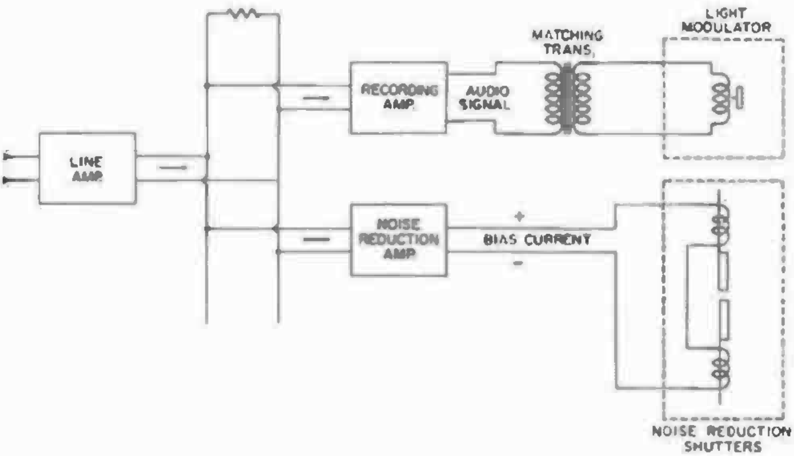
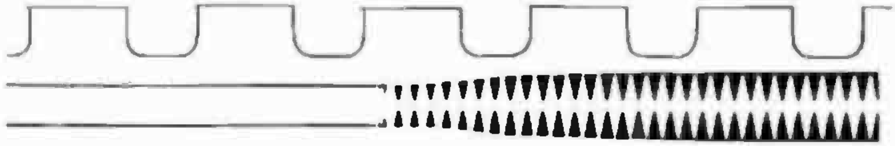
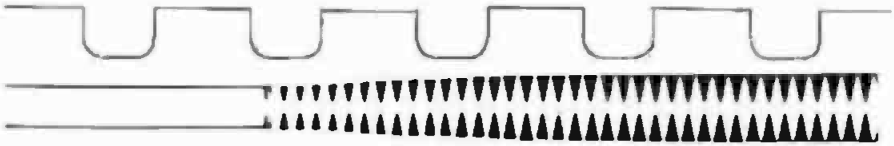


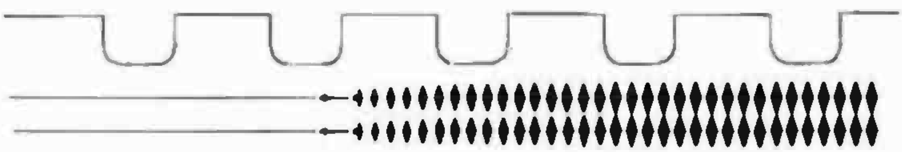
Fig. 18-59. A typical noise-reduction system for variable-area recording using noise-reduction shutters (now obsolete.)



(a) 35-mm duplex track, opening time 15 milliseconds.



(b) 35-mm duplex track, opening time 30 milliseconds.



(c) 35-mm double bilateral track, opening time 28 milliseconds.

Fig. 18-61. 1000-Hz noise-reduction shutter opening tests.

area of exposure for the bias lines. The noise-reduction action is similar to that described for the shutter type. (See Questions 18.12, 18.78 and 18.79.)

18.60 What is a bias line?—A narrow line of exposure between the modulations of the sound track as shown in Fig. 18-290. In all systems, except the class-B push-pull, the bias line is obtained by letting a small amount of light fall on the sound track during periods of low or no modulation. In the class-B push-pull optical system, the bias lines are created by the tail of the image in the optical mask as shown at (d) in Fig. 18-77.

18.61 Define the terms "attack time" and "release time."—These are the times required for a noise-reduction amplifier to open or close. Standard terminology in the industry defines attack time for a noise-reduction amplifier to be the elapsed time from the instant of application of the audio signal until 90 percent of full opening or deflection of the light modulator is achieved. Release time is the elapsed time from the instant of removal of the signal until 90 percent of the complete closing time. These terms apply equally to biased galvanometers, shutters, or to a light valve, and variable-area or variable-density recording. Three noise-reduction system opening tests are shown in Fig. 18-61. At (a) the opening time is 15 milliseconds; at (b) 30 milliseconds; and at (c) 28 milliseconds.

The first and second tests are for noise-reduction shutters, the third for a double bilateral track, the recorder using a biased galvanometer. The opening time test of a compressor having an opening time of 3 milliseconds is shown in Fig. 18-100. Only a negative is required for these tests.

18.62 What is decay or release time? — Time required for a noise-reduction system or compressor-limiter amplifier to close or return to a steady state.

18.63 What are the average attack times and release times for a noise-reduction system?—The average times for both a light valve and a galvanometer are tabulated in Fig. 18.63.

18.64 What is the effect if the attack time is too long?—An excessive amount of modulations will be clipped at the start of the sound track causing distortion.

18.65 How are the opening and closing times of a noise-reduction system measured?—To illustrate the procedure for this measurement, a typical variable-area noise-reduction timing test circuit is shown in Fig. 18-65A.

A 1000-Hz signal is applied to the input of the noise-reduction system am-

LIGHT VALVES

Type	*Opening Time (Milliseconds)*	*Closing Time (Milliseconds)*
Variable-area Standard Single Track	16 to 23	180 to 260
Variable-area Push-pull	7 to 10	180 to 260
Variable-density Standard Single Track	16 to 22	50 to 80
Variable-density Push-pull	7 to 10	50 to 80

GALVANOMETERS

Type	*Opening Time (Milliseconds)*	*Closing Time*
Variable-area Standard Biased Galvo.	28	4 in. of track
Variable-area Standard Shutters	13 to 42	3.5 to 4.5 in. of track

Fig. 18-63. Tabulation for average attack and release times.

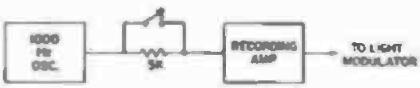


Fig. 18-65A. A circuit for measuring the opening and closing times of a noise-reduction system.

plifier and rapidly broken. If the oscillator used for the test is accurately calibrated for 1000 Hz, each modulation on the film will represent one millisecond of time.

The noise-reduction opening time is measured by means of a calibrated microscope from a normal bias line width to a modulation representing 90 percent of the full opening. The number of modulations between these two points of measurement is the time in milliseconds required for the noise reduction system to open from a normal bias line to 90 percent of full opening.

Release time is measured in a similar manner, except the time is taken from the last full amplitude modulation to a point on the bias line where the line returns to its normal width.

The opening time and characteristics for a pair of *maladjusted* noise-reduction shutters are shown in Fig. 18-65B. Note the bounce in the first few modulations before the shutters completely open. Only a negative is required for timing tests. A negative closing test is shown in Fig. 18-65C.

Fig. 18-65B. A maladjusted noise-reduction shutter opening test, showing shutter bump at the start of the opening (negative).

Fig. 18-65C. A noise-reduction shutter closing test (negative).

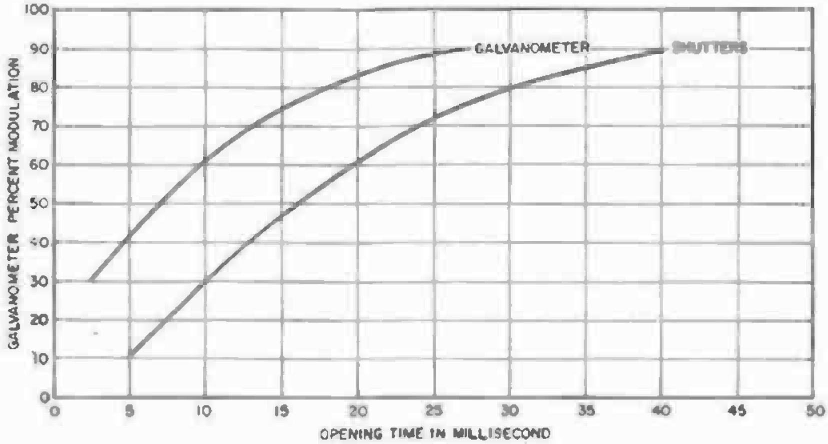


Fig. 18-65D. Opening time for noise-reduction shutters as compared to that of a biased galvanometer.

The difference in the opening time for a biased galvanometer and noise reduction shutters is shown in Fig. 18-65D.

18.66 What is a noise-reduction thump?—A low-frequency thumping noise caused by a noise-reduction system when the attack time is too short. (See Question 18.65.)

18.67 How can noise-reduction thump be checked?—If the noise-reduction system attack or release time is not correct, it is possible that a thumping noise will be heard in the reproduction, particularly in dialogue. If the recorder is equipped with a phototube monitoring system, thumping can be checked by patching out the audio signal and listening to the opening and closing of the noise-reduction system. If a photocell monitoring system is not available, dialogue is recorded with the audio leads to the light modulator disconnected. When this track is played back, if noise reduction thumps or breathing noises are present, they will be heard between pauses in the dialogue. Noise-reduction shutters are checked in a similar manner. If the channel includes a 40-Hz high-pass filter, it should be removed from the circuit for both the recording and reproduction tests.

18.68 What effect does noise-reduction bias current have on the frequency response of a galvanometer or light valve?—It has no effect if the current is an inverse of the input signal-level change. For a variable-area NRA, this may be measured by connecting the amplifier as shown in Fig. 18-68. The dc output is terminated in a resistive load of 600 to 1000 ohms. The bias current is adjusted for 30 mA without an input signal. A frequency of 1000 Hz is then applied to the input and increased for an indication of about 5 mA on the bias-current meter.

Constant-amplitude signals from 30 Hz to 10,000 Hz are then applied to the amplifier inpnt, and the bias current read for each frequency of interest. If the hias current does not remain constant, the input level should be increased or decreased 1 dB. For a reference current of 5 mA, the variation in milliamperes of bias current versus frequency should remain well within 1 dB over the entire frequency range.

The frequency-response characteristics are plotted thus: frequency versus decibels, as for any other type amplifier. The same method of measurement can also be applied to a light valve, except that the initial current and the closure current will vary with different type light valves.

18.69 Is noise reduction required with a variable-area class-B recorder?—No. A class-B recording system is inherently a noiseless recording system

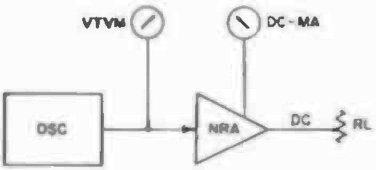


Fig. 18-68. Test circuit for measuring the frequency response of a noise-reduction amplifier (NRA).

because the bias line disappears during the periods of no modulation. However, due to slight variations in the optical and modulation systems, it has been found advisable to permit a small amount of light to fall on the film between modulations, thus creating bias lines. Generally, they are in the order of 1 mil and are visible in Fig. 18-29B.

The bias lines in a class-B variable-area recording system are created by the image mask and not by a bias current flowing through a bias coil as used in a normal recording system. Class-AB recording systems, illustrated at (e) in Fig. 18-29, also require no noise-reduction equipment.

18.70 What is margin?—Margin is the amount of gain the noise-reduction amplifier is increased after setting the noise-reduction bias current. It is expressed in decibels. The increased gain causes the noise-reduction system to open a greater amount than is actually required, thus reducing clipping of the initial modulations. The amount of margin (lead) required will vary with the type recording system, program material, and the attack time of the noise-reduction amplifier. Typical margius used in the industry are: variable-area, 3 to 6 dB; variable-density, 2 to 4 dB.

18.71 What is an anticipatory noise-reduction system?—An early noise-reduction system used with direct-positive recording. A separate sound track of the noise-reduction envelope was recorded in advance of the normal sound track at a sufficient distance to compensate for the delay-time of the noise-reduction amplifier. This system is now obsolete.

18.72 Show a schematic diagram for a noise-reduction amplifier (NRA) suitable for variable-area or variable-density recording.—Noise-reduction amplifiers (NRA) have undergoue little change in design over the past years, and are much the same except for a few minor changes. A schematic diagram for a Westrex Model 1610-A high-frequency carrier type NRA is shown in Fig. 18-72A. Its design is suitable for recording either variable-area or variable-density sound tracks, and standard or push-pull sound tracks.

Starting at the input, the signal is brought from the recording amplifier to the input through terminals 2 and 3 of J1, and out through terminals 1 and 3. The external circuit, 1 and 3, goes through the light-valve equalizer and to the light valve. (See Fig. 18-26E.) With light-valve switch S1 in the on position, the input to the NRA is bridged across the recording line. Since the bridging input is 10,000 ohms, the bridgiug loss is less than 0.5 dB. With the light-valve switch in the off position, the circuits to the light-valve equalizer and to the noise reduction circuit are shorted, and a 600-ohm termination is applied to the recording line. The signal now goes through the input control R2, through transformer T1 and auxiliary input control R4, which is used for adjusting the range of R2.

Tube V1-A is an amplifier stage and V1-B is a cathode-follower output stage. The output from this stage passes through transformer T2 and through a full-wave rectifier V2. The low-pass filter at the output of V2 supplies an attack time of approximately 18 to 22 milliseconds, and a release time of approximately 85 to 100 milliseconds. Tube V3-A is an LC oscillator, operating at a frequency of 30 kHz, with the output being controlled by R17. The output of the oscillator is modulated in V3-B by the rectified signal envelope from the low-pass filter in the output of V2. The 30 kHz carrier voltage at the output of V3-B decreases as the applied signal envelope dc voltage increases. The carrier output is controlled by R18. Control R20 adjusts the range of R18.

The 30-kHz output of modulator V3-B is applied to the grid of V4, amplified, and then stepped down by transformer T3 and rectified by CR1 and CR2. The rectified 30-kHz signal is then filtered and brought to the meter shunt R24, the noise-reduction switch S4, and to terminals 10, and 11, of J1. The V.A. test position of S4 applies reverse bias to fully open the valve ribbon, and provides a wide exposed track for sensitometric measurements, when recording variable-area tracks.

Bias meter M1 is a zero-center movement, calibrated in 100 divisions each side of center. Four ranges of current for the meter are provided by switch S3-A. The second section of S3-B increases the gain of V4 for the two higher scales of the meter positious.

In use, certain controls are placed at a remote position as pictured in Fig.

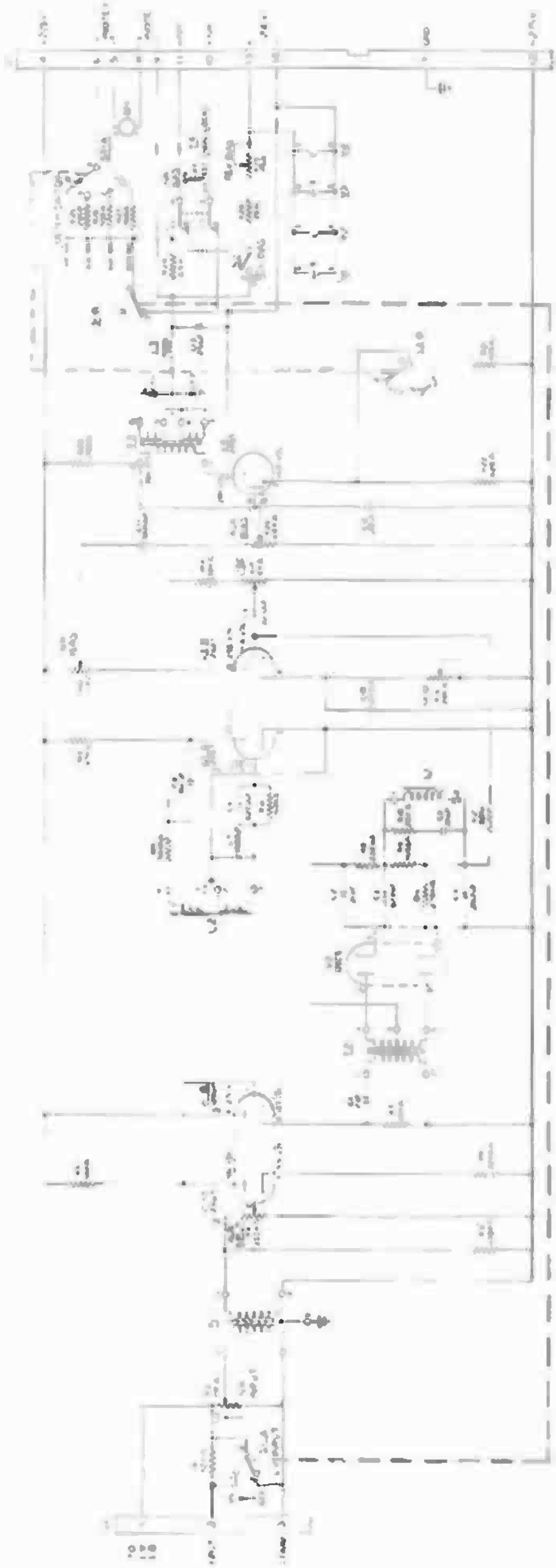


Fig. 18-72A. Westrex Corp. Model 1610-A noise-reduction amplifier (NRA).

18-26D, and are connected to terminal strip (J1). Light valves in current use require a noise-reduction current of 52 to 600 milliamperes, depending on the type used. The frequency response of the amplifier is plus or minus 1 dB 50 to 5000 Hz, and 1.5 dB from 5000 to 10,000 Hz, with reference to 1000 Hz.

A second noise-reduction amplifier schematic diagram is given in Fig. 18-72B. This NRA is designed for use with only variable-area sound track using a

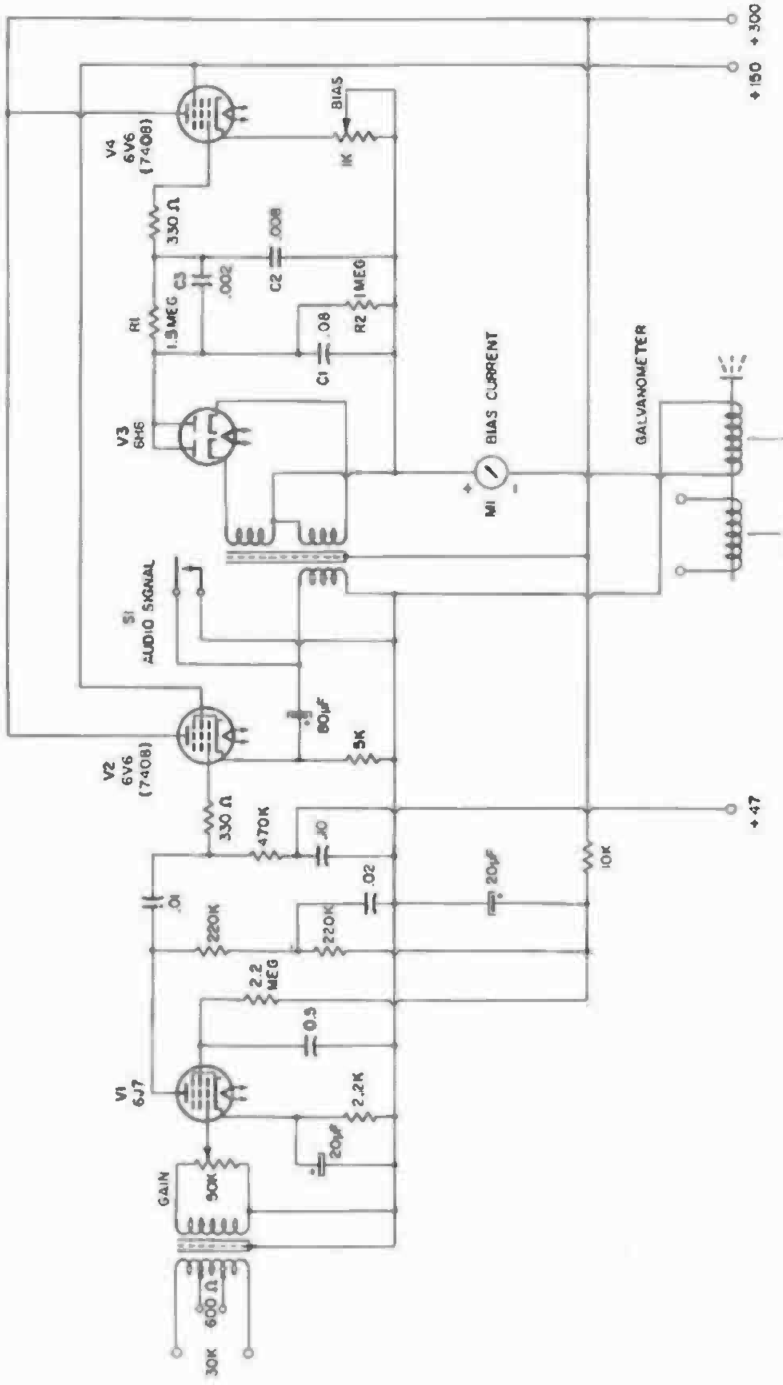


Fig. 18-72B. Noise-reduction amplifier for variable-area recording.

biased galvanometer or noise-reduction shutters. The circuitry consists of a voltage amplifier tube V1, a driver stage V2 connected as a cathode follower, a full-wave rectifier V3, and a dc amplifier stage V4. An input transformer T1 supplies both a matched 600-ohm and 25,000-ohm bridging input. Across the secondary of T1 is a gain control calibrated in 1-dB steps. The first stage is RC coupled to the second stage. The driver stage is transformer-coupled to the rectifier stage. The driver stage is employed as a cathode follower in order to present a low impedance to the rectifier input transformer, thus permitting the charging of capacitor C7 from a very low impedance to retain good peak-reading ability. The output from rectifier V3 is filtered and supplied as a control voltage to the grid of V4, the dc amplifier stage. The load (galvanometer NRA coil or shutter coils) is connected in series in the negative leg of the high voltage. The initial or no-signal bias current is indicated in the meter M1. Key switch S1 is used to short-circuit the audio signal for checking the bias current. The power supply must be of good regulation and capable of supplying fairly large amounts of current without affecting the operating voltages to any great extent. A plus voltage of 47 volts is supplied to the grid return of V2. Because the cathode is plus 57 volts above ground, tbis makes the control grid of V2 10 volts negative with respect to the cathode.

18.73 What is the width of the bias line for 35-mm variable-area recorders? —For standard sound tracks, the width of the bias line is 2 mils.

18.74 What is the width of the bias lines for 16-mm variable-area recording? —The width of the bias line is 2 to 4 mils. However, if it is possible, the width should be 2 mils. In the early model 16-mm recorders, the bias line was generally on the order of 4 to 6 mils, due to the difficulty of maintaining a 2-mil line.

18.75 What is the width of the bias lines for a 35-mm class-B, variable-area recorder?—Unmodulated, 0.75 to 1 mil. Modulated, 0.25 to 0.35 mil. For single system class-B newsreel cameras, 2 to 2.5 mils (unmodulated) and 1 to 1.5 mils (modulated). The bias lines for a class-B recording may be clearly seen in Fig. 18-298.

18.76 Give the connections for a variable-density noise-reduction amplifier.—Such a diagram is shown in Fig. 18-76. It is somewhat similar to that of the variable-area noise-reduction amplifier described in Question 18.72. The principal difference is the method used for applying the bias current to the light modulator.

The positive side of the bias current is connected to a center tap on the light valve impedance-matching transformer. The negative side of the bias current is connected to the junction of the light valve and a resistance R, which is equal in value to the light valve impedance. This method of connection forms a bridge circuit which is unbalanced by the action of the noise-reduction current.

With the noise-reduction amplifier properly set and no modulation, the negative exposure appears as a light gray exposure the full width of the sound-recording slit. The density of exposure varies with the amount of bias current and the opening of the light-valve ribbous.

18.77 Describe an optical mask.— A small metal mask in which is cut an

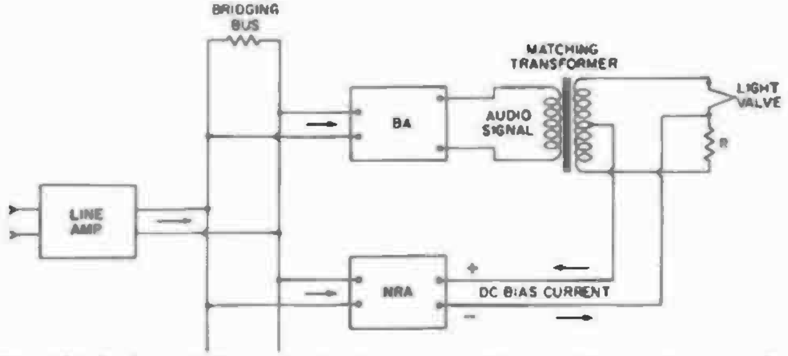
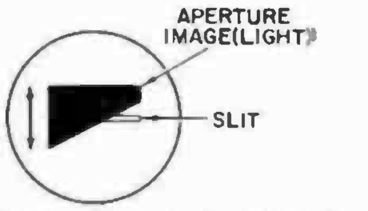


Fig. 18-76. A typical noise-reduction system for variable-density recording.

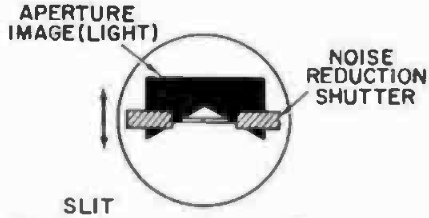
image for forming the sound track in a variable-area photographic film recorder.

The image mask is mounted at (c) in the optical train drawing of Fig. 18-338A. Typical sound track images used by RCA in their early- and present-day recorders are given in Fig. 18-77 (a) through (h).

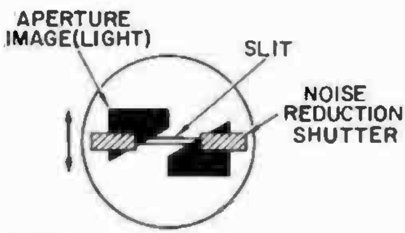
At (a) is a mask used in the first variable-area recorders. The sound tracks appeared as shown in Fig. 18-



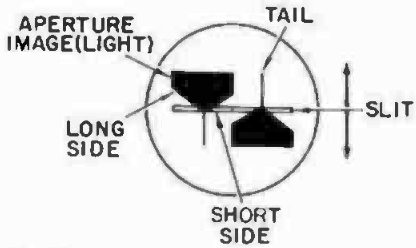
(a) *Unilateral negative 50% deflection.*



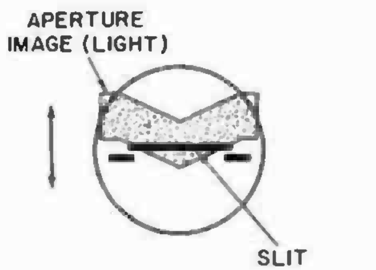
(b) *Duplex negative standby position.*



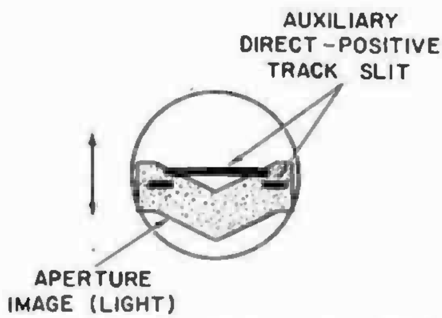
(c) *Class-A, push-pull negative standby position.*



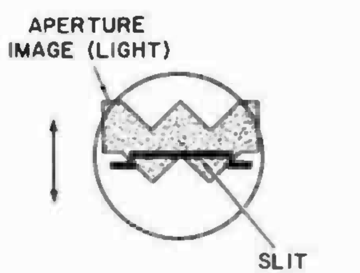
(d) *Class-B, push-pull negative standby position.*



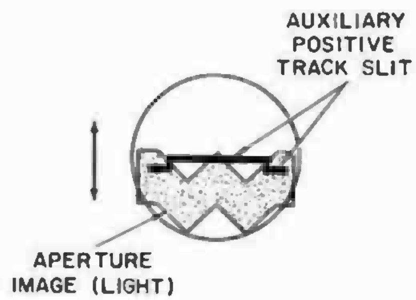
(e) *Bilateral negative 100% deflection.*



(f) *Bilateral direct-positive 100% deflection.*



(g) *Double-bilateral negative 100% deflection.*



(h) *Double-bilateral direct-positive 100% deflection.*

Fig. 18-77. Sound track images used by RCA both past and present, for negative and direct-positive recording. The black areas represent the light image as seen looking at the illuminated end of the optical system.

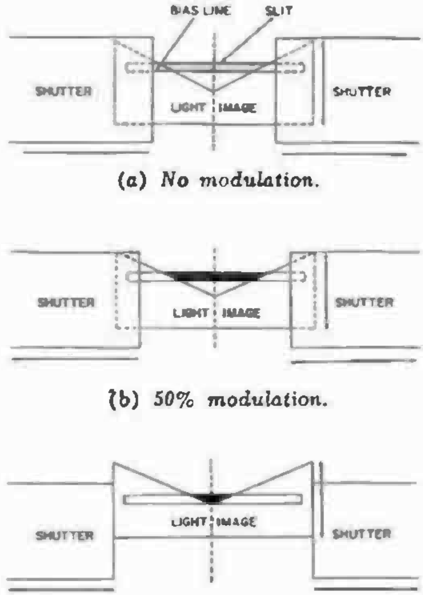
29A. At (b) of Fig. 18-77 is a duplex mask which produces a sound track as shown in (c) of Fig. 18-29.

The image mask at (d) in Fig. 18-77 is used to create the class-B push-pull track shown at (f) in Fig. 18-29. The class-AB mask has a similar appearance, except the angles of the long and short sides of the image are slightly different. The tails in both masks form the bias lines. A bilateral negative image appears in Fig. 18-77e, and for direct-positive recording in (f). A double-bilateral image for negative recording appears at (g), and for direct-positive recording at (h).

Bilateral and double-bilateral images are used for both 16- and 35-mm recording and are the present type in use. The signal-to-noise ratio for the double bilateral is somewhat better with lower distortion and higher output. Either type may be used for recording negative or direct-positive by tilting the galvanometer image above or below the recording slit, as shown in (f) and (h) of Fig. 18-77. Basically, the image recorded on the film for the double-bilateral track operates in the same manner as for the bilateral, as explained in Question 18.78. For direct-positive recording the process is reversed.

18.78 Describe the operation of a galvanometer with noise-reduction shutters.—An image that will form the desired sound track image is cut in the optical mask. This image is projected by an objective lens on the galvanometer mirror which reflects the image of the mask on the mechanical slit at the rear of the recording optical system. The mask image is deflected upward and downward over the face of the recording slit by the motion of the galvanometer which is caused to move by the applied audio frequency currents. The galvanometer mirror (and the light beam) are deflected in proportion to the applied signal voltage. The film is in motion, and, as the galvanometer moves the image across the slit, an oscillographic-type sound track is exposed on the film.

As the galvanometer is modulated, the noise-reduction shutters are moved in and out, the amount depending on the percentage modulation of the galvanometer, as the galvanometer, amplifier, and noise-reduction shutter amplifiers receive the audio signal simultaneously.



(a) No modulation.

(b) 50% modulation.

(c) 100% modulation.

Fig. 18-78. Light image positions for a standard duplex variable-area sound track and noise-reduction shutters.

For periods of low modulation, the shutters are moved inward and as the percentage of modulation increases, they move outward.

The mechanical slit at the rear of an optical system, with an image for a standard duplex variable-area sound track projected across the face of the slit, is shown at (a) in Fig. 18-78. (This is the same as shown at (b) in Fig. 18-77 except the image is inverted.) With the galvanometer recorder switch in the record position, the noise reduction shutters are pulled inward by the fixed dc bias from the NRA, to a half-track position. It will be noted the shutters are closed down to a position where only a very small amount of light is permitted to pass through the recording slit to the film. These small beams of light at the upper corners form the bias lines when the galvanometer is in a steady-state condition. When viewed through a periscope placed in the optical system, they appear as two pinpoints of light and are often referred to as "snake eyes."

The light image appears as shown at (c) in Fig. 18-78 when the galvanometer is modulated 100 percent. Here it will be noted the shutters are pulled outward beyond the ends of the slit and the light image has moved upward to a position where the slit is completely covered, except for the small dark portion in the center of the slit. The center portion of the slit is purposely left dark so as not to expose the film in the center. This prevents clipping of the modulation peaks. If the slit were completely covered with light, the peaks of the negative image would meet in the center. When printed, they are reversed. Thus, the modulation peaks would be clipped. As peak clipping causes excessive harmonic distortion the peaks are prevented from meeting by leaving the center portion of the slit dark. The position of the light image for a condition of 50-percent modulation is shown at (b) in Fig. 18-78. The light image is in the same position as for the standby position, except the noise-reduction shutters have been pulled partially outward by the action of the noise-reduction bias current. Only the ends of the slit are illuminated. The center portion is dark. For 25-percent modulation, the light image does not rise as far as for 50-percent modulation. Also, the noise-reduction shutters do not open as wide.

Because the recording amplifier that drives the galvanometer and the noise-reduction amplifier receive the signal simultaneously, the noise-reduction amplifier pulls the shutters outward as the percentage of modulation increases. However, the shutters are never completely out of the way; thus, the first few modulation peaks are clipped and appear as shown at (a) and (b) in Fig. 18-61.

18.79 How does the optical mask form the sound track image when using a biased galvanometer?—Fig. 18-79 shows the image reflected from the galvanometer mirror as seen on the end of the optical system barrel. The lower point of the reflected image is brought to the edge of the slit by adjustment of the fixed bias in the noise-reduction amplifier. The width of the bias line is determined by recording several current settings, developing the negative, and then measuring the width of the line.

With no modulation applied to the galvanometer, the image rests as shown by the solid lines. With modulation applied, the image moves up, down, and around a mean average, depending on the amount of cancellation of the bias

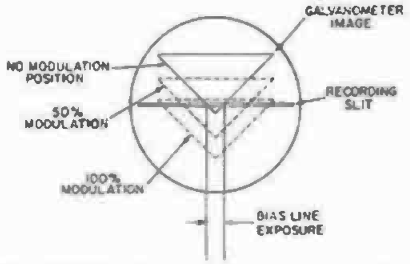


Fig. 18-79. Biased galvanometer using a bilateral sound mask.

current and the percentage of modulation. In no case should the modulation be great enough to pull the image upward to the point where the bias-line exposure is removed. This would result in a loss of the bias line. Two other positions of modulation are shown by the dotted lines. As it may be seen, the area of exposure across the slit is varied with the percentage of modulation. With the film traveling at a constant velocity past the slit, an oscillographic picture is obtained on the film.

18.80 Is an optical mask required with a light valve?—Yes, a mask is required. The type mask used depends on the type light valve used. In the Westrex Corp. recorders, the appropriate mask is put in place by turning a thumb screw in the modulator unit in which the light valve is mounted.

Different type light valves are used for variable-area recording than are used for variable-density push-pull recording. For variable-area recording, the light-valve ribbons are placed in a vertical plane, as may be seen in Fig. 18-26B. For variable-density recording the ribbons are placed in a horizontal plane.

18.81 What are the factors governing the recording of dialogue?—Because dialogue is generally reproduced at a higher sound level than that at which it was originally recorded, it will sound boomy and unnatural if the human ear characteristic is not considered when making the original recording. To obtain a listening quality pleasing to the ear, dialogue equalization is used in the recording channel as shown in Fig. 18-81. It will be noted the frequencies below 800 Hz are slowly attenuated to reduce the amplitude of 100 Hz 8 to 12 dB compared to the level of 1000 Hz. The frequencies above 1000 Hz are accentuated to add presence to the voice.

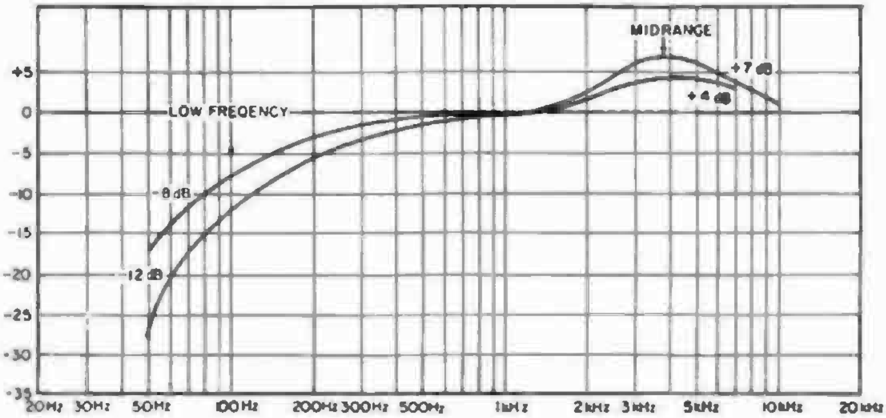


Fig. 18-81. Dialogue equalization cures.

However, the amount of midhigh frequency equalization will be governed by the stage acoustics, the type microphone employed, and the type recording channel used. As a rule, each recording activity develops its own standards of equalization to produce an overall frequency response which will meet the accepted reproducing standards of the industry. Meeting the standards of reproduction is highly important because theaters throughout the world are treated acoustically and the sound systems equalized electrically to produce a high rate of intelligibility, and to provide the overall best reproduction possible. The low end of the dialogue must be rolled off in order to achieve intelligibility.

In addition to the equalization, compression may be used when rerecording which also induces a certain characteristic. This subject is discussed in Question 18.84. Microphone placement is discussed in Question 4-114, and dialogue equalization is treated in Question 6.122. (See Question 18.169.)

18.82 Is the peak energy of music and dialogue the same as for a sine wave?—No. The peak energy of music and speech is from 8 to 12 dB greater than that of a sine wave for a given level. A complex waveform compared to a sine wave is shown in Fig. 18-82. This subject is further discussed in Question 17.163.

18.83 Are the waveforms of music and dialogue of sine wave character?—No, the greater percentage of recorded sound is not symmetrical in character as the pressure half of the waveform is generally greater in amplitude. It is highly important in an optical film recording system that the complete recording channel starting at the microphone and continuing through to the light modulator be in phase electrically and that if any piece of equipment is removed its phasing be checked before putting it into permanent operation. Phasing is discussed in Question 23.104.

18.84 What is compression?—A signal is said to be compressed when it is reduced or held within a given amplitude at the output of a device, for a given increase in signal level at the input. Compressor amplifiers are employed with recording systems, particularly with the optical recording system, to prevent the overloading of the light modulator. Compression is of great value when recording high-level dialogue or sound effects. Compression should not be confused with overloading of the system. When properly applied, compression reduces overloading distortion, and generally improves the overall recording characteristic.

Experience has shown that certain types of recorded dialogue and sound effects are of such volume range they cannot be reproduced to the best advantage in a theater. When reproduced, the loud passages are too loud and the

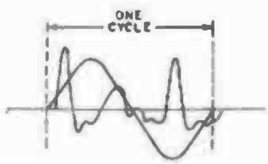


Fig. 18-82. One cycle of a complex waveform superimposed on one cycle of a sine wave.

low passages are too low. Many times the low-level passages are lost in the ambient noise level of the theater. Also, the high-level material appears exaggerated, producing an unnatural quality. Because manual compression of program material is impractical, electronic compressors (sometimes called electronic mixers) have been developed which do a very satisfactory job. Although a compressor restricts the volume range to some extent, it aids in flattening out excessive peaks and valleys in the overall characteristic. Compression is seldom used with music recording, although a compressor can be used as a limiter for supplying 1 to 3 dB of limiting when transferring magnetic sound track to photographic sound track. This is necessary because of the greater dynamic range of the magnetic track, compared to the optical sound track. Variable-area photographic sound track is limited by the mechanical deflection of the galvanometer, while variable-density is limited by the exposure. Limiting is also used in the transfer of magnetic sound track to disc records. Here again, the peaks must be controlled to permit increasing the average level to an acceptable value. (See Question 18.101.)

18.85 What are the characteristics of the average compression used for film recording?—Two types of compression are in general usage, gradual and limiting. Gradual compression is set to give a slope of 2 to 1 as shown by the curve (a) in Fig. 18-85. Limiting curves have slopes between 4 and 6 to 1 as shown by curve (b). Compression ratios are expressed by stating the breakaway point which is relative to 100-percent modulation of the system. The breakaway point is defined as that point where compression can just be observed or it affects the system 0.5 dB. Curve (a) represents a compression ratio of 20 dB into 10 dB at an output level of plus 4 dBm. Curve (b) is a ratio of 17 dB into 3 dB at an output level of plus 4 dBm. Curve (c) is the input-output characteristic of the compressor amplifier uncompressed below its overload point.

18.86 Show a schematic diagram for a compressor amplifier.—The schematic diagram for a Westrex Model RA-1593A compressor is given in Fig. 18-86A, and its control panel RA-1594A is shown in Fig. 18-86B. Fig. 18-86A shows that the amplifier consists of three push-pull stages connected in tandem. The first stage is comprised of two 12BA6 remote cutoff tubes, V1 and V2. The second stage, V3A and V3B, is transformer-coupled to the output stage V4A and V4B. Compression and limiting are obtained by varying the dc bias applied to V1A and V2A. At the lower part of the

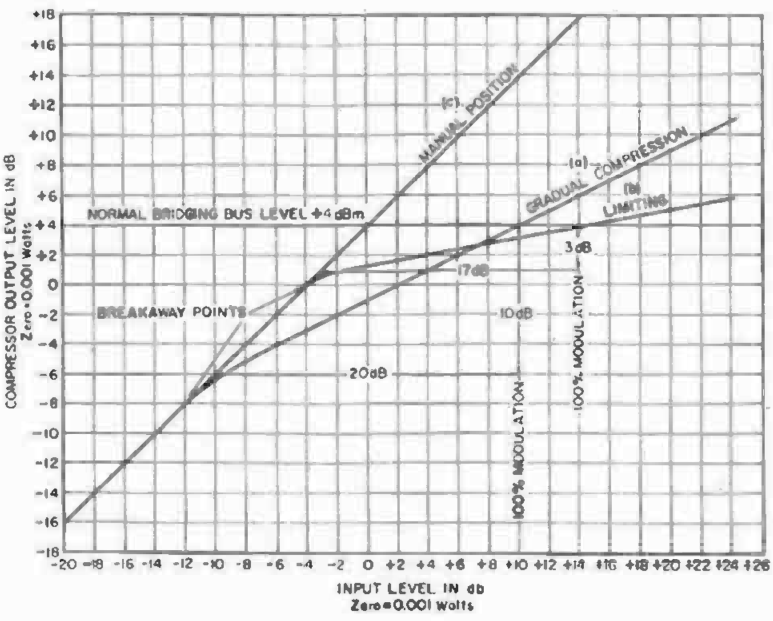


Fig. 18-85. Compressor characteristics.

diagram is a control or side amplifier. The input and output circuits are transformer-coupled and are designed to operate from and into 600-ohm circuits. The input signal for the control amplifier is taken from one side of the output transformer T3, through capacitor C9 and resistor R31.

The control amplifier consists of tubes V7A and V7B, with the second

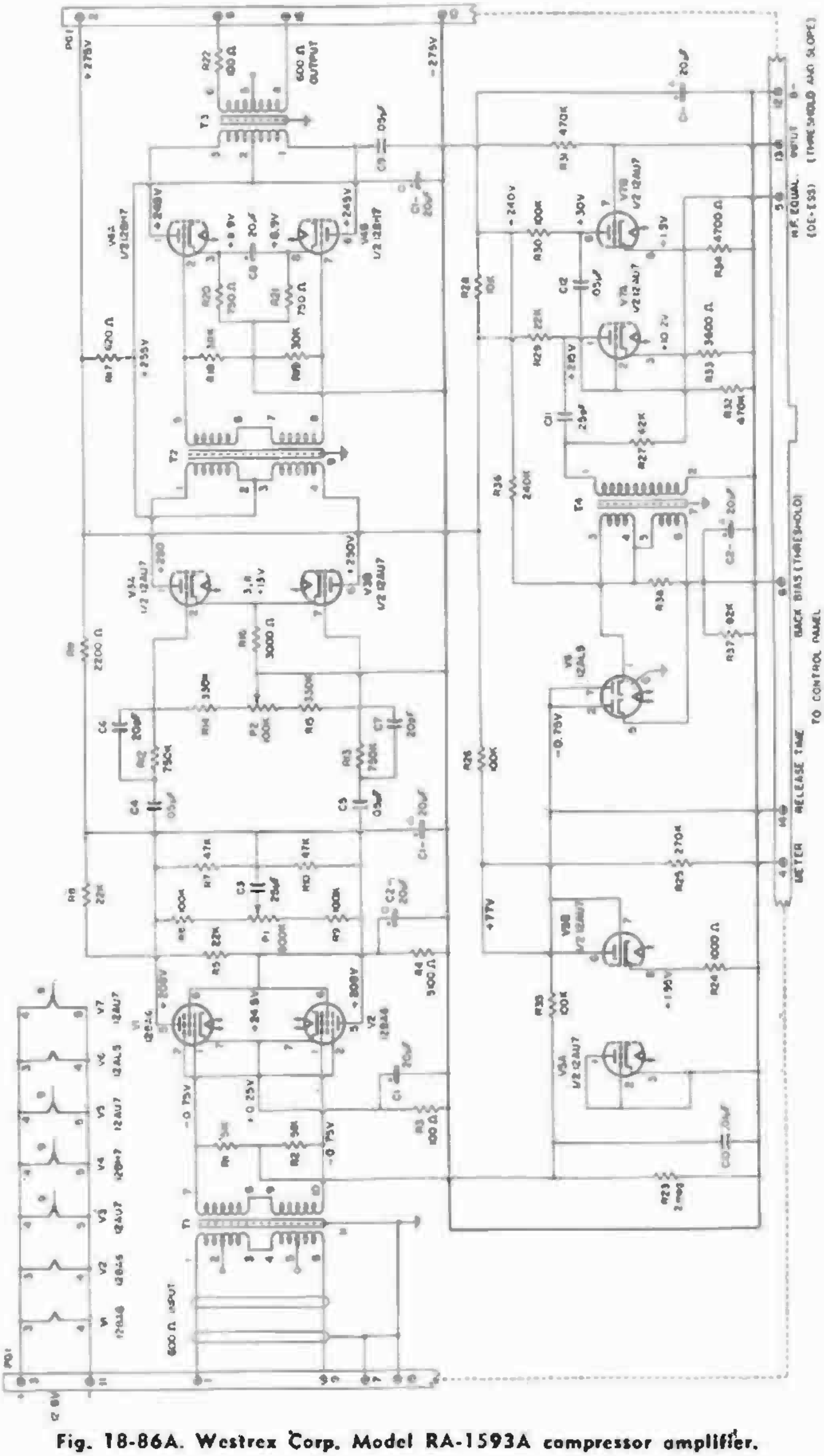


Fig. 18-86A. Westrex Corp. Model RA-1593A compressor amplifier.

stage transformer-coupled to a full-wave rectifier V6. The dc voltage from the rectifier is fed to the control-grid return circuit of V1A and V1B, which acts as a variable gain stage. A dc amplifier is connected across the output of rectifier V6 to drive the meter or meters which indicate the amount of compression. Potentiometers P1 and P2 provide a means of dynamically balancing the stage following the compression stage, to minimize transients (generally referred to as thumps) resulting from the compression of a steep wave-front signal.

The operating controls are mounted on remote control panel RA-1594A (Fig. 18-86B). A selector switch in this panel provides a means of varying the compression slope, and the threshold at which compression starts. The compression meter is calibrated from 0 to 15 dB, and it indicates the amount of compression while the compressor is operating.

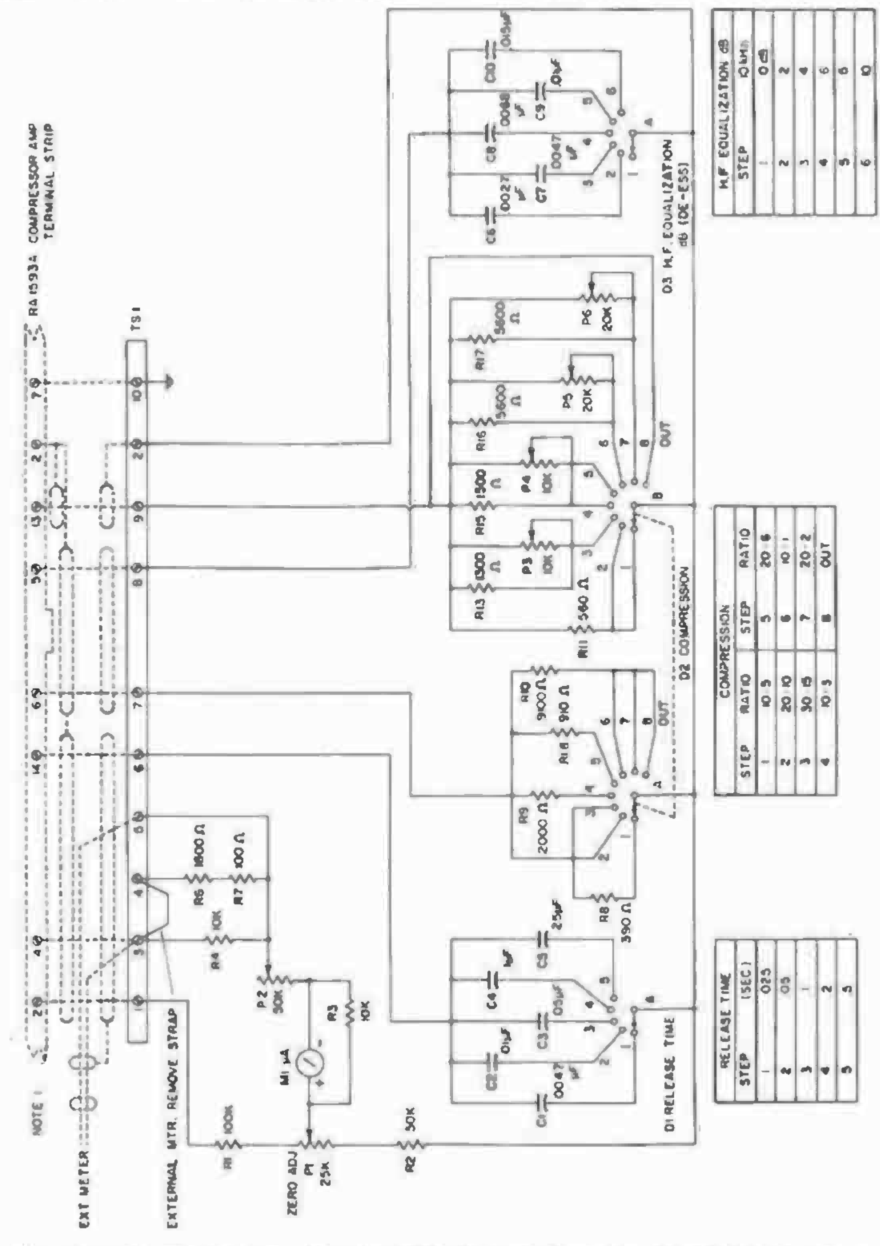


Fig. 18-86B. Westrex Corp. Model RA-1594A control unit for RA-1593A compressor amplifier shown in Fig. 18-86A.

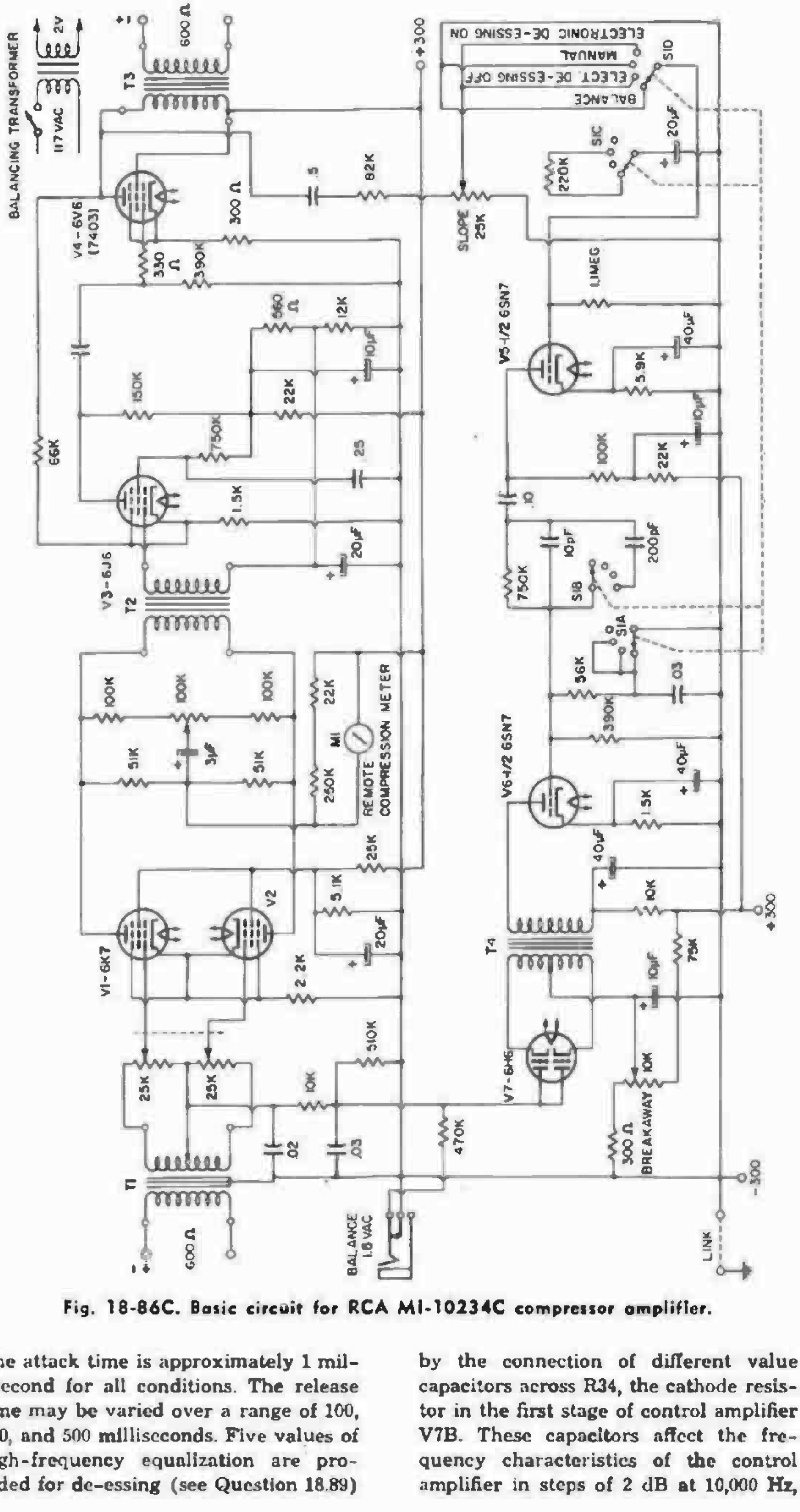


Fig. 18-86C. Basic circuit for RCA MI-10234C compressor amplifier.

The attack time is approximately 1 millisecond for all conditions. The release time may be varied over a range of 100, 250, and 500 milliseconds. Five values of high-frequency equalization are provided for de-essing (see Question 18.89) by the connection of different value capacitors across R34, the cathode resistor in the first stage of control amplifier V7B. These capacitors affect the frequency characteristics of the control amplifier in steps of 2 dB at 10,000 Hz, causing the high frequencies to be compressed more than the lower frequencies. A flat position is also provided. The range of compression, release time, and high-frequency equalization is given in Fig. 18-86B. A special 20,000-Hz transistor oscillator automatically keyed on and off at a level sufficient to drive the amplifier into compression by approximately 2 dB is patched into the input and used to balance the compression amplifier. A VU meter is connected across the output of the compressor amplifier, in parallel with a circuit containing a 200-ohm resistor and a 4-μF capacitor connected in shunt. Potentiometers P1 and P2 are then adjusted for a minimum thump on the VU meter.

The amount of compression to be used will be governed by the type of program material to be recorded. A properly designed compressor-limiter amplifier adds very little distortion to the overall recording and may be disregarded when compared to the overall results obtained.

A second schematic diagram for a compressor amplifier appears in Fig. 18-86C and is similar to that in Fig. 18-86A, except for certain circuit changes. The first stage consists of two 6K7 variable-mu tubes, V1 and V2, connected push-pull. A dual input attenuator, calibrated in 1-dB steps, is connected across the secondary. It will be observed that a fixed resistor, R5, is connected in the cathode circuit of these tubes to establish an initial bias. In the plate circuit of the tubes is connected balancing potentiometer R9. The first stage is transformer-coupled to a two-stage amplifier, and then to an output transformer.

At the lower portion of the diagram is the control or side-amplifier, consisting of a two-stage amplifier and a rectifier. The center tap of the terminating resistors across the secondary of the input transformer T1 is returned through an RC network to the rectifier V7. With an audio signal at the primary of the input transformer T1, the secondary applies the signal to the control-grids of V1 and V2 simultaneously, but 180 degrees out of phase. The signal is amplified and passed on to the second and third stages and then to output transformer T3. A portion of the output signal is fed to the control amplifier input stage V5, through capacitor C10 and resistor R28, to slope control R29. After amplification, the signal is again amplified by V6 and rectified by full-wave rectifier V7. Here the audio signal after rectification is applied to the grid-return circuit of V1 and V2. As the amplitude of the signal rises and falls, the variable bias voltage supplied by the diode rectifier rises and falls in proportion to the average value of the input signal. Thus, the gain of the input stage is controlled, holding the output to a predetermined level as set by the adjustment of the threshold, slope, and compression controls.

The amplifier may be dynamically balanced to remove the effect of transients and the tendency to thump by applying a 60-Hz signal of approximately 2 Vac to the grid return of the first stage by means of the balance voltage jack. Because the balancing signal is connected in series with the return circuit of control grids of V1 and V2, the signals at the grid of each tube will be in phase; that is, both grids will have the same instantaneous polarity. Therefore, the signal at the plates will also be in phase. For all practical purposes, the signal can be completely balanced out by connecting a vacuum-tube voltmeter and a 600-ohm terminating resistor across the output winding of transformer T3, and adjusting potentiometer R9 for a minimum reading on the vacuum-tube voltmeter. The minimum level is greater than minus 52 dBm.

The restoring time is 100 milliseconds for 99-percent restoration. This may be increased to 500 milliseconds by changing resistor R4 in the grid-return lead of V1 and V2 to a value of 2.4 megohms. The power supply, although not shown, is straight forward with low ripple voltage and good regulation. In the uncompressed position (Manual), the frequency response is plus or minus 1 dB, 20 to 10,000 Hz. In the compressed position, it will approximate that shown in Fig. 18-91. Total harmonic distortion in the uncompressed position is less than 0.7 percent for an output of plus 28 dBm. Harmonic distortion in the compressed position, using a compression ratio of 20 into 10, is 0.9 percent with a 100-millisecond release time, or 0.2 percent for a 500-millisecond release time. Both measurements are at 400 Hz. The reference frequency for testing compression ratios of the amplifier in Fig. 18-86C is 300 to 400 Hz.

18.87 What does the term 20 into 10 compression mean?—It is a term used in stating the characteristics of a compressor amplifier and means that for an increase of 20 dB in level at the input the output level will only increase 10 dB.

18.88 What are the compression ratios most commonly used for motion picture rerecording and transfer?—30 into 10, 20 into 10, and 30 into 15. For transfer work only limiting is used, with the limiter set for 1 to 2 dB at 95-percent modulation of the light modulator.

18.89 What does the term "de-essing" mean?—It means that the high frequencies have been compressed a greater amount than the lower frequencies to remove sibilance. This is accomplished by equalizing the rectifier amplifier (side amplifier) so that greater compression is obtained above 1000 Hz than below 1000 Hz.

It has been shown from experience that a more natural and pleasing reproduction of dialogue can be achieved if the tendencies toward accentuation of sibilants is reduced or eliminated. Equalization used in compressors for this purpose is derived from consideration of the average relative-spectral energy distribution of the male and female voice, speaking in the English language. For languages other than English, it may be necessary that the characteristics of the compressor equalizer be changed slightly. Equalizers used for the suppression or elimination of sibilants are termed "de-essing equalizers" and are connected in the control or side amplifier of the compressor.

18.90 When are compressors and limiters essential? — With the present policy of recording original dialogue and music on magnetic tape or film, and because of the dynamic range of magnetic sound tracks, it is essential that a compressor be employed when rerecording dialogue, and limiting be employed when transferring a master magnetic sound track to an optical sound track.

The overload point of variable-area sound track is reached quite suddenly, and peak clipping occurs, squaring-off the peaks of the waveforms. If the track is composed of dialogue or music, considerable distortion may result.

Variable-density sound track overloading is rather gradual; however, if the levels are high, compression or limiting is also called for. It is next to impossible to hold down manually a high-level dialogue track when rerecording, and still attain a fairly high average level of dialogue. This is particularly true for 16-mm film. (See Question 18.88.)

18.91 What is the frequency response of a compressor amplifier using a de-esser equalizer?—In Fig. 18-91 two compressor frequency characteristics are shown, with and without de-essing equalization. It will be noted that in the de-essing position the response at 10,000 Hz is reduced 9 dB, and at 40 Hz it is reduced 4 dB. Without de-essing the frequency characteristics are essentially flat.

18.92 What frequency is used when adjusting the compression ratio of a compressor amplifier?—For compressors using a de-esser equalizer and having a frequency characteristic as shown in Fig. 18-91, 300 to 400 Hz is used. Compressors with other frequency characteristics will require a frequency in the flat portion of the frequency response. If a de-esser equalizer is not used, either 400 Hz or 1000 Hz may be used.

The compression ratio should be checked before each recording session. This may be done by applying a normal signal to the recording channel and throwing the compression key from manual to compress and noting the drop

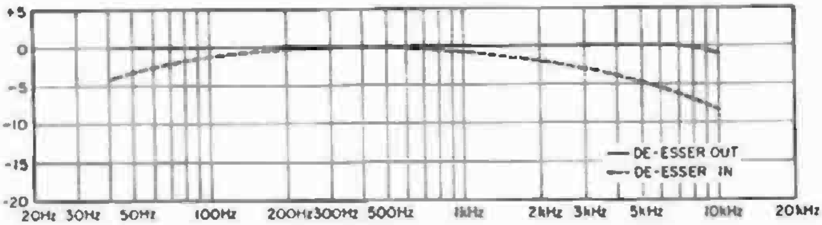


Fig. 18-91. Compressor de-essing characteristics.

in level on a VU meter. For a ratio of 20 into 10, this will be 5 dB; for 8 into 4, it will be 2 dB; and for 32 into 16, 8 dB.

18.93 Will the use of compression when set for a high ratio increase the harmonic distortion? — Only a small amount. The benefits gained by the use of compression are such that the slight increase of distortion is negligible.

18.94 When should compression be used? — When recording dialogue and certain types of sound effects. Generally, for dialogue about 20 into 10 will be suitable. Compression is seldom used for music recording; however, a small amount of limiting is sometimes used as a ceiling control.

When recording original dialogue sound tracks on magnetic tape or film, no compression is used. However, when rerecording, compression is used in the dialogue tracks. Limiting may be required when transferring from a dubbing master sound track to an optical sound track for release prints. (See Question 18.88.)

18.95 Describe the procedure for balancing a compressor amplifier.—For the RCA compressor amplifier, a 50- or 60-Hz signal is applied to the balancing voltage input, in series with the grid returns of the first-stage variable-mu tubes. A vacuum-tube voltmeter is connected across the output, and the balancing control (R9 in Fig. 18-86) is balanced for a minimum indication on the meter. As a rule, at least 50 dB or better is obtained. Indications of minus 60 dB are not uncommon.

To balance a Westrex compressor, a special oscillator is used that supplies a 20,000 Hz signal pulsed about once per second. While watching a meter connected across the output of the compressor, the balancing control is adjusted for a minimum thumping indication on the meter. If a compressor amplifier is not balanced, a breathing or thumping sound may be heard between pauses in dialogue. (See Question 18.86.)

18.96 What is the gain of a compressor amplifier?—As a rule, it is operated as a no-gain device. However, the manufacturer generally provides some means for increasing the gain so that the amplifier may serve the purpose of both a line and a compressor amplifier.

18.97 What precaution should be taken when using a compressor?—If the channel employs high- and low-pass filters, they must be connected after the compressor amplifier. This is to prevent a change in their frequency characteristics by the action of the compressor. Compressors have a tendency to smooth out the frequency characteristics of the recording system preceding it, resulting in a form of "automatic effort equalization." Thus, when the voice is raised in level, the amount of low-frequency attenuation is reduced, thus preventing the voice from becoming harsh and shrill. Normal low-frequency equalization used for dialogue prevents thumping due to unbalance that may exist in the push-pull stages of the compressor amplifier.

18.98 Where is a compressor amplifier normally connected in a recording channel? — For dialogue recording, at the output of the mixer network, as shown in Fig. 18-98. A separate pot called a ceiling control is connected in the output of the compressor and may be used to change the amount of compression. Changing the compression in this manner does not affect the slope of the compressor characteristic, only the amount of compression.

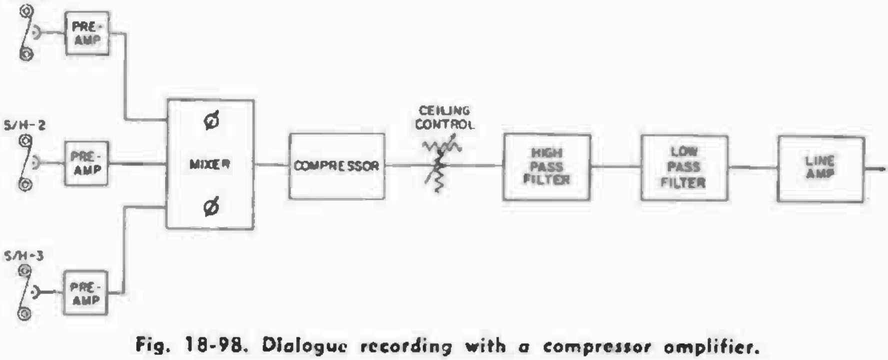


Fig. 18-98. Dialogue recording with a compressor amplifier.

18.99 At what percentage of modulation does a compressor take hold?—It will depend entirely on the setting of the compression ratio and the level selected for the breakaway point. Compression may be smoothly controlled over a wide range of levels.

18.100 What are the average attack and release times for a compressor amplifier?—Attack time: 1 to 10 milliseconds. Release time: 100 to 500 milliseconds. A typical timing test of the opening characteristics of a compressor amplifier is shown in Fig. 18-100. The test is made by applying 5000 Hz at an amplitude 20 dB below the normal input level and then suddenly applying a level that will result in the maximum output level. Not more than three peaks of the 5000-Hz modulations should be clipped. Only a negative sound track is required for the above test.

18.101 What is the difference between compression and limiting?—Compression is used to hold down or limit the peaks of dialogue in the ratio of 30 into 10, or 20 into 10; limiting is generally set for 1 to 3 dB and is used as a ceiling control when transferring from a high-level track to a photographic sound track. In both these instances, the dynamic and average levels are increased due to limiting of the peak values. (See Question 18.85.) Limiters are used extensively in broadcasting, as they permit the transmitter to be fully modulated, thus increasing the coverage without overmodulation.

18.102 What is a ceiling control?—A variable attenuator connected at the output of a compressor-limiter amplifier to control the amount of compression. Increasing the loss of the ceiling control causes the compressor amplifier to be driven harder to secure the same outpnt; thus, the amount of compression is increased. However, the slope characteristics are not affected.

18.103 What type volume indicator is recommended for photographic film recording?—The standard meter for monitoring is a VU meter with ballistic characteristics that meet the USASI (ASA) Standard C16.5-1961. However, for photographic film recording, a peak-indicating instrument, if available, should be used. It is adjusted to indicate the maximum deflection when the light modulator indicates 100-percent deflection. In this manner, the peak excursions of the light can be monitored. In the absence of a volume indicator of this type, an oscilloscope can be used as it is a peak-indicating instrument.

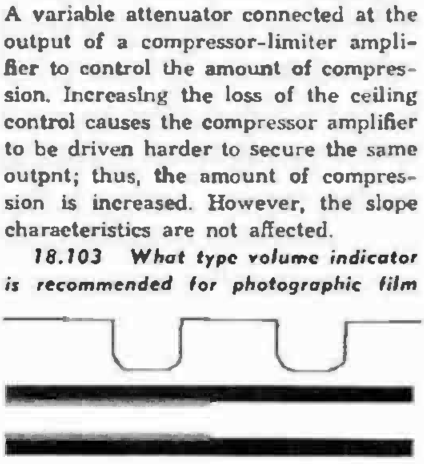


Fig. 18-100. Attack-time test for compressor amplifier, opening time 3 milliseconds.

If a peak-indicating device is not available, at least an 8-dB lead must be inserted in the VU meter circuit to prevent the overloading of a light valve or galvanometer after lining-up the channel. (See Questions 10.3, 10.7, and 17.163.)

18.104 What is a background-noise suppression amplifier (BNSA)?—It is a special type amplifier that operates somewhat like a noise-reduction amplifier (NRA) used for photographic film recording. However, in the instance of the background-noise suppression amplifier, it is connected early in the recording circuits to reduce unwanted noise in the sound track to be rerecorded. The amplifier is set up in such a manner that it closes down during pauses in the dialogue and other portions of the sound track. A typical example of its use follows. Assume that dialogue has been recorded on a street with considerable traffic noise in the background. During rerecording (dubbing), the BNSA is set to close down a given amount during dialogue pauses; thus, the background noise is reduced. Precautions must be taken to keep the reduction in the background from becoming great enough to be noticeable, since too much reduction at a pause between sentences become quite annoying to the listener. Also, if the amplifier is set up too tight, it causes a thumping noise at the end of the spoken word, which also becomes annoying. A small amount of noise must be left in each pause to avoid listening fatigue. Such amplifiers are not used with music.

When employed properly, a BNSA may be used to salvage sound tracks previously considered unusable. Because such amplifiers are provided with an adjustable release time, they may be used to reduce reverberation to an ex-

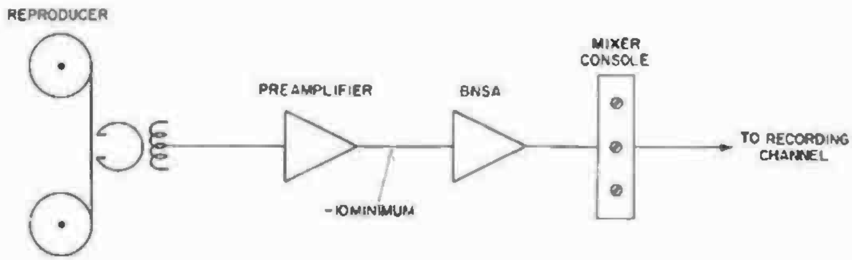


Fig. 18-104A. Position of a background-noise suppression amplifier (BNSA) in a rerecording channel.

tent. It should be understood that this amplifier does not actually remove the noise recorded in the dialogue, but because the listener is concentrating on the dialogue, a reduction of noise between words makes it appear that the background noise has been reduced. If the original sound track contains background noise that is equal to or greater in level than the dialogue, little can be done to improve the situation.

Devices of this type are not discriminatory to frequency and operate regardless if the background noise is of low or high frequency in nature. The frequency response is generally greater than that of the recording channel. The small amount of added distortion is of little consequence.

A block diagram showing the position of a BNSA in a rerecording channel is shown in Fig. 18-104A. Background suppression amplifiers as a rule have sufficient gain to be placed in a fairly low-level position; however, this is not always true. Therefore, to attain the most efficient operation of the device, the manufacturer's data sheet should be consulted.

The term "background noise suppression amplifier" (BNSA) should not be confused with the background amplifier (up and downer). This was an amplifier used in the early days of rerecording (1931) to reduce the music and sound effects behind the dialogue when the actor spoke, without disturbing the level of the dialogue. The control or side amplifier had frequency characteristics that would allow it to respond to frequencies only within the dialogue range. The first amplifiers of this kind used voice-operated relays; later, electronic control was added. For operational purposes, the sound effects and music had to be on separate sound tracks, with a separate background amplifier in each track. The use of such a device was annoying, as the sound effects and music levels were continually being increased or decreased, which detracted from the dialogue.

Modern background-noise suppression amplifiers are generally constructed to use five stages of noise reduction of 5 dB each, that may be inserted singly or together, reducing the noise by an equivalent amount. During

Fig. 18-104B. Magna-Tech Model 348, solid-state background-noise suppression amplifier.

periods of no modulation, the attenuation of these stages is reduced rapidly to zero, to permit the program signal to be passed without attenuation. The noise thresholds of the various stages are staggered to remove the background noise gradually.

During its operation, by inserting the first stage the signal-to-noise ratio is increased by 5 dB. Inserting the second stage increases the ratio by another 5 dB, and so on for the additional stages. In this manner, a total of 25-dB background reduction can be attained. The attack time is on the order of 1 millisecond, and the release time is adjustable in three steps—fast, medium, and slow. By selecting the proper release time, sibilant distortion and reverberation in sound tracks can be considerably reduced. As mentioned previously, in the initial adjustment for a given situation, care must be taken to avoid clipping the end of the words, or using too much background reduction.

18.105 What is an electronic mixer? —It is another name for a compressor-limiter amplifier.

18.106 What is a dynamic noise suppressor?—A method of suppressing noise in recording circuits by controlling the bandwidth of the reproducing circuits. The control circuits make use of a reactance tube, electronic gating, and filter circuits. The bandwidths of the reproducing circuits are automatically controlled by the amount and character of the noise in the recorded material.

18.107 What is comprex recording? —A system of recording in which the original sound track is compressed a given amount reducing the dynamic range. When reproduced, the sound track is expanded the exact amount it was compressed. Because of the difficulty of controlling both the recording and reproducing characteristics, it has not been accepted commercially.

18.108 What is overmodulation?—Overloading or overshooting the maximum amplitude for a given recording system. Overmodulating increases harmonic and intermodulation distortion.

18.109 What does the term "percentage modulation" mean?—It is the percentage of the applied signal with reference to the maximum signal that may be applied to a recording system. One hundred-percent modulation of a recording system is the maximum signal amplitude that may be applied without overloading or with reference to a given amount of distortion. Fifty-percent modulation of a recording system is an input signal that is one-half the maximum signal amplitude, or 6 dB down from 100-percent modulation. In Fig. 18-24A is shown percentage modulation of a light modulator for optical film recording with reference to the maximum signal amplitude or 100-percent modulation.

18.110 How is percent modulation of a recording system calculated?—Percent modulation may be calculated:

$$\text{Percent modulation} = \frac{A}{B} \times 100$$

where,

A is the signal amplitude,
B is the carrier amplitude (the clear area of the film base).

18.111 Define the term "rerecording."—Whenever a sound track is recorded a second time it is said to be rerecorded. In the motion picture industry, it is common practice to rerecord all sound tracks into one composite track called a master. During the rerecording operation, music and sound effects are added as required and the levels smoothed out.

18.112 Define the term "dubbing." —In the early days of sound motion pictures when only disc records were used, the process of rerecording disc records was termed duplicating. The area where this work was performed was called the duplicating room, or "dupe-room." In some studios, the term "doubling" was also used. This term then degenerated to dubbing and has remained so throughout the years. Therefore, dubbing has become the expression for rerecording and vice versa.

18.113 What is lip-sync?—A recording procedure for recording dialogue by actors while watching a silent motion picture. The words desired are spoken in synchronization with the lip motion of the characters in the picture and recorded. Sometimes, previously recorded dialogue is used for timing and inflection.

18.114 Describe the mechanographic system of recording.—This is a system where the sound track is engraved in an opaque film. The finished sound track is similar in appearance to a variable-area sound track and is reproduced using a photocell. This system was invented by James A. Miller of the Phillips Miller Co., of Holland.

18.115 What is a rerecording monitor low-pass filter?—A low-pass filter used in rerecording channel monitoring systems, having the frequeucy characteristics of the average motion picture theater. This filter was used when photographic film was the only medium for recording sound for theater use. The characteristics of this filter were based on hundreds of acoustical measurements in theaters conducted throughout the United States. A composite was then made of the measurements and the filter designed accordingly. The rerecording mixer was then in a position to better judge the final recording. The frequency response is that shown in curve (a) Fig. 18-115. Since magnetic sound tracks are used extensively in theaters, a rerecording filter is seldom used.

In the rerecording of 16- and 35-mm optical sound track, it may be desirable to include in the rerecording monitor system a low-pass filter, having the characteristic given in curves (b) or (c). In some installations, only a 10,000-Hz low-pass filter is used and is left in the monitoring circuit permanently.

18.116 What is a monitor card?—A scale on a film recorder for indicating the percentage modulation of the light modulator to the recordist. Calibrations are also included to aid in the setting of the noise-reduction bias current and margin.

18.117 What is the purpose of a 45-Hz high-pass filter in a rerecording channel?—To limit the response of the channel at the low frequencies. The original optical film rerecording channels contained such a filter to eliminate the effect of low-frequency rumble in reproducers, which geuerally occured below 45 Hz. Unless very wide-range recording is being done, many recording activities still use a 45-Hz high-pass filter to remove the possibility of low-frequency noise. High-pass filters are also used in transfer channels in combination with optical recording equipment. (See Question 18.331.)

18.118 What are the bandwidths used for production recording and rerecording?—For production recording involving only dialogue and using ¼-inch magnetic tape, the bandwidth is 40 to 8000 Hz, with dialogue equalization being used below 800 Hz, and the mid-range high-frequency equalization above 1500 Hz. If the production channel uses 16-mm magnetic film, the high-frequency response will be about 7000 Hz, and for 17.5 mm, about 8000 Hz. The best dialogue recording is obtained when this frequency range is used.

For rerecording, the bandwidth is generally 40 to 8000 Hz. This is particularly true if the final product to be released uses an optical sound track. If the release print uses magnetic sound track, the music is sometimes extended to 10,000 Hz. However, this will depend entirely on the type of music, and whether the distortion is low in the

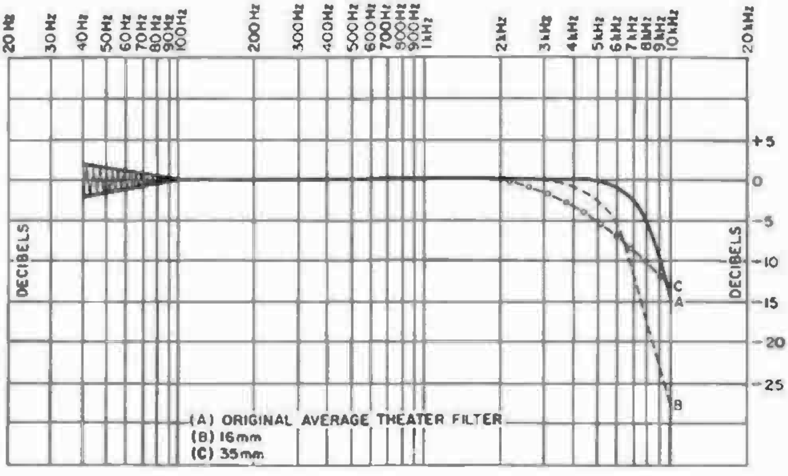


Fig. 18-115. Rerecording low-pass filters for 16- and 35-mm optical film recording.

original sound track. Generally speaking, for either type release print, a frequency range of 40 to 8000 Hz will be quite satisfactory. (See Question 6.122.)

18.119 Describe a photocell (PEC) monitor system used in photographic film recorders.—Photocell or PEC monitoring, as it is sometimes called, is a method used to monitor directly from the modulations of the light modulator in a photographic film recorder and is used by both RCA and Westrex. A portion of the modulated light beam is split off and passed through an optical system, amplified, and is available over the monitor system for comparison with the direct sound and that generated by the light modulator.

Early type film recorders monitored through the film base by placing the phototube behind the film. However, this method is now obsolete, as the film created a high-frequency hiss in its passage in front of the phototube, and the monitoring system required considerable high-frequency equalization to compensate for the loss of high frequencies.

A modern PEC monitoring system used by Westrex is illustrated in Fig. 18-26B. The method used by RCA is quite similar.

18.120 What is 35/32-mm optical film recording stock?—It is a standard 35-mm film base with 16-mm perforations, as shown in Fig. 18-120A, and is used in 35-mm film recorders that have been modified to run at a speed of 36 fpm.

Two methods of recording the sound tracks are used. In the first method, the

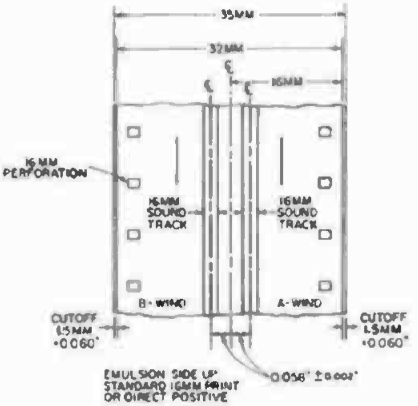


Fig. 18-120A. A 35/32 film base with two 16-mm sound tracks.

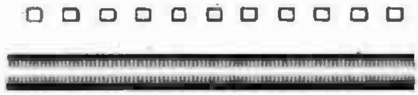


Fig. 18-120B. A 35/32 film base with a single 16-mm sound track.

sound track is recorded near the center at a distance equivalent to sound track placement for 16-mm film. After recording the full length of the roll, without rewinding, the film is rethreaded on the recorder and a second track is recorded.

After the film has been processed, it is split down the center and 1.5-mm of film is removed from the two outer edges. This results in two 16-mm sound tracks from a single base. Hence, the name 35/32-mm film.

In a second method, a single sound track is recorded in the exact center of the film (Fig. 18-120B). After processing, the film is not split. Both these methods result in considerable saving of storage space, as the storage of film becomes quite a problem in a large studio where several million feet must be stored every year.

It is the policy of most studios to modify their older recorders for this type recording. Because of the demand for this type recording, film manufacturers now supply a 32-mm base film with 16-mm perforations.

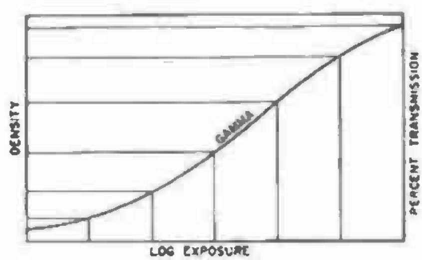


Fig. 18-121. A typical H and D curve.

18.121 What is an H and D curve?—A curve showing the relationship of exposure to density for a given film emulsion. This curve was originated by Hurter and Driffield, hence its name. It is also known as a D Log E curve A typical H and D curve is shown in Fig. 18-121.

18.122 What is a D Log E curve?—See Question 18.121.

18.123 Define the term "density"?—It is the common logarithm of opacity.

18.124 What is fog density?—An increase in density caused by development and is not due to exposure.

18.125 How is the correct density for sound track negative and print determined?—For variable-area recording, by the cross-modulation method. For variable-density, by the intermodulation method.

18.126 What is a cross-modulation test?—A test used with variable-area recording to determine the correct negative and print densities by measuring the fill-in of the valleys between modulations.

18.127 What is an intermodulation test?—A test used in variable-density recording for determining the correct negative and print densities. The intermodulation distortion is measured for a combination of negative and print density that will result in the lowest distortion. The combination which results in the lowest distortion is the optimum value of negative and print density.

18.128 How are the optimum negative and print densities determined, if test equipment is not available?—By making a series of exposure tests starting at a low-exposure lamp current setting and increasing the exposure current in fractions of an ampere in the following manner:

Assume the manufacturer's recommendation is that the exposure lamp be operated at 5.6 amperes. Start at 5.4 amperes and record sibilant words such as, "sister Susie is sewing shirts for soldiers," or "she sells sea shells by the seashore." Do this for every 0.20-ampere increase of the current until 6.6 amperes is reached.

Develop several samples of unexposed sound recording stock at developing times of 4, 5, 6, 7, 8, 9, and 10 minutes. This test is to determine how long the film may be developed without the fog exceeding 0.06. Next, develop the exposure tests to the maximum time determined by the fog test and then make prints for each sound track negative at the following densities: 1.20, 1.30, 1.40, 1.50, and 1.60. The print fog density must not exceed a density of 0.30.

Play back the prints, starting with the lightest density, on a reproducer having a known frequency response and pay particular attention to high-frequency distortion, sibilance, signal-to-noise ratio, and whether noise reduction breathing can be detected. Sibilance should not be confused with high-frequency distortion as there is a considerable difference between the two. Improper combinations of print and negative densities will cause high-frequency distortion, which is similar in sound to excessive sibilance.

The output level will vary with different combinations of negative and print density. Select a combination for the best reproduction of the high frequencies and overall level. Although the foregoing method is not entirely satisfactory, it can be used for either variable-density or variable-area sound tracks when test equipment is not available. However, whenever possible, the densities should be determined by the use of an intermodulation analyzer or cross-modulation oscillator for best results. It is the policy of most recording activities to hold the print density to a given value. If subsequent tests indicate a change in density because of a change in emulsions, the negative density is changed rather than the print density, as this method simplifies the making of future prints. Also, in this way the print density is always an optimum value.

18.129 Are tests required every time the emulsion is changed?—Yes. Every time a new emulsion number is used, tests must be made to determine the correct negative density, lamp current, and high-frequency loss. (See Question 18.160.)

18.130 What is image spread?—It is a term used to indicate the growth of the silver image beyond the outline of the optical image which originally formed it. A negative which has been properly exposed and processed will appear clean and sharp under a microscope. A good print will appear almost as sharp as the negative, with only slight deterioration.

If a variable-area negative is examined under a microscope, it will be noted the peaks of the modulations appear to be slightly rounded when compared to the valleys which appear to be sharp. This nonsymmetry is caused by the high density of the negative. Image spread is necessary in the negative to compensate for the image spread in the print which is in the opposite direction. For variable-area recording, the print density should be approximately 1.40 for satisfactory reproduction and good signal-to-noise ratio.

18.131 What is the principal cause of image spread?—Reflection and refraction in the film base between the halide grains and the surrounding gelatin causing the light to be scattered. Scattered light produces a latent image in the grains outside the exposure area. Therefore, as the density is increased, distortion of the modulation image increases. When developed at low density, peaks of modulations are reduced. For high densities, they are increased.

18.132 What is the effect on reproduction if the image spread is too great?—The reproduction appears to have excessive sibilance. However, this is actually not the case, but is high-frequency distortion.

18.133 How does the density of a print affect the electrical output from a photocell in a reproducer?—The output voltage from the photocell is directly proportional to the change in transmission between the light and dark areas of the sound track.

18.134 What routine tests should be made to maintain a uniform product?—Exposure tests to determine whether any change has taken place due to aging in the exposure lamp. This may be ascertained by running a series of wide-track exposures and reading the densities, then comparing them with previous tests against a standard developing time, or by means of a photometer mounted on the recorder.

18.135 What type film is used for recording sound?—Although the film used for sound recording is called sound-positive, when it is exposed and developed, the image is black. A negative sound track is shown at (a) in Fig. 18-288 and a positive one is shown at (b). (See Questions 18.142 and 18-151.)

18.136 What is a negative blow-up?—A 35-mm sound track made from a 16-mm sound track by optical enlargement. This practice is not recommended because of the increased background noise, image, spread, and distortion. However, reducing a 35-mm negative to 16 mm is common practice, as very satisfactory results are obtained.

18.137 What is the dynamic range of a 16-mm sound track?—Approximately 35 to 40 dB.

18.138 What is the dynamic range of a 35-mm sound track?—From 40 to 50 dB. 45 dB is about average when good quality control is maintained.

18.139 What is an "A" or "B" film winding?—A method of winding 16-mm optical or magnetic recording stock to bring the sprocket holes into the right position for a given optical or magnetic film recorder. When purchasing 16-mm film, the type winding must be specified. The USASI (ASA) Standard for 16-mm film winding is given in Fig. 18-139. It will be observed that the emulsion or magnetic coating is inside for either type winding.

18.140 Is the "A" or "B" winding used with 35-mm film?—No. Only with 16-mm film.

18.141 What is the color sensitivity of sound recording film?—It is blue-sensitive only.

18.142 Can the same recording stock be used for both variable-area and variable-density?—No. They require different emulsions and characteristics. For variable-area recording, the recording stock has fine grain and high contrast to permit the attaining of a sharp image over a wide range of frequencies. The control gamma generally falls between 3.0 and 3.20. Recording stock used for variable-density recording is also fine grain, but with lower contrast, with the control gamma falling between 0.40 and 0.60. Although the foregoing stock is referred to as sound-positive stock, the word positive is used only to indicate that the speed of the film is similar to release print stock and can be handled under the same safe-light conditions. Positive perforations are used in both types of film. (See Questions 18.135, 18.150, and 18.151.) Both emulsions are blue-sensitive only, and employ a gray antihilation base.

18.143 Show a flowchart for 16-mm black and white negative and sound track.—See chart in Fig. 18-143.

18.144 Show a flowchart for 16-mm color reversal original and sound track.—Such a chart is given in Fig. 18-144.

18.145 Show the method of producing 16-mm positive prints from black and white film, using 35/32 method, with sound track.—Such a method is given in Fig. 18-145.

18.146 Give a flowchart for 35-mm color negative and sound track.—Such a chart is pictured in Fig. 18-146.

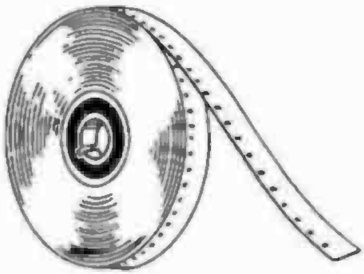
American Standard

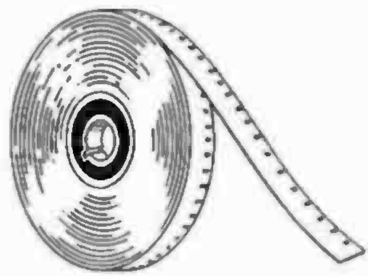
A and B Windings of 16mm Film, Perforated One Edge

ASA
Reg. U.S. Pat. Off.

PH22.75-1953

*UDC 778.5





Winding A
Emulsion side in

Winding B
Emulsion side in

(With the types of winding described below, the emulsion side of the film shall face the center of the roll.)

1. Scope

1.1 The purpose of this standard is to insure a uniform method of designating the type of winding (the location of the perforated edge) when ordering or describing 16mm raw-stock film with the perforations along one edge.

2. Film on Cores for Darkroom Loading

2.1 When a roll of 16mm raw stock, perforated along one edge, is held so that the outside end of the film leaves the roll at the top and toward the right, winding A shall have the perforations along the edge of the film toward the observer, and winding B shall have the perforations along the edge away from the observer. No preference for either type of winding is implied, since both types are required for use on existing equipment.

3. Film on Spools for Daylight Loading

3.1 When the film is wound on a spool with a square hole in one flange and a round hole in the other flange, it shall be specified as winding B when wound as described for B above and with the square hole on the side away from the observer. Windings other than winding B, on spools, are considered as special-order products.

Appendix

(This Appendix is not a part of American Standard A and B Windings of 16mm Film, Perforated One Edge, PH22.75-1953.)

A1. The types of winding covered by this standard are limited to those which are in general use.

A2. It is recognized that film on spools, with a square hole in one flange and a round hole in the other, can be wound in other ways than that described as winding B, and that for special purposes these windings may be supplied commercially.

Approved December 17, 1953, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers

*Universal Decimal Classification

Printed in U.S.A.
[illegible]

Price, [illegible] Cents

Fig. 18-139. USASI (ASA) Standard for A and B wind 16-mm film. This Standard was originated in 1953 and reaffirmed in 1961.

18.147 What is raw stock?—Unexposed film.

18.148 What is safety base film?—A film base made from a slow burning material called acetate. This base is used in the manufacture of all present-day 35-mm motion picture film. Safety base film may be identified by small marks placed at a distance of one frame along the edge of the sprocket holes and the words safety film. The marks are *parallel* to the film length and distin-

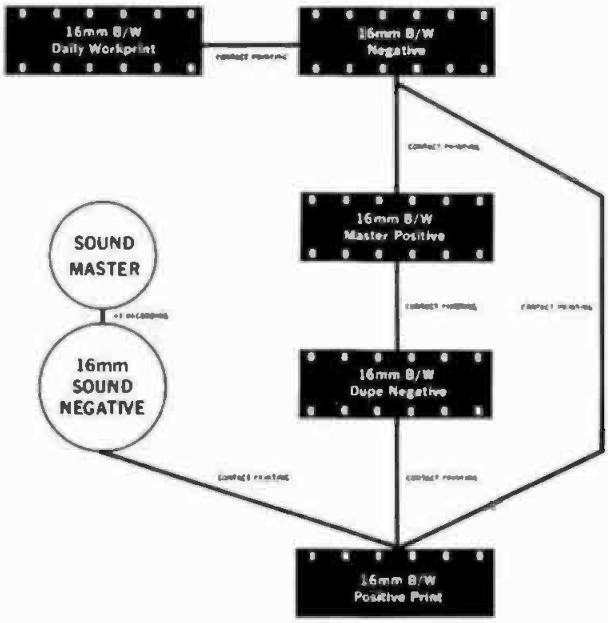


Fig. 18-143. Flowchart for 16-mm black and white negative and sound track. (Courtesy, General Film Laboratories, Division of DeLuxe Laboratories, Inc.)

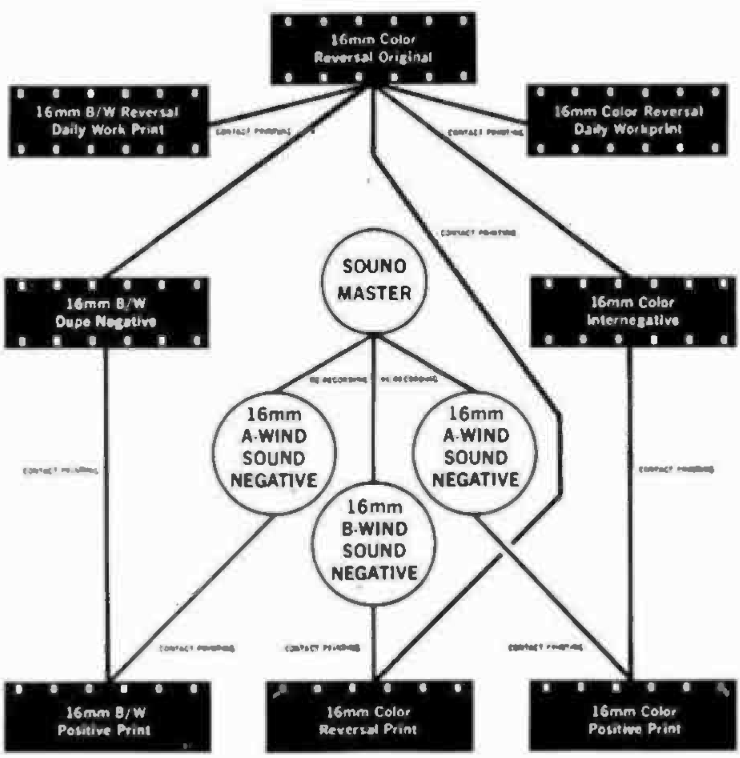


Fig. 18-144. 16-mm color-reversal original and sound track. (Courtesy, General Film Laboratories, Division of DeLuxe Laboratories Inc.)

guish safety base film from the older nitrate base type. The marks may be seen in Fig. 18-148.

18.149 What is a nitrate base film?—A highly inflammable material formerly used for the making of motion picture film base before the introduction of safety base film. Nitrate base film may be identified by small marks in the sprocket hole area placed at *right angles* to the film length. Nitrate film is no longer manufactured in the United States because of its inflammatory nature. However, many valuable nitrate base films are still in archival vaults throughout the world; therefore, if there is contact with nitrate film, it *must be handled with extreme care.*

18.150 What is fine-grain sound recording stock?—A fine-grain emulsion with an *antihalation* base with a grain size much finer than emulsions used prior to 1939. Although fine-grain film does not require a blue or ultraviolet filter, a somewhat better image is obtained with its use. Such emulsions may be loaded, using a red light in the dark room, inasmuch as they are insensitive to red light. The antihalation backing is a gray dye which serves to prevent halation. If the dye were not present, light penetrating the emulsion would be reflected at an angle from the back of the base to strike the emulsion again and cause halation around the image of the sound track. By using an antihalation backing, the light not absorbed by the emulsion undergoes absorptive action of the dye twice in order to return to the emulsion. Under these circumstances, the chances of halation occuring are practically nonexistent. The dye does not bleach out when the film is processed and it has no effect on the printing, except that the lamp current must be increased slightly.

18.151 Why is a positive stock used for sound recording?—The negative

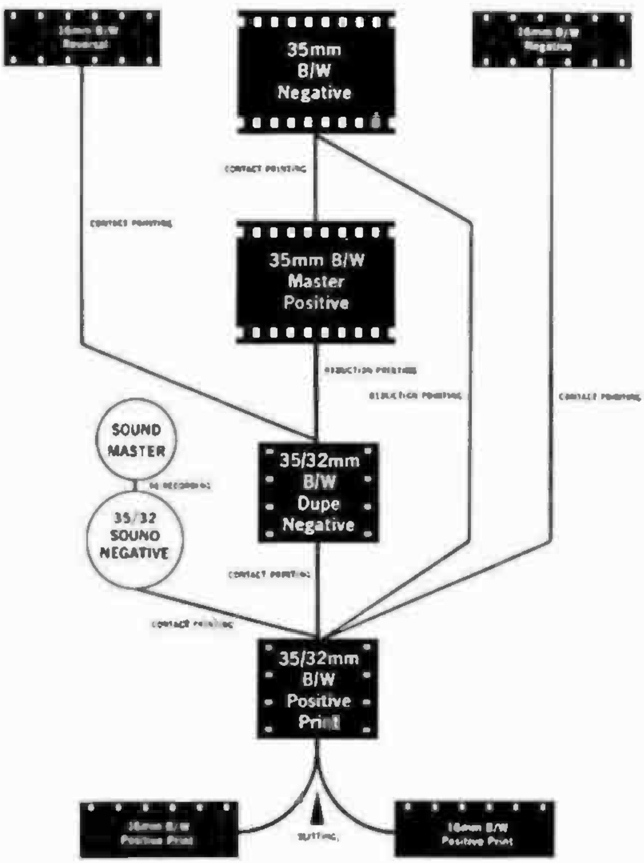


Fig. 18-145. The 35/32-mm method of producing 16-mm positive prints from black and white films. (Courtesy, General Film Laboratories, Division of DeLuxe Laboratories, Inc.)

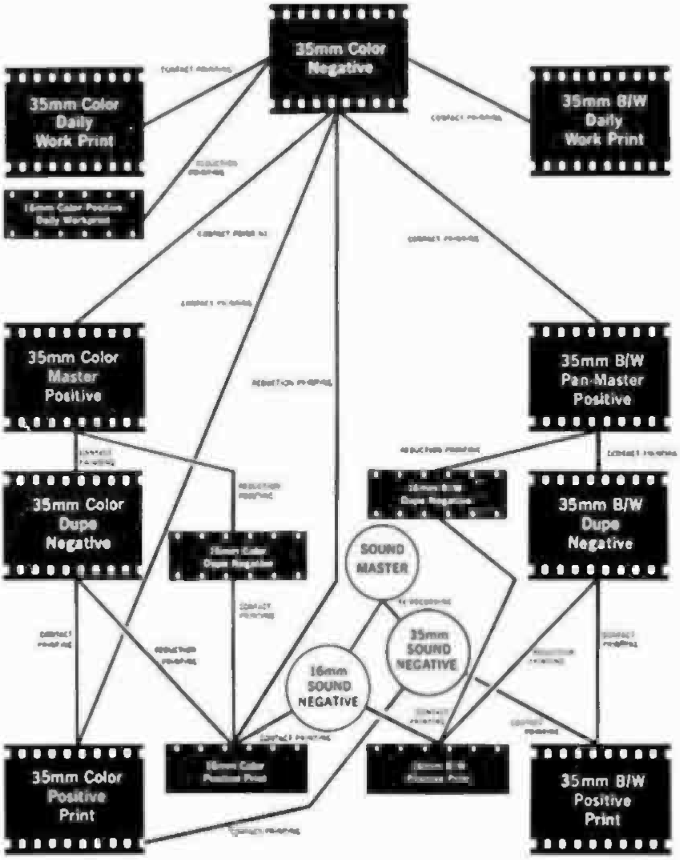


Fig. 18-146. 35-mm color negative flowchart with sound track. (Courtesy, General Laboratories, Division of DeLuxe Laboratories, Inc.)

stock used for the picture is not suitable for sound reproduction because of the higher background noise. The finer grain of a positive emulsion has a greater signal-to-noise ratio.

18.152 What is a reversal film?—One which, after exposure, is processed to produce a positive rather than a negative image. (See Question 18.287.)

18.153 Under what conditions should sound recording film be stored?—Negative reversal and sound film may be stored for a period of 6 months in a temperature not to exceed 55 degrees Fahrenheit, provided it is in the sealed cans as received from the manufacturer. For positive stock, the maximum storage temperature is 65 degrees Fahrenheit. In both instances, the relative humidity must not exceed 50 to 60 percent. Sound stock, when exposed, should be processed as soon as possible. Film may be subjected to temperatures from minus 65 degrees F to 160 degrees F. However, at the lower temperatures, the film speed is reduced and the film requires a much longer exposure time. Above 70 degrees sensitometric deterioration of the emulsion is much more rapid. At temperatures of 200 degrees, raw stock becomes badly fogged, with shrinkage and deterioration increasing rapidly.

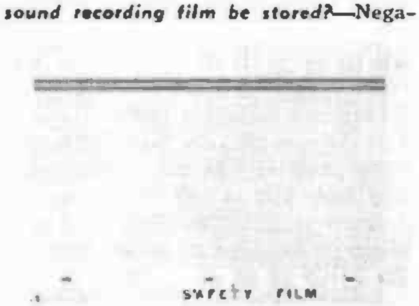


Fig. 18-148. Method of identifying 35-mm safety base film.

18.154 What is projection printing?—A method of reducing a 35-mm sound track or picture by optically projecting the image of the negative on the print stock. It is more commonly known as optical reduction printing. (See Question 18.157.)

18.155 Define the terms "inserted" and "wound-on" relative to motion pictrue film raw stock.—The expression "wound-on" is used to denote that the film is wound tightly on a core which cannot be removed from the roll, except by rewinding the film. Inserted means that the film initially was wound using a collapsible mandrel, and the core has been inserted in the roll. Thus, the film is not actually attached to the core. When received from the manufacturer, sound recording stock is wound-on. When purchasing 16-mm film, the type winding "A" or "B" must be specified. (See Question 18.139.) The cores for 35-mm stock are supplied with or without a keyway in the core, and they are generally referred to as male and female cores. Therefore, when ordering 35-mm stock, the type core must be stated.

18.156 Can a negative sound track be developed in a positive-type solution?—Yes. This is often done in small laboratories where it is impractical to have both negative and positive developing machines. The developing solution is slightly different from that used for normal positive development. Cross-modulation tests are made in the same manner and plotted as for normal sound track negative development.

18.157 What are the basic principles of an optical reduction sound track printer?—The basic components are shown in Fig. 18-157. The 35-mm sound track A to be optically reduced to 16-mm is placed at the left end of the machine. The image of the sound track is projected on a prism C through the optical system B by the exciter lamp D. The projected image of the sound track is turned 90 degrees by the prism both at the head end and at the point where it enters the 16-mm section of the printer. A second optical system E focuses the sound track on the 16-mm raw stock at F.

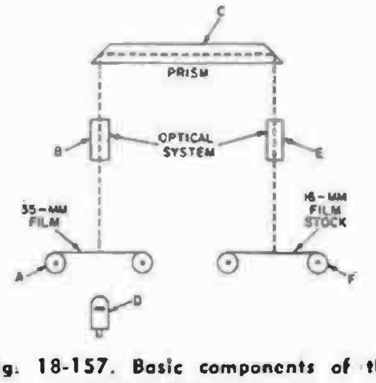


Fig. 18-157. Basic components of the optical reduction printer.

18.158 What are the principal objections to optical reduction sound tracks?—Possible increase in the background noise and distortion, as well as loss of high frequencies as a result of print slippage. Also, the printer may induce flutter in the final print. However, all of these defects may be controlled by close maintenance of the machine.

18.159 What is a contact sound-track printer?—A sound printer in which the negative sound track is held in close contact with the print stock on which the image is to be printed.

18.160 What method is used by the manufacturer of raw stock to indicate a new emulsion run?—The manufacturers of raw stock number their emulsions in different manners. Eastman Kodak uses three groups of figures. The first group designates the type emulsion, the second group the emulsion run, and the third group the emulsion cut. A typical number might be:

Type	*Run*	*Cut*
5375	101	21

If the second group of figures is different from those on the stock previously used for recording, a new set of cross-modulation or intermodulation tests is in order to determine the correct negative and print densities. Under certain instances different emulsion cuts will require a different exposure lamp current; therefore, a lamp test is required.

18.161 What are the laboratory processing tolerances for negative and print sound-track densities?—Most laboratories can hold the requested densities within plus or minus 0.05 points. Thus, if a negative density of 2.50 is requested, the density could vary from 2.45 to 2.55. For a print of 1.40, the density could vary from 1.35 to 1.45. Under no circumstances should the density be permitted to exceed 0.10 points from the optimum density.

18.162 Is it permissible to hold the print density at a given value and vary the negative density when changing

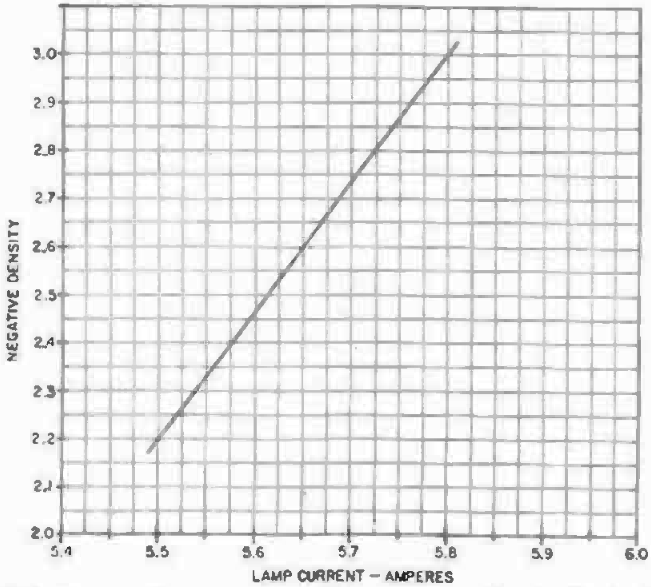


Fig. 18-163A. Exposure lamp current plotted versus negative density for 16-mm film.

emulsions?—Yes. The print density should be held at a given value and small variations, as required, made in the negative density. By this method prints may be ordered at any future time knowing that the print density will always be correct. If such a system is not adopted, it will be necessary to specify a different print density for each sound track. With different print densities, the output level will change and make it difficult to intercut sound tracks made at different times.

18.163 What is a photometer and how is it used with a film recorder?—It is a photocell of the self-generating type which is placed in such a position that a portion of the exposure light in the recording optical system falls on its light-sensitive target. Although the exposure lamp is initially set by its current, the final adjustment is made by the photometer. As the exposure lamp ages, the emitted light falls off and, for a given ampere setting, less light is obtained. If the reading of the photometer is noted when the exposure lamp is first calibrated, it may be used to set the

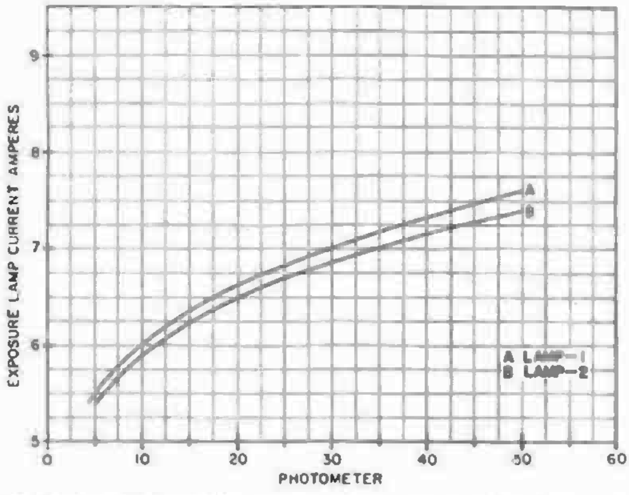


Fig. 18-163B. Typical photometer reading versus exposure lamp current for 35-mm film. Readings for two lamps are given to show the difference in exposure between the lamps.

lamp current for the same amount of illumination each time. Thus, the recordist is assured of the correct exposure at all times. It is also a good indicator as to when the exposure lamp should be changed. Recorders such as those in Fig. 18-26A, F, and G are equipped with photometers. A typical plot of exposure lamp versus negative density for 16-mm film is shown in Fig. 18-163A. The actual readings are shown by the dots above and below the line. For all practical purposes, the line can be assumed to be straight. Fig. 18-163B is a plot of photometer readings versus exposure lamp current for 35-mm film. Two lamps are plotted to illustrate the difference in exposure for a given lamp. This indicates the importance of a photometer.

18.164 What is a working print?—A print used by the editorial department for editing purposes. When completed, the negative is cut and a new print made for rerecording.

18.165 What does the term "opacity" mean?—It is the amount of opposition offered to light transmission by exposed and unexposed film. It is the reciprocal of transmission. A table of opacities is given in Section 25.

18.166 What is percent transmission of film?—The amount of light that will pass through an exposed and processed film. Percent transmission may be expressed:

$$\text{Transmission} = 100 \times \frac{L_t}{L_i}$$

where,

L_o is the total incident light,
L_t is the total light through the film.

Generally, the transmission of film is measured in terms of its opacity:

$$\text{Density} = \text{Log Opacity}$$
$$= \frac{1}{T}$$

where,

T is the transmission.

(See Fig. 25-160.)

18.167 What are dailies or rushes?—The first print made from the previous day's shooting. The picture and sound track are placed in sync by the cutting department and run for the director, camera, and sound departments to check picture and sound quality.

18.168 What is the aperture effect?—A term applied to the high-frequency harmonic distortion caused by the height of the recording and reproducer apertures. For recording, the slit height should be as small as practicable to reduce the effects of harmonic distortion. The slit height in a modern film recorder is approximately 0.25 mil. To reproduce 9000 Hz with minimum distortion, the reproduce slit should be 1.0 mil, in order to cancel second harmonic distortion.

Motion picture theater reproducing equipment as a rule employs slits of 1.2 to 1.3 mils in height. Such heights may be used because the frequency response is limited to approximately 8000 Hz.

18.169 What is film-loss equalization?—Film-loss equalization is equalization inserted in the recording circuit of a photographic film recorder to compensate for the high-frequency loss of the film. High-frequency loss is due to the characteristics of the film, linear speed, optics, printer loss, and processing losses. (See Question 6.127.)

18.170 How is film loss measured?—When a new recorder is installed or the optical system is overhauled, it is necessary to make film-loss measurements to establish the amount of film-loss equalization required to compensate for the characteristics of the film, linear speed, optical printer, and processing losses. All equalization and filters normally used in the recording channel (except in the instance of a light valve equalizer) are removed. Frequencies of interest, generally 400 Hz to 10,000 Hz, are applied to the input of the recording amplifier, and the light modulator is deflected by 80-percent modulation (2 dB below 100-percent modulation) as accurately as possible for each frequency.

After the negative has been recorded and processed, the peak amplitudes of each frequency are measured with a calibrated microscope or optical comparitor. The results are plotted, frequency versus amplitude, as shown in Fig. 18-170A. The amplitudes of the print are measured in a similar manner. Before measuring the print, a frequency-response measurement of the reproducer must be made, using a multifrequency test film. From its frequency-response test, necessary corrections are determined and applied to the print measurement to compensate for

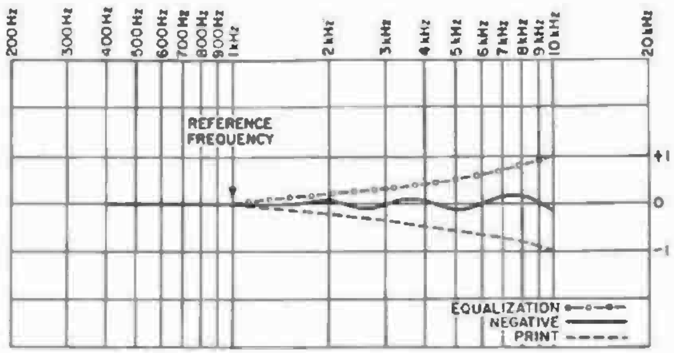


Fig. 18-170A. Negative and print losses for 35-mm variable-area film.

the variations in the reproducer response.

If both the negative and print are measured using a comparitor, the difference between the two is the printer loss. The negative cannot be successfully measured in a reproducer.

If the print density is satisfactory, the required film-loss equalization will be an inverse characteristic of the film loss. The measurements given in Figs. 18-170A and B are for galvanometer recording. Typical film-loss characteristics using a light valve for both variable-area and variable-density film are given in Figs. 18-170C and D.

To make the initial recording for variable-density film, an exposure is made for a negative density of about 0.48 and a print of 0.6. For variable-area film, a negative density of about 2.40 and a print density of 1.40 will be satisfactory. After the film equalization has been established, cross-modulation or intermodulation measurements are then made, as discussed in Questions 18.232, and 18.274 through 18.277. Film-loss equalization requires no further attention when once established, unless there is some radical change in the method of processing, printing, or other devices directly affecting the recording on the negative. Film-loss equalizers are discussed in Question 6.127.

The variations in the negative response of Fig. 18-170A are caused by the slight variations in reading the optical comparitor. It will be observed that the graduations for this particular graph are 0.2 dB per graduation.

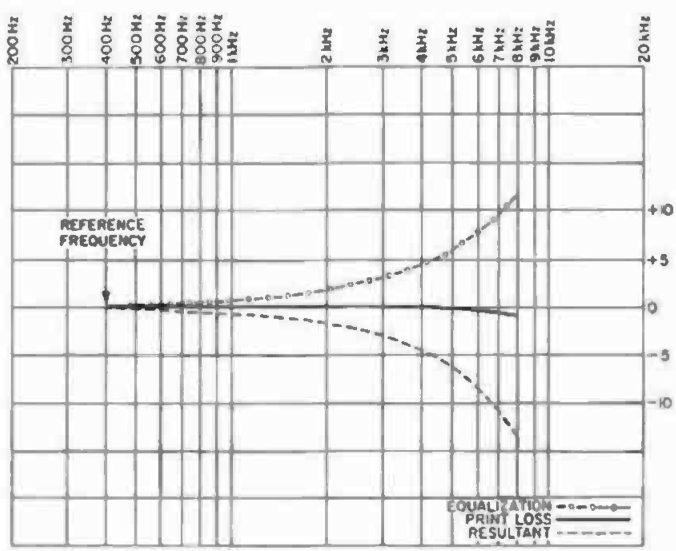


Fig. 18-170B. Print losses for 16-mm film, variable-area recording using a galvanometer. The amplitudes of the recorded frequencies were measured using an optical comparitor.

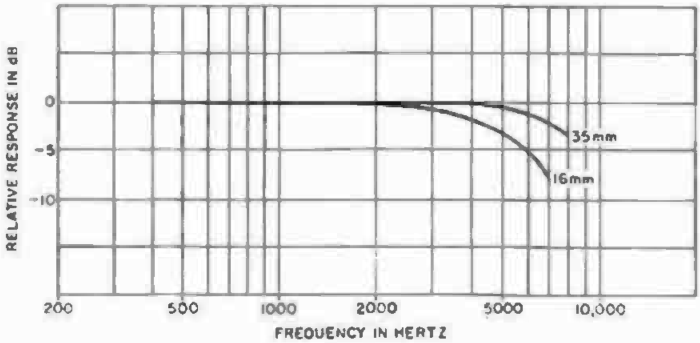


Fig. 18-170C. Typical recorded variable-area frequency characteristics using a light valve. Response is measured electrically.

18.171 How is a sound head equalized for a uniform frequency response?—By playing back a multifrequency film and adjusting the sound head optical system for maximum output at 9000 Hz. The equalization in the photocell preamplifier is then adjusted for a uniform output frequency characteristic.

18.172 What is the loss per generation of rerecorded film?—For each generation, approximately 85 percent of the resolving power remains.

18.173 Show the characteristics of a light-valve equalizer.—Light valves have a rising frequency characteristic caused by the resonant frequency of the ribbons, starting at about 3000 Hz and continuing to a resonant peak of approximately 6 dB at 6000 Hz. To compensate for this rising characteristic, a light-valve equalizer is employed, with a frequency characteristic inverse to that of the light valve. In Fig. 18-175 is shown the frequency characteristic for such an equalizer, strapped for a 7-dB insertion loss. The amount of equalization depends on the type of light valve. (See Question 18.15, and Fig. 18-26E.)

18.174 Give a recording characteristic suitable for optically reducing 35-mm sound track to 16-mm sound track.—Such a characteristic is given in Fig. 18-175. It will be noted that an 80-Hz high-pass filter is used to limit the low-frequency end. A small amount of midrange high-frequency equalization is used to compensate for losses in this region and to add presence to the reproduction. The magnitude of the equalization will vary with the optical reduction machines and laboratory processing; therefore, the characteristic shown is only for guidance. Listening tests must be conducted and the characteristic altered to determine the final characteristic.

18.175 What are the techniques used for rerecording 16-mm sound track?

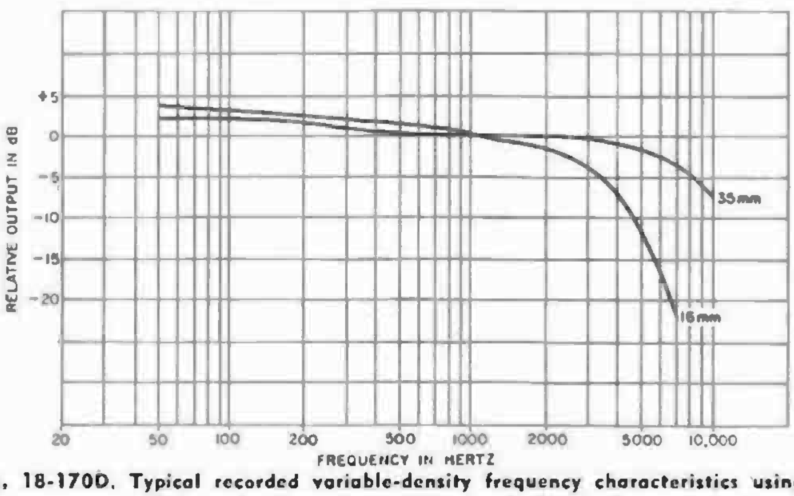


Fig. 18-170D. Typical recorded variable-density frequency characteristics using a light valve. Response is measured electrically.

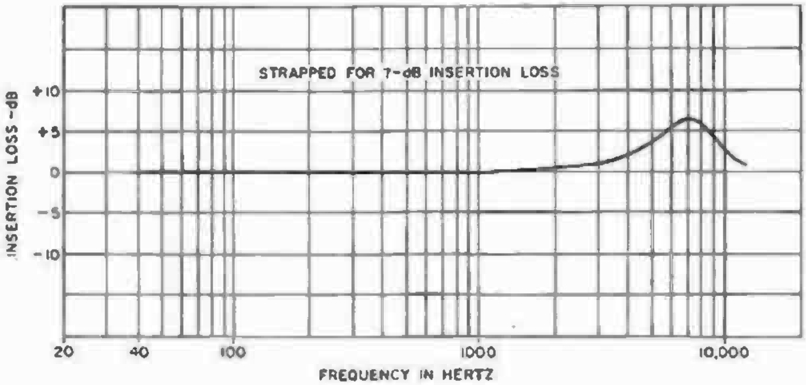


Fig. 18-173. Light-valve equalizer characteristic for Westrex light valves.

—Because the dynamic range of 16-mm film is considerably less than 35-mm film, it is necessary that 16-mm dialogue be recorded at approximately 75- to 80-percent modulation. To aid the mixer in obtaining a correct balance and to bring the lower levels up to an intelligible level (for television), the monitor level is lowered 4 dB below that normally used for 35-mm rerecording. The frequency response of the monitor should be cut off at about 6000 Hz. For dialogue, an 80-Hz high-pass filter is employed to limit the low-frequency end. A suggested recording characteristic is given in Fig. 18-175.

18.176 Describe the characteristics of a direct-positive recording.—Direct-positive recordings are made using a variable-area sound track, recorded with either a light valve or galvanometer. The track is developed to a gamma of 2.5 to 3.0, and exposed for a density as determined by cross-modulation or sibilant tests. In some instances, a special device called a cross-modulation compensator is used in the recording circuit to permit the making of a high-density print (1.5 to 1.9), with lower cross-modulation products. This is accomplished by predistorting the recorded signal in order to cancel the cross-modulation components produced by image spread due to the high density. A typical frequency response from a 16-mm direct-positive recording is shown in Fig. 18-176. The cross-modulation compensator is a development of RCA, and can only be used with a biased galvanometer. (See Questions 18.201 and 18.246.)

18.177 Describe the various optical test films available for photographic sound equipment.—Test films for optical devices, like those for magnetic equipment, are many and varied. Among these are: multifrequency test films, Academy test film (with picture) consisting of products from the majority of the major studios, focusing azimuth, flutter, and many others. Also available are special test films for television

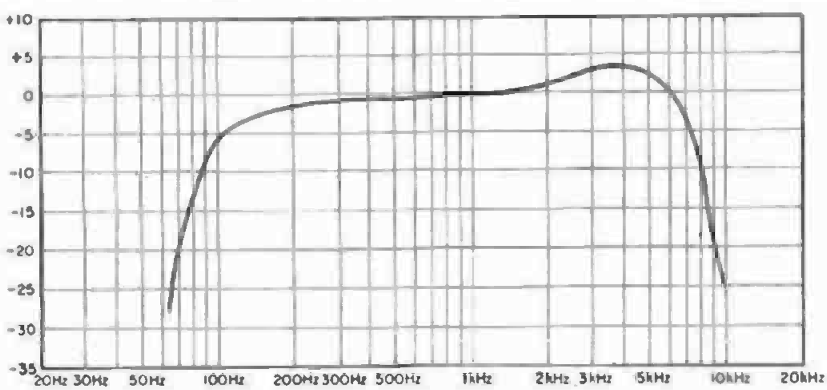


Fig. 18-175. Suggested recording characteristic for rerecording 35-mm sound tracks to 16-mm photographic film.

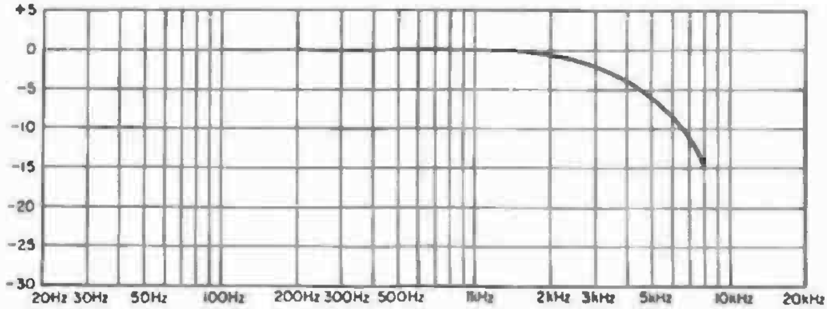


Fig. 18-176. Recording frequency characteristic for 16-mm variable-area, direct-positive recording.

equipment. These films may be obtained from the Society of Motion Picture and and Television Engineers (SMPTE).

18.178 What does the term "gamma" mean?—The gamma of a photographic material is the slope of the straight-line portion of an H and D curve, as shown in Fig. 18-121. Gamma represents the rate of change of the photographic density with the logarithm of exposure, and does not by itself define the contrast of the negative or print, but is only one of the many contributing factors in the processing chain of events. Both gamma and density are specifications used in the development of photographic sound track.

18.179 What does the term "print-up" or "print-down mean?—It is associated with a method used in printing variable-density souud tracks to increase or decrease the output level by changing the average transmission. The purpose of the change from normal is to increase the output level for main or end title music and to obtain the greatest dramatic effect in certain portions of the film. By this method, an increase of 4 to 6 dB may be gained in the release print. The output increases when the sound track is printed lighter than normal, thus increasing the dynamic range.

18.180 What is a film buckle?—It is a condition which occurs when the film in a camera or sound recorder jumps the sprockets and piles up in the transport system. As a rule, a switch is provided to protect the equipment when this occurs.

18.181 What is a scratch track?—A temporary sound track used for cuing purposes only. It is generally made by a person other than the one who will do the final recording. One-quarter inch-tape machines are often used for this purpose.

18.182 What is a blimp?—A sound deadening cover placed over a motion picture camera on a motion picture set to prevent the noise of the camera movement from being picked up by the microphone. It is also called a barney (horse blanket).

18.183 What is grandeur film?—A 70-mm film introduced by 20th Century Fox several years ago for wide screen projection.

18.184 What is a loading bag?—A cloth bag with arm sleeves for loading film in daylight.

18.185 What is papering a sound track?—Dropping small pieces of paper in the reel while it is running to mark a cue or rewind point.

18.186 What is a montage?—A group of dissolves and superimpositions. A term used by the special-effects department of a motion picture studio to describe a jumble of pictures or sounds.

18.187 What are edge numbers?—A series of numbers with key letters appearing along the edge of negative film. Their purpose is to identify the footage of sound track or picture. Such numbers may be seen on the film outside the sprocket holes in Fig. 18-191.

18.188 What is the purpose of waxing film?—To prevent the film from sticking to the shoes of the projector aperture plate. The wax is applied to the edges of the film.

18.189 What is green film?—Film which has not been properly dried and waxed. Green film when run through a projection machine will stick to the aperture plates causing the emulsion to pile up and damage the sprocket holes. Green film also causes the picture to go in and out of focus. Sticking to the

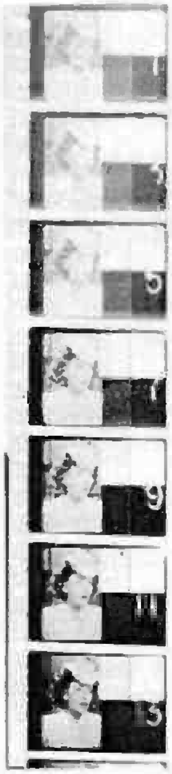


Fig. 18-190. A cinex strip.

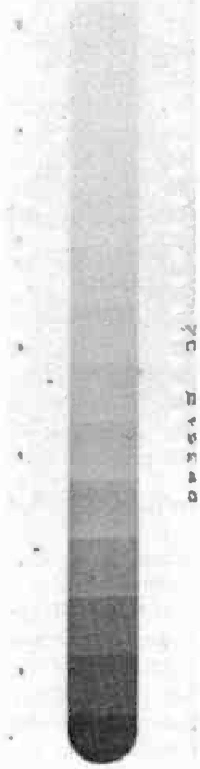


Fig. 18-191. A sensitometric strip.

aperture plate can be prevented by applying a mixture of beeswax and carbon tetrachloride, or special solutions developed for this purpose, to the film.

18.190 What is a cinex strip?—A film strip with a series of exposures corresponding to standard printer lights from 1 to 21. Each succeeding exposure is given a greater exposure than that of the previous one, as shown in Fig. 18-190.

18.191 What is a sensitometric test strip?—A strip of film containing a series of graded exposures as shown in Fig. 18-191. It is used for quality control in film-processing laboratories. It is also called a gamma strip.

18.192 What are printer light steps?—Varying degrees of printer light intensity for increasing or decreasing the density of exposure on a printing machine. The film illustrated in Fig. 18-190 is a typical example of the different light steps of a printer. Printers for sound-track printing include half-step lights.

18.193 What is a gamma strip?—A film strip exposed in a sensitometric printer and used in a motion picture processing laboratory for the control of gamma. A typical gamma strip appears in Fig. 18-191.

18.194 What is a control strip?—A gamma strip as described in Question 18.191.

18.195 What is a loading hook?—A metal rod bent in the form of a square hook and used for releasing the light trap in a film magazine when loading the magazine.

18.196 Describe the construction of a densitometer.—A densitometer is basically a sensitive light meter for measuring the density of exposure of motion picture film, in either the picture or sound track areas, for both black and white or color.

The Westrex Model RA-1100H, pictured in Fig. 18-196A is a direct-reading integrating-sphere type electrical densitometer using two photocells for measuring the ratio between the light transmission with and without the film in the beam of light. Light from an exciter lamp, held constant by a voltage regulator, is interrupted by a light chopper (rotating disc) and directed through the film and a limiting aperture by suitable lenses. After passing through the film, the light enters an integrating sphere where it produces a signal in the photocell, proportional to the amount of light transmitted. The amplified and rectified output of the photocell is indicated on a meter calibrated to read directly in terms of density. A schematic diagram of the integrating sphere and photocells is shown in Fig. 18-196B.

This densitometer is designed to provide three types of integral diffuse-density measurements. The integrating sphere essentially collects all the light passing through the film and aperture. A type 929 blue-sensitive and a type 925 red-sensitive photocell are mounted in the sphere in a manner in which they receive only the light that has been reflected at least once from the surface of the sphere. Three light filters mounted on a filter disc provide transmission of limited spectral ranges for the type of measurement desired. Setting the filter disc in place, with the desired filter in the optical path, automatically operates switches. In turn, these switches control a relay that connects the appropriate photocell to the input circuit of the pre-

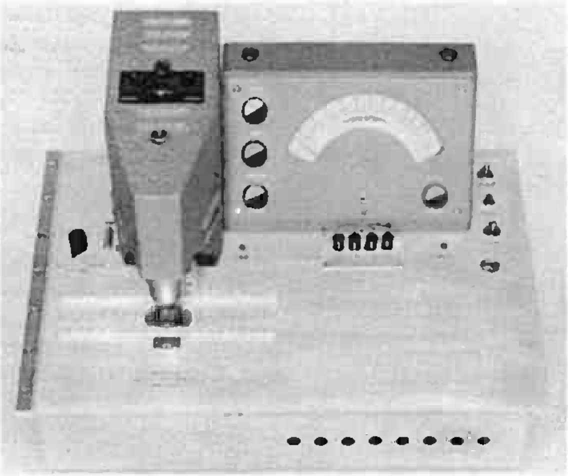


Fig. 18-196A. The Westrex Corp. Model RA-1100H densitometer.

amplifier. The resulting reading of density, as defined by Hurter and Driffield, is based on the differential with and without film sample in the light beam path.

Three filters are mounted in the head, with the chopper wheel. Filter "V" is used for measuring the visual density of black and white film. It has been selected so that the resultant spectral response closely simulates that of the human eye.

A Filter "P" is used for measuring the printing density of black and white film. It provides a density reading which is essentially the same as is seen by the positive film.

A Filter "IR" is an infrared filter and is used in the measurement of silver or silver sulfide with dye sound tracks. This filter in conjunction with the photocell characteristics provides a narrow bandpass in the spectral region of 800 millimicrons. Its use results in a thoroughly reproducible density reading of tracks on color films measured at the peak wavelength of the spectral sensitivity of the average theater reproducer.

Densities from 0 to 4.0 may be read with the visual or printing filter in use.

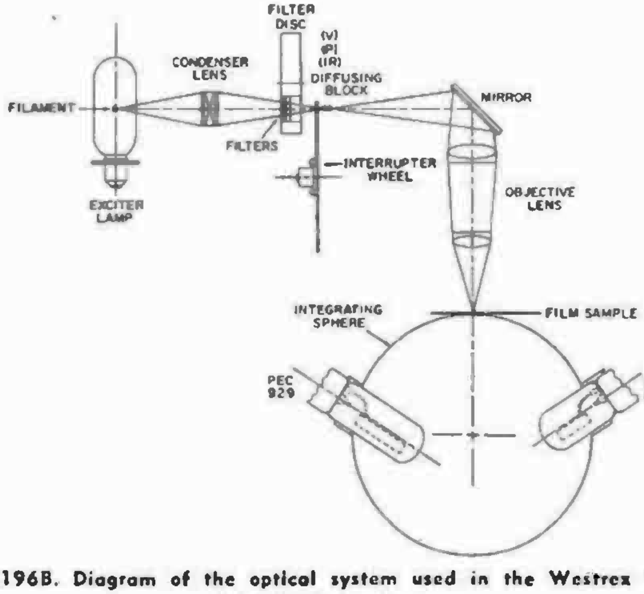


Fig. 18-196B. Diagram of the optical system used in the Westrex Corp. Model RA-1100H densitometer.

The infrared filter passes only a very narrow band at 800 millimicrons and consequently reduces the sensitivity of the instrument by approximately 20 dB. This limits the readings with this filter to a density of 3.0.

The head assembly at the left of the instrument (Fig. 18-196A) contains the optical system and filters. Below the head assembly is an aperture plate in the form of a metal slide. Just above the aperture plate is a film gate which is designed to accommodate 35-mm film. The film gate may be locked in position by a locking lever located at the left of the head assembly. Immediately below the aperture plate is the integrating sphere.

To the right of the head assembly is a meter having a logarithmic response, calibrated in density. It, therefore, reads linearly in density. The scale is calibrated 0 to 1.1. Four density ranges: 0 to 1, 1 to 2, 2 to 3, and 3 to 4, are covered by the use of pushbnttons below the meter. An extra 0.1-density calibration point at the upper end of the scale provides overlap between ranges.

Fig. 18-196C. Macbeth Instrument Corp. Model TD-102 Quantolog densitometer.

Three potentiometers are provided on the left side of the meter case for presetting the meter zero point for the three light filters. Contacts, operated by the filter wheel (on top the optical housing) automatically select the appropriate preset meter zero point for each filter in the optical path. On the right side of the meter case is a potentiometer for calibrating the high end of the meter scale.

The light source is a straight filament lamp, operated at a relatively low temperature to insure long life. The current for this lamp is maintained constant over a wide voltage range, hy a voltage-regulator transformer. The condenser lens assembly consists of a pair of plano-convex lenses which image the lamp filament on a block of optical glass, the length of which is chosen to eliminate the coil pattern at its exit surface. The cone of light falling on this block is interrupted by a synchrononsly driven interrupter wheel (chopper) which gives a frequency of 375 or 450 Hz on a power supply of 50 to 60 Hz respectively. The light from the glass block is reflected downward by an aluminum-coated first surface mirror to the objective lens which brings the exit face of the glass block to a focus at the film plane. The filter wheel previously discussed is located between the condenser lens and the interrupter wheel. Pittsburg HA-2043 heat-absorbing filters are mounted in front of the visual and print light filters in the filter disc.

The aperture plate contains a wide and narrow aperture. The wide aperture clears the scanning beam, the narrow aperture limits the width of the light beam.

Pictured in Fig. 18-196C is a Macbeth Instrument Corp. Model TD-102 Quantalog densitometer. This instrument may be used for measuring both picture

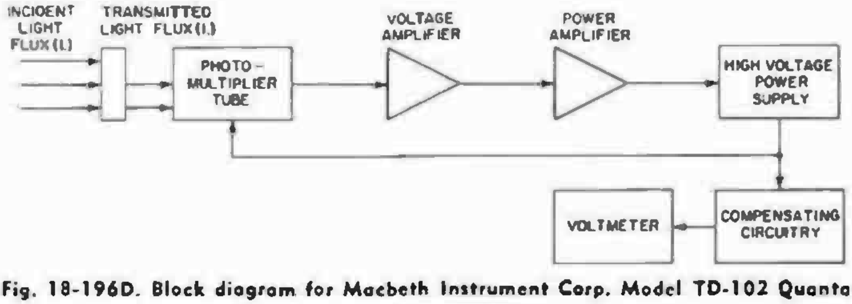


Fig. 18-196D. Block diagram for Macbeth Instrument Corp. Model TD-102 Quantalog densitometer.

and sound track. A block diagram of its basic components is shown in Fig. 18-196D, with the schematic diagram shown in Fig. 18-196E. The operation of the instrument is based upon the use of a photomultiplier tube in a feedback circuit, which automatically adjusts the dynode voltage applied to the photomultiplier tube, so as to maintain constant anode current. The dynode voltage required to maintain a constant current for any particular amount of

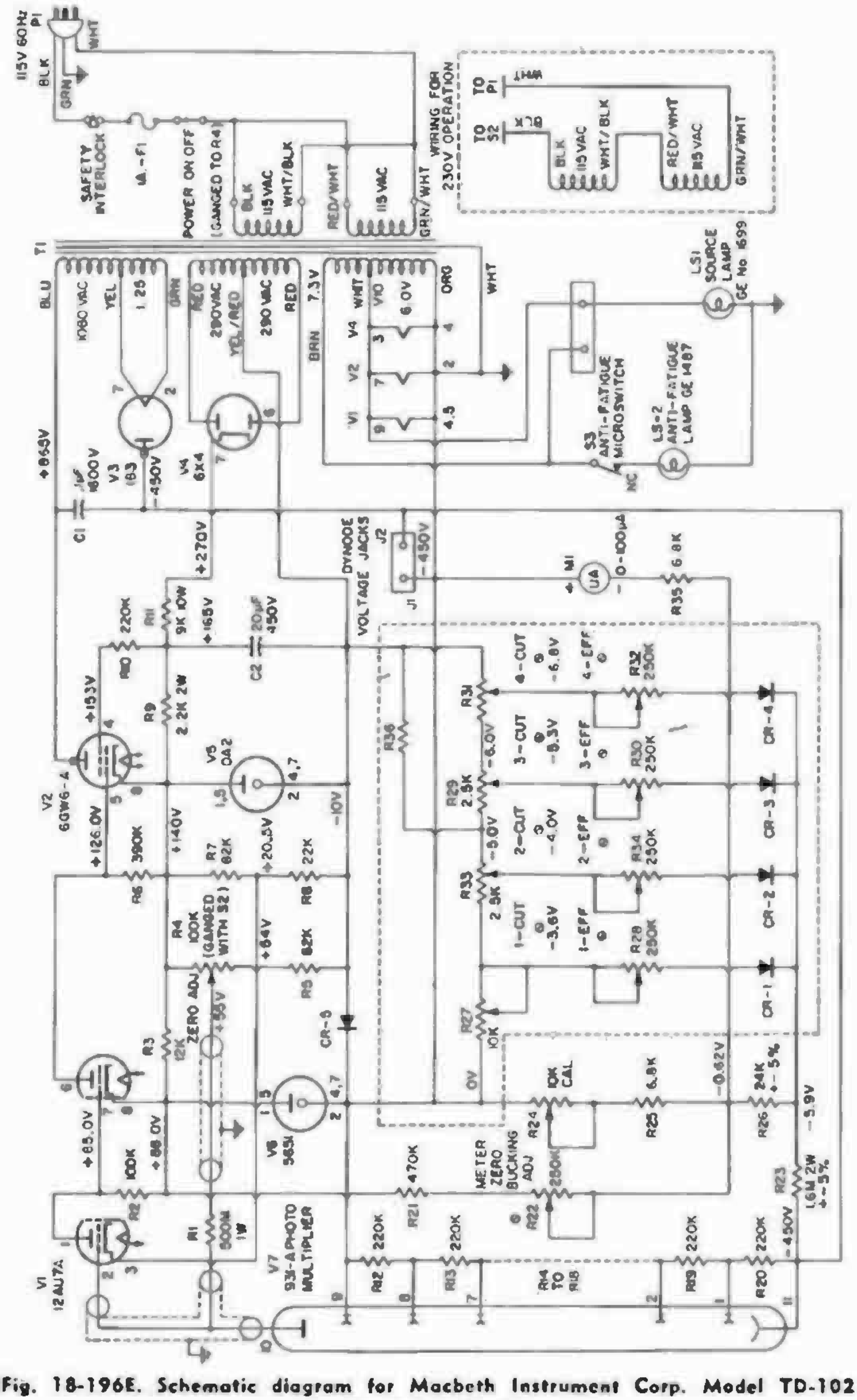


Fig. 18-196E. Schematic diagram for Macbeth Instrument Corp. Model TD-102 Quantalog densitometer.

light falling on its photosensitive cathode is used as a measure of the optical density of a sample placed between a light source and the photomultiplier tube itself.

If the light flux falling on the photosensitive surface of the photomultiplier decreases when a sample of film is introduced into the light path, the input signal of the voltage amplifier decreases. This signal is amplified and fed into the input of a power amplifier, which controls the output voltage of the high-voltage power supply. With a film sample in the optical path, the output of the high-voltage supply increases in order to maintain a constant anode current within the photomultiplier tube. The meter indicating the density is in reality a voltmeter, which constantly measures the voltage applied to the dynode element. Since the meter is calibrated in terms of density, the voltage increase indicates an increase in density. The dynode voltage for a typical photomultiplier tube will vary between 220 and 750 volts for samples whose optical densities fall between 0 and 4.0. As the relationship is not completely linear, a compensating circuit is used to calibrate the density readings to a standard. This instrument may be used for both color and black and white densities. (See Question 11.32.)

18.197 Can any type densitometer be used for the measurement of color film sound tracks?—No. A special densitometer must be used. (See Question 18.196.)

18.198 What is a Moviola?—A motion picture film editing and synchronizing machine manufactured by the Magnasync-Moviola Corp. The machine pictured in Fig. 18-198A is designed for editing 35-mm picture and sound track. Various combinations of 16-mm, 17.5-mm and 35-mm film heads may also be obtained using sound reproduction from either optical or magnetic sound track.

The sound track which is on a separate film is placed on the sound head section at the left and the picture film at the right. The picture and sound head sections are connected by a mechanical clutch, which may be disengaged to permit either head to be run independently of the other. Also, the machine may be run backward or forward at standard sound speed, or at a variable speed controlled by means of a foot pedal. The sound-pickup unit employs a photocell or a magnetic head, amplifier, and loudspeaker. Headphones may also be used when required.

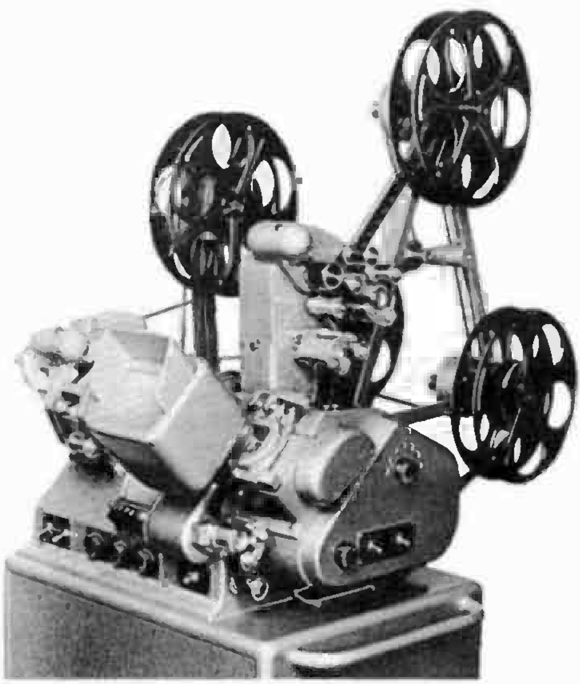


Fig. 18-198A. The Magnasync-Moviola Corp. Model UD-20-CS editing machine for synchronizing the sound track with the picture.

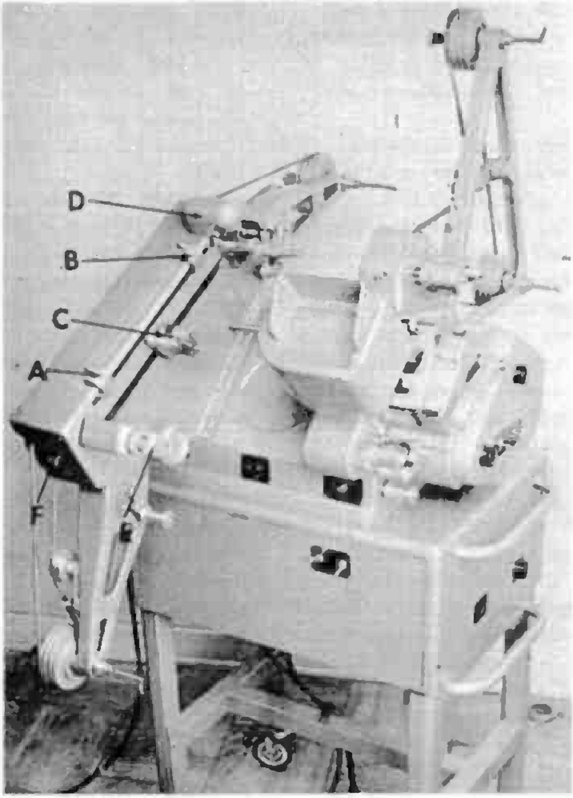


Fig. 18-198B. Magnasync-Moviola Corp. Model UD-20-S editing machine with the Carlos Revas magnetic search head at the left.

Another machine also manufactured by Magnasync-Moviola Corp. and equipped with a Carlos Revas magnetic searching head for synchronizing magnetic sound tracks to the picture is shown in Fig. 18-198B. The magnetic sound track is threaded in the head on the left and passes over two sprockets placed approximately 14 inches apart. Between the sprockets is a magnetic pickup head C mounted on a movable arm.

In normal operation, the magnetic head remains stationary and the film is passed over it for regular sound reproduction. In the event it is desired to locate a particular sound or modulation, the film is stopped with the section of interest between the sprockets. The magnetic pickup is then oscillated between the sprockets at sound speed by means of an electric motor. This enables the operator to hear the sound recorded on the magnetic sound track between the lower position indicator A and the upper position indicator B. An automatic switch is provided which cuts out the sound during the return trip of the magnetic pickup head. The start and ending of a particular sound or group of modulations may be accurately located by moving the position indicators A and B. Sound and picture action may be synchronized without stopping the machine by shifting the magnetic pickup head through use of hand wheel E.

The machine also contains a complete reproducing system for optical sound tracks; however, the optical reproducing unit D remains in a fixed position at all times. The switch for the motor driving the pickup head is shown at F. This machine may also be obtained for various combinations of film size and speed and for multiple sound track and picture reproduction.

An editing machine manufactured by the Westrex Corp. and designed for editing both optical and magnetic sound tracks with picture appears in Fig. 18-198C. Although the machine shown is for 35-mm film, it may be obtained for use with any size film. Normally, the picture is viewed on a small screen

Fig. 18-198C. Magnasync-Moviola combination 16/35-mm preview-type editing machine.

in the center of the machine; however, if desired, the picture may be viewed considerably enlarged on a screen placed at the rear of the machine. Facilities are provided so that the machine may be run either backward or forward at the sound speed or a variable speed. If the variable speed is used, the speed is controlled by means of a foot pedal. Footage counters and rewinds can be seen at the right- and left-hand sides.

One of the features of this machine is the elimination of the conventional intermittent sprocket employed in the picture head section. The picture is projected by means of a 12-sided prism which rotates at a continuous speed to provide continuous projection without the use of a shutter or intermittent mechanism.

18.199 What is the difference between a diagonal and a straight splicing machine?—As far as the machine itself is concerned, there is little difference, except for the shape of the cutting knives. A straight splice is one which is made at right angles to the motion of the film. A diagonal splice is made at an angle running from one corner of a frame to the other. This means the splice crosses the sound track at an angle.

Diagonal splices may be used with any standard type sound track, either variable-area or variable-density. However, they cannot be used with push-pull sound tracks, because part of the splice is out of phase with the sound track and will cause a noise when it passes before the photocell. Blooping is discussed in Questions 18.209, 18.210, and 18.211. A 35-mm automatic film-splicing machine manufactured by Bell and Howell is shown in Fig. 18-199A. A close-up of the splicing mechanism is shown in Fig. 18-199B, and typical splices appear in Fig. 18-199C.

18.200 What is sprocket-hole modulation and what is its cause?—During the development of the sound track, the sprocket holes, in passing through the developing solution, cause eddy currents to be set up in the solution. This action causes an increase in the development around the sprocket hole area

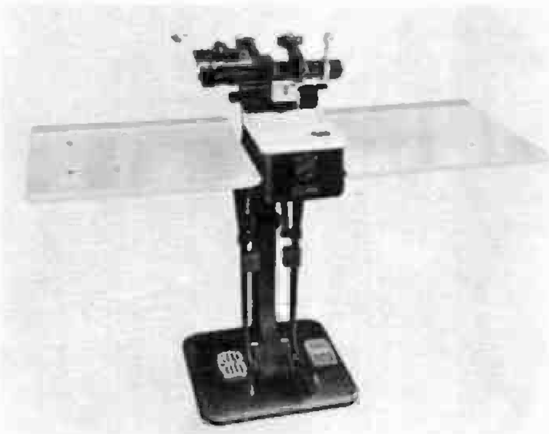


Fig. 18-199A. The Bell and Howell 35-mm automatic film-splicing machine.

compared to other portions of the film. When played back, a frequency of 96 Hz will be heard (35-mm film running at 90 feet per minute).

Sprocket-hole modulation may be prevented by turbulating the developer around the area of the sprocket holes to prevent the formation of eddy currents.

Sprocket-hole modulation may also be created by the misalignment of the film as it passes through the transport system of the projector.

18.201 What is meant by the term electronically printing a sound track?—The term is another name for direct-positive recording, and is a method of transferring a prerecorded sound track to a release print without intermediate steps. In the procedure used, the laboratory exposes the picture on the print stock but does not process it. A punch mark is put on the film where the sound is to start. The picture is then loaded in a sound magazine and placed on a recorder, with care taken to line up the punch mark in the film with the light beam of the recorder optical system. The sound is then recorded in the usual manner for direct-positive recording, as discussed in Question 18.176. Ektochrome reversal print stock may be used for this purpose.

As there are no intervening generations and the final sound track is a direct transfer from the original, considerable time is also saved. The principal drawback to this method of recording release prints is that each print must be recorded individually.

18.202 What is toe recording?—A method used in the early days of variable-density sound recording. The exposure was confined to the toe portion of the H and D curve for both negative

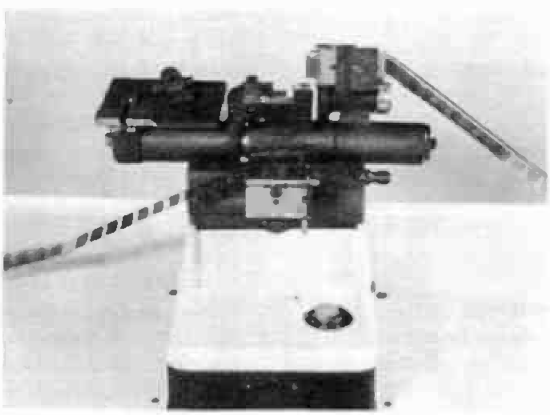


Fig. 18-199B. Close-up of the splicing mechanism of the Bell and Howell automatic film splicer.

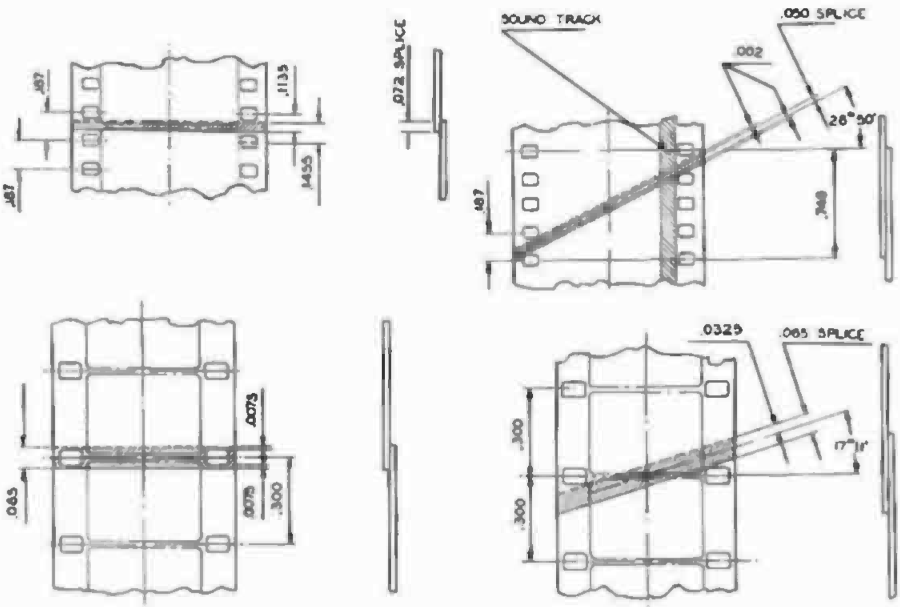


Fig. 18-199C. Optical film splices on 16-mm and 35-mm film, as made on a Bell and Howell automatic film-splicing machine.

and print. Such prints were used for playback purposes and tests. Their use was similar to the direct positive of today. The output level is about 6 dB greater than for a normal print. However, the signal-to-noise ratio is lower and the distortion is higher.

18.203 What is painting out a sound track?—The removal of unwanted noises of trailing sibilants by the use of an air brush and indelible ink. (This is not very satisfactory.)

18.204 What is a glow tube recording system?—An obsolete film recording system which used for its light modulator a vacuum tube containing two or more elements in a combination of rare gases. A high voltage is used to bias the elements on which the audio signal is superimposed. The audio signal causes the gas to ionize and vary its light intensity, thus producing a variable-density recording.

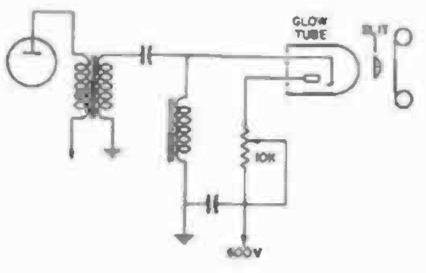


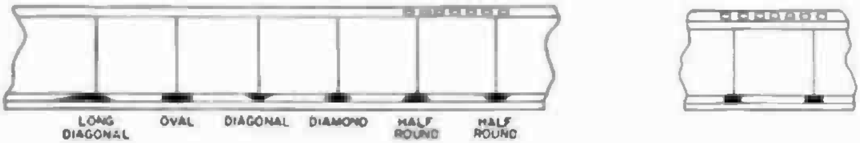
Fig. 18-204. The coupling circuit for the glow tube.

This method of recording is now obsolete because of its low light intensity, large wattage consumption (causing a considerable amount of heat), very short life, lack of uniformity, low sensitivity, low signal-to-noise ratio, and erratic electrical characteristics. The outline of this tube is shown in Fig. 18-4B and the method of coupling it to the output of an amplifier is shown in Fig. 18-204. About 100 volts of audio signal are required to properly modulate the tube.

18.205 What is a periscope?—A device used for checking the optical system of a film recorder. It is also known as a focusing microscope.

18.206 What is a silent sound track?—A sound track containing no modulation and used for intercutting between sections of dialogue to prevent having a completely dead period between the sections. For variable-density sound tracks, it consists of a light gray exposure the full width of the track. For variable-area sound tracks, it consists only of the bias lines.

18.207 What is a room noise track?—A sound track of the noise of a room or set. This track is made when everyone on the set is quiet while the ambient noise is recorded. This track is later



(a) *Variable-density and variable-area standard sound tracks.*

(b) *Variable-density and variable-area push-pull sound tracks.*

Fig. 18-209A. Bloop configurations.

used by the editorial department for intercutting between scenes for a uniform background noise. (See Question 18.206.)

18.208 What is a Kerr cell?—A light modulator containing two electrodes immersed in a solution of nitrobenzol. The exposure light is passed through a Nicol prism and focused on a position between the electrodes of the Kerr cell. The pale polarized light is rotated under the influence of a strong magnetic field external to the cell. This action results in a variable-density sound track. It was used in the early days of the development of film recording.

18.209 What is a bloop?—An opaque patch or painted image over a splice where two sound tracks are joined together (Fig. 18-209A). Its purpose is to prevent a sudden impact or click, caused by the splice, as it passes the photocell.

Sound-print bloops are painted over the splice with indelible ink, using an artist's air brush. Negative bloops are made by punching a hole in the negative, using an oval-shaped punch. When prints are made from the negative, the punch hole appears as an opaque oval-shaped patch over the splice. A flash bloop is a photographic bloop made during the making of prints by exposing the splice area to a strong light as it passes through the printing machine. The density variation is such that a smooth transition is obtained from one section of a sound track to another.

The blooping operation is automatically controlled by a notch punched in the edge of the sound negative which operates a microswitch turning on the blooping light. A typical flash bloop is shown in Fig. 18-209B.

18.210 How long should a bloop be for a standard sound track?—One wavelength longer than the lowest frequency to be recorded. If the lower limit of the system is 45 Hz, the bloop image must be 0.409 inches in length or one wavelength of 44 Hz. The wavelength for film running at a speed of 90 feet per minute or 18 inches per second may be calculated:

$$\lambda = \frac{18{,}000 \times 10^{-3}}{\text{frequency}}$$

18.211 Is it necessary to bloop a push-pull sound track?—Generally, no. In the push-pull system of recording, the two halves of the sound track are recorded 180 degrees out of phase with respect to each other. Therefore, their outputs as seen by a push-pull photocell are additive. When a splice appears

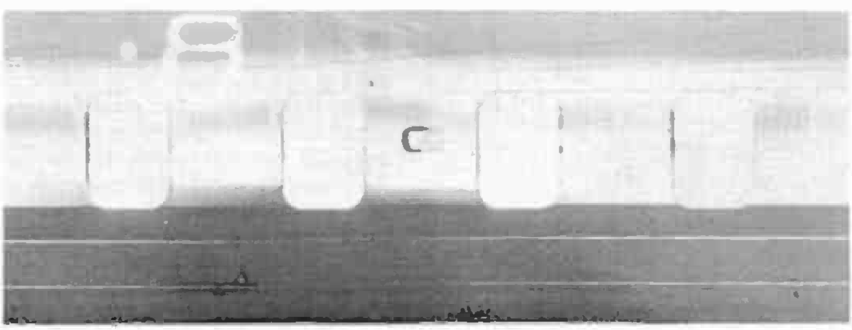


Fig. 18-209B. Flash bloop.

Slit Height	Loss in dB										
	0.5 kHz	1.0 kHz	2.0 kHz	3.0 kHz	4.0 kHz	5.0 kHz	6.0 kHz	7.0 kHz	8.0 kHz	9.0 kHz	10.0 kHz
1.00 Mil	0.1	0.3	1.1	2.6	4.9	8.5	14.0	31.0	—	—	—
0.75 Mil	—	0.2	0.6	1.5	2.6	4.3	6.5	9.7	14.0	23.0	—
0.50 Mil	—	0.1	0.3	0.6	1.1	1.8	2.6	3.7	4.9	6.5	8.5

Fig. 18-213. Loss for different slit heights with reference to frequency.

across a push-pull sound track, the noise caused by the splice appears across both plates of the photocell simultaneously (see part (a) of Fig. 18-209A) causing a simultaneous voltage to be generated by each half of the photocell. These two signals being equal in amplitude and in phase are cancelled out in the output circuit of the photocell; thus, no signal or noise is generated.

The foregoing statements apply only to a push-pull reproducer in which the cancellation is at least 30 dB. The subject of push-pull sound heads is discussed in Questions 19.97, 19.98, and 19.99

18.212 What is slit loss?—The loss caused by the aperture size in an optical system used for either the recording or the reproduction of sound tracks.

18.213 What is the loss for different slit heights with reference to a given frequency?—See Fig. 18-213.

18.214 What is the cutoff frequency for slits of different heights?

Slit Height in Inches	*Cutoff Frequency in Hz*
0.0012	6000
0.0010	7200
0.00075	9600
0.00050	14,400
0.00040	18,000
0.00014	51,000

18.215 What is the scanning beam height used in film recorders?—For variable-density recording, 0.5 mil. For variable-area, 0.2 mil. The mechanical slit in the optical system is considerably higher than the actual beam on the film. The final height at the film is obtained by optical reduction.

18.216 Show the details of a motion picture film processing machine.—An automatic film developing machine manufactured by Houston-Fearless is shown in Fig. 18-216A. It may be used for the development of reversal and negative or positive 16-mm motion picture film. Although the machine shown is designed primarily for the processing of reversal film, a simple rearrangement of the processing solutions or the film travel seqnence will enable it to process either negative or positive film.

The machine is designed to be portable, if necessity requires. Casters, leveling devices, and locks are provided at the front end of the cabinet. It may also be loaded in the daylight, making special partitions unnecessary. External water and power connections are included at the rear. The housing contains eight solution tanks, two water tanks, a drying cabinet, a film driving mechanism with a variable-speed transmission, a solution circulating pump, an air compressor, a refrigeration system, a solution heater, thermometer, and footage counter. It is equipped with three electric motors. Heat lamps for drying the film and an air filter are also provided. A 1200-foot magazine which is loaded in a dark room and placed on the machine is also provided. This type operation permits the machine to be operated in white light.

A continuous minimum tension on the film, that eliminates breakage and assures uniform processing results, is provided by a clutch-controlled film-driving mechanism. This mechanism eliminates stretching and contraction due to the process cycles, and makes mechanical control and manual adjustments unnecessary.

Replenishers are furnished for the first and second developers, the bleach, and hypo tanks. The replenisher bottles are fitted with glass petcocks and the rate of replenishment is adjusted by calibrations on the replenisher brackets. Fresh water replenisher solution is added at the bottom of the tanks and the solution level is maintained by overflow drains at the top of the tanks.

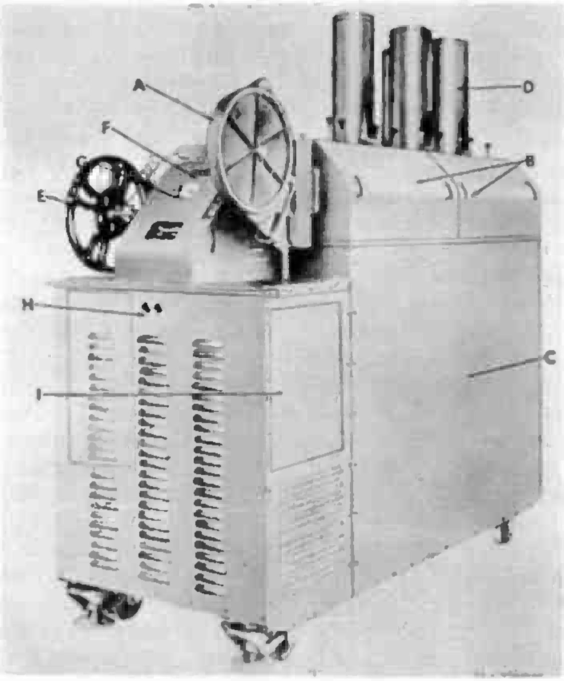


Fig. 18-216A. The 16-mm automatic film developing machine manufactured by Huston-Fearless.

The overflow is piped to a drain at the rear of the machine. The principal components, as indicated are: A film magazine for unprocessed film, B and C solution tanks in which rollers pass film through processing stages, D replenisher bottles, E take-up reel for processed film, F footage counter, G temperature gauge, H start and stop buttons, and I access door to solution valves.

The interior view of the machine with the top covers and replenisher bottles removed is shown in Fig. 18-216B. Starting at the lower right corner, the loaded film magazine (not shown) feeds the unprocessed film to the first group of rollers A. The film passes downward over another group of rollers and then upward to the next set B and so on until the last group H is reached. The direction of the film is then changed and passed over roller I and to the other side of the machine where it passes over other groups of rollers K to Q and to the take-up reel. Shaft R drives a pulley which drives the take-up reel spindle S by means of a V-belt. The transmission system consisting of a gear box T and a group of clutches U driven by chains Y is shown in the center compartment. The machine is driven by three motors, operating from 220 volts ac, single-phase, and consumes about 3.5 kW.

Fig. 18-216C shows the methods used for reversal or negative processing of film. The loading elevator holds 20 feet of film, the first developer tank 7 gallons of solution and 40 feet of film. The wash, bleach, clear, and second developer each hold 3.5 gallons of solution and 20 feet of film. The stop bath holds 3.5 gallons of solution and 26 feet of film. The fixer tank holds 4.5 gallons of liquid and 26 feet of film. The two wash tanks are filled with running water and will hold 26 feet of film each. The dry box will hold 110 feet of film. The total amount of leader required to run through the machine is 354 feet. The net weight of the machine is approximately 1100 pounds.

The interior of a film processing laboratory installed by Houston-Fearless in Hollywood, California, is shown in Fig. 18-216D. At A are the processing tanks with their rollers, and at B the dry boxes. The processed film is wound onto a reel at the extreme left (not

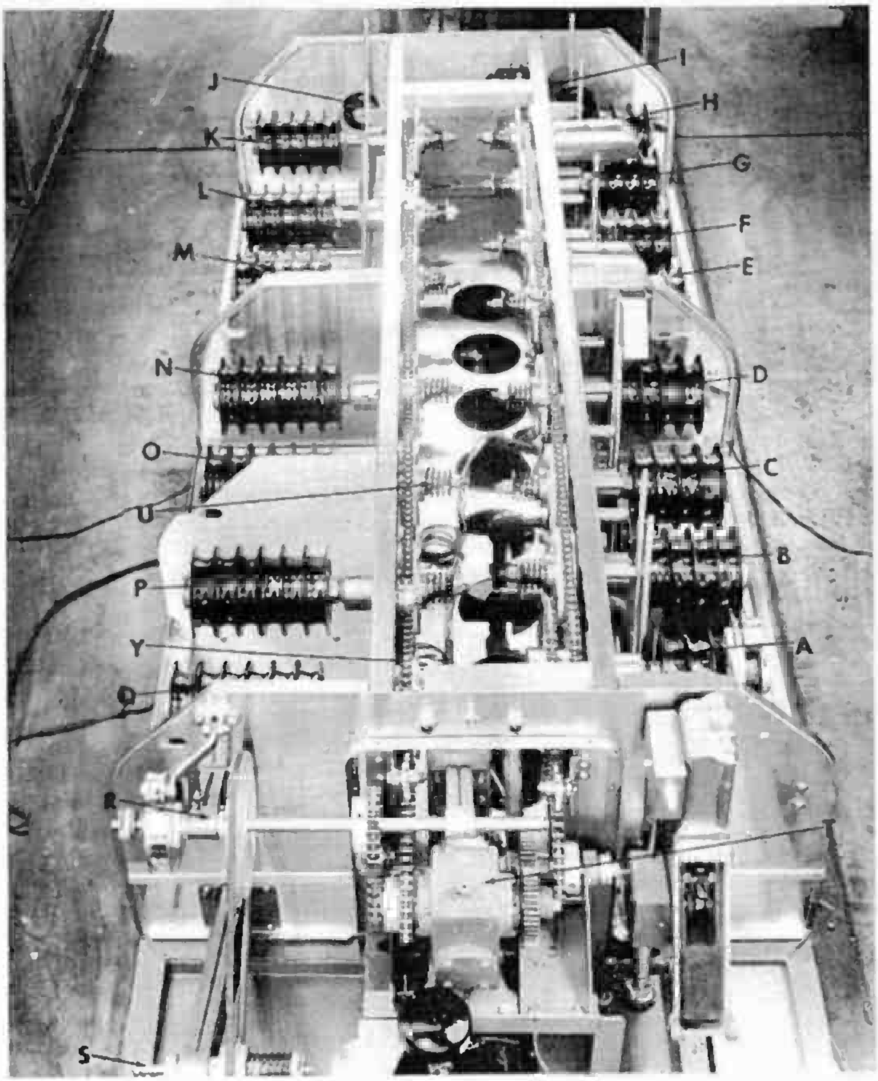


Fig. 18-216B. Interior view, from the operating end, of the Huston-Fearless 16-mm automatic film developing machine.

shown). Duplicate equipment appears at the right.

Fig. 18-216E shows the laboratory developed by Hi-Speed Equipment Division of the Artison Industries. This laboratory is designed for either color or black and white processing. The control equipment is pictured in Fig. 18-216F.

18.217 How are variable-area sound tracks processed?—Variable-area sound tracks are processed in the same solutions as used for regular positive film development. (See Question 18.218.)

18.218 How are variable-density sound tracks processed?—In the past years, it was necessary to process variable-density sound track in a special developer having low activity, in order that the development time for the required low gamma would not be too short for the linear speed of the average processing machine. With the emulsions in use, this practice is unnecessary, and they are processed in a regular negative developer. A typical developer is the Eastman Kodak D-76 developer.

18.219 Define the term "speed of film."—It is the inherent sensitivity of the emulsion for a specified set of conditions, relative to its exposure and development. If such conditions of exposure are known precisely, such numbers may be said to describe the absolute speed of the material. As a rule, only the relative speed is required. The term "speed" is defined as the reciprocal of the exposure required to produce a specific result and can be equated:

$$S = \frac{K}{E}$$

where,

K is constant,
E is the exposure corresponding to a point on the toe or shoulder of the curve at a specified density gradient, over a specified log E range. Various values of K are also used.

18.220 At what speeds are sound tracks developed?—The time of development and speed of the developing machine depend on the type of sound track being processed, the developing solution, and the required density. A variable-area sound track developing time could run from 4 to 12 minutes, with machine speeds of 30 to 250 feet per minute.

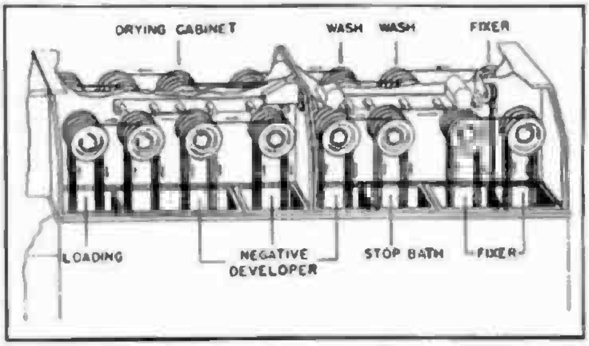
18.221 What are hand tests?—They are hand-developed recording tests for checking optical focus, track placement, azimuth, or any other tests requiring only a negative. They are not critical, in regard to a specific density. Hand tests are generally developed in the same solution used by the laboratory.

18.222 What is the average negative-fog density for variable-area sound track?—Approximately 0.01 to 0.05. This figure does not include the base density. With the base density included, it will measure about 0.26 and will vary with the time of developmeut. (See Question 18.271.)

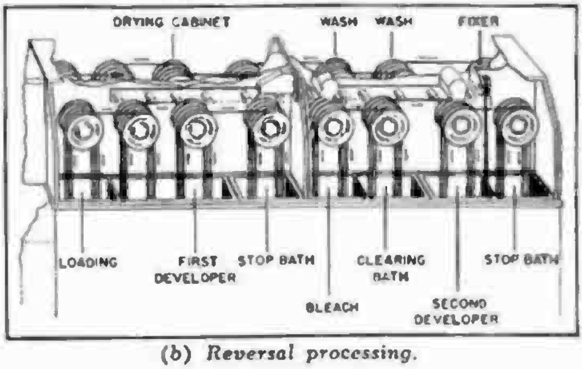
18.223 What is the average print fog density for variable-area sound track?—Approximately 0.12, including the base density. As for the negative, the amount of fog will depend on the time of development. (See Question 18.272.)

18.224 What is the cause of fog density?—It is a slight density produced on the film in areas where there has been no exposure. The net fog value for a given development time is obtained by subtracting the total base density and fix-out emulsion from the density of the unexposed, but processed, film. With modern films, the fog density is quite small.

18.225 What are the basic components of a cross-modulation oscillator?—A high-frequency oscillator for gener-



(a) Negative processing.



(b) Reversal processing.

Fig. 18-216C. Methods of setting up the Huston-Fearless 16-mm automatic film developing machine.

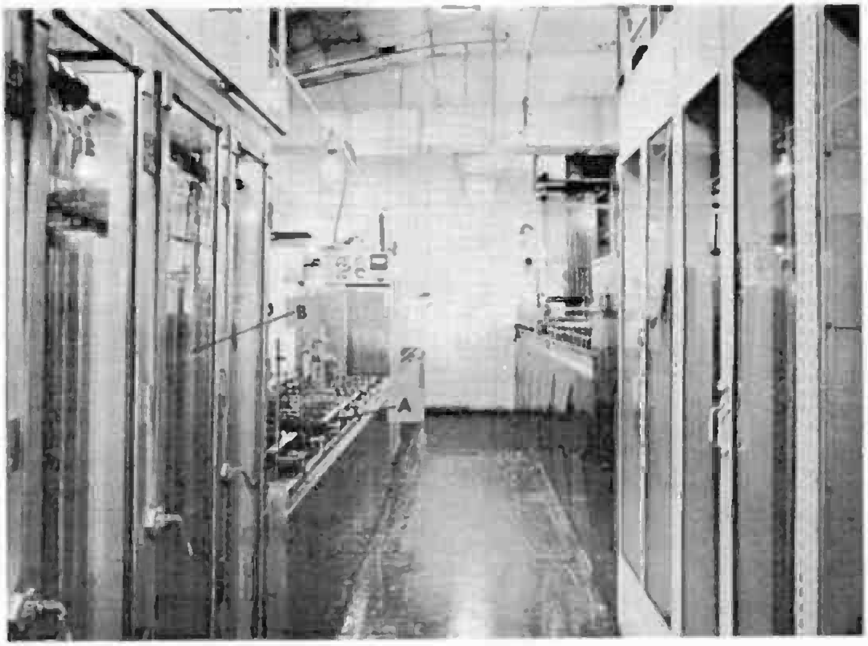


Fig. 18-216D. The interior of a typical film processing laboratory.

ating carrier frequencies and a second oscillator to modulate the carrier frequency at a lower frequency, generally around 400 Hz at 75-percent modulation.

18.226 What are the carrier frequencies employed for 35-mm cross-modulation tests?—6000 or 9000 Hz.

18.227 What carrier frequency is used for 16-mm cross-modulation tests?—A frequency of 6000 Hz modulated by

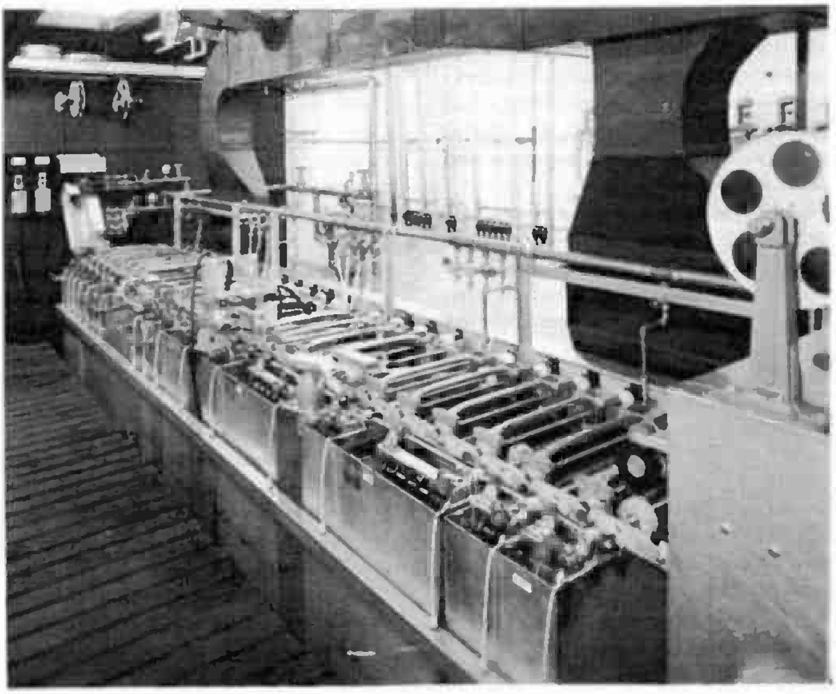


Fig. 18-216E. Film processing laboratory equipment manufactured and installed by Hi-Speed Equipment Division, Artison Industries Inc.

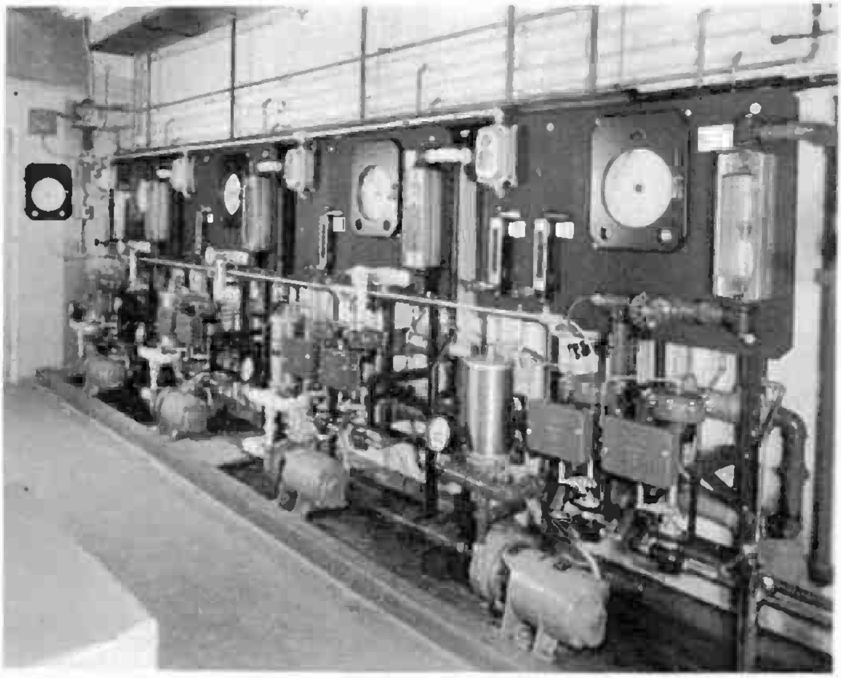


Fig. 18-216F. Processing control equipment for the film processing equipment shown in Fig. 18-216E.

400 Hz, at 75 percent (Westrex) or 80 percent (RCA).

18.228 To what percent is a light modulator deflected for cross-modulation tests?—Eighty-percent modulation. This modulation or deflection must not be confused with the 75-percent (Westrex) or the 80-percent (RCA) modulation of the cross-modulation oscillator carrier frequency. These are two separate and distinct functions.

18.229 Why is the light modulator deflected 80 percent for cross-modulation tests as opposed to 100 percent?—The use of 80-percent modulation prevents any possibility of overmodulation of the sound track.

18.230 Describe the design of cross-modulation oscillators.—Cross-modulation tests are used only with variable-area photographic film recording and are particularly well adapted to the measurement of photographic film distortion (image spread). Since the distortion occurs principally as a fill-in on the clear half of the recorded waveform, it is an unsymmetrical form of distortion. (See Question 18.232.)

Basically, a cross-modulation oscillator consists of two internal oscillators, one generating a carrier frequency, and the second generating a carrier-modulating frequency. Different carrier frequencies are employed, depending if the test is to be recorded on 16-mm or 35-mm film. At the present time, a carrier frequency of 6000 Hz is employed for 16-mm film, and 9000 Hz for 35-mm film, although older equipment may use 4000 and 6000 Hz, respectively. However, regardless of the carrier frequency, the modulating frequency is always 400 Hz. The oscillators must not have a THD greater than 0.5 percent.

The schematic diagram of an RCA MI-10803-B cross-modulation oscillator is given in Fig. 18-230A. Tube V1 functions as both a carrier oscillator and a mixer tube for the 400-Hz modulating frequency, with switch S1 selecting the desired carrier frequency. Tube V2 functions as a modulator, operating at 400 Hz and a reference frequency of 1000 Hz, and is coupled to V1 by means of two windings on the carrier oscillator coil. Switch S2 selects the type test to be recorded.

Mounted on the front panel are three continuously variable attenuators, P1, P2, and P3. These attenuators are for adjusting the amplitude of the carrier frequency, 1000-Hz reference frequency, modulated carrier, and the overall output level. As pointed out in Question

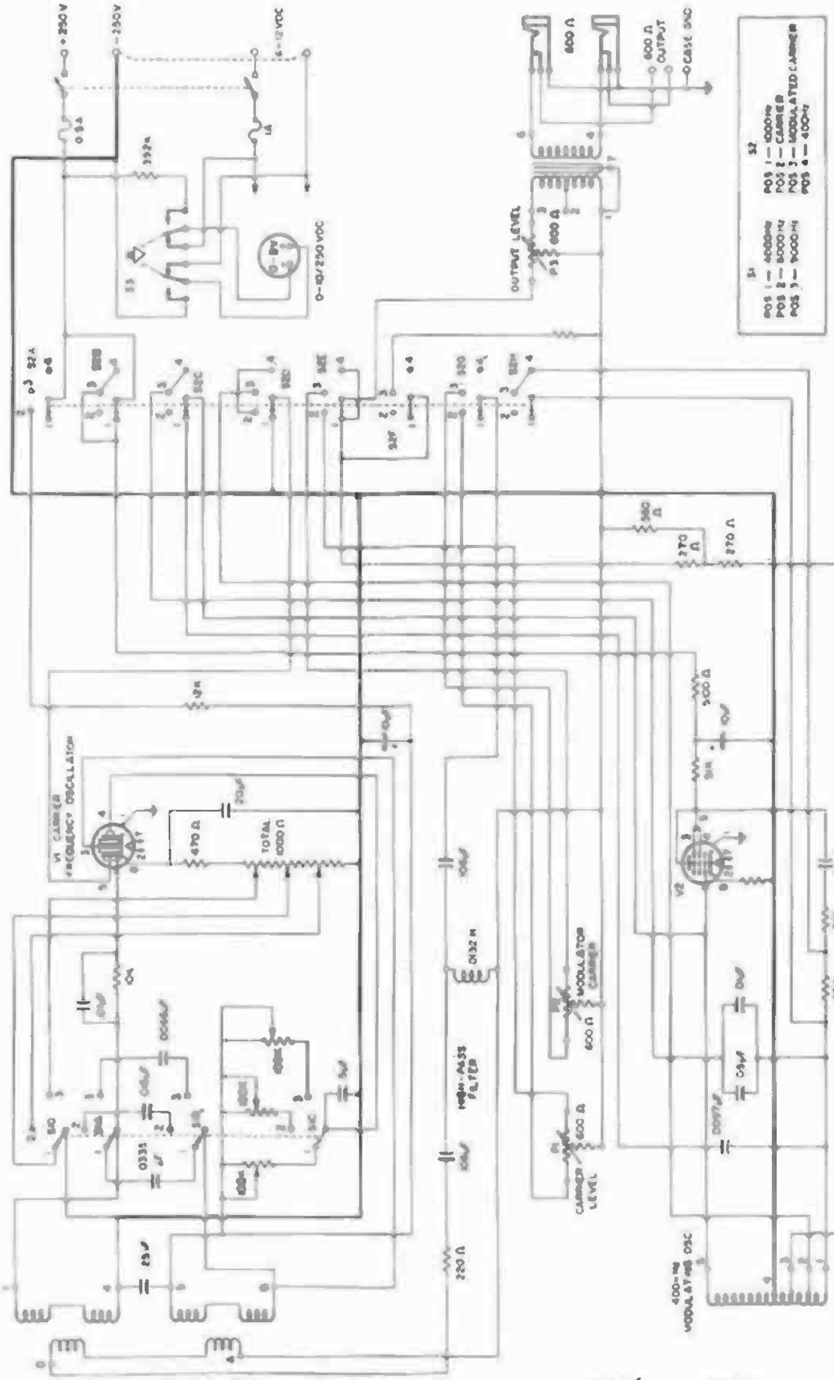


Fig. 18-230A. Schematic diagram for a cross-modulation oscillator.

18.232, the amplitude of the individual oscillators must be precisely set when recording tests. Attenuator P3, the output control, is always in the circuit, regardless of the tests being conducted.

When only the carrier frequency is being recorded, it is first passed through the carrier level attenuator, and then to output level attenuator P3. When recording the modulated carrier, the signal passes through the modulated carrier attenuator P2, then to the output attenuator P3. When recording a 400-Hz or 1000-Hz reference signal, the output level of V2 is fixed, but the overall level is adjustable by means of P3. Switch S1 is mounted inside the instrument and is generally left at a given carrier frequency. The use of the various controls is discussed in Question 18.232.

The schematic diagram for a Westrex Corp. Model RA-1397 cross-modulation signal generator is given in Fig. 18-230B, and although the circuit is somewhat different from the oscillator previously described it will produce the same results. Fig. 18-230C shows that two frequencies, A and B, are generated: one at 8600 Hz and a second at 9000 Hz (35-mm film). These frequencies are added together, resulting in a difference frequency C, with an envelope varying at a rate corresponding to the difference between the two frequencies, or 400 Hz.

When such a signal is applied to a recording system having nonlinear distortion characteristics (photographic

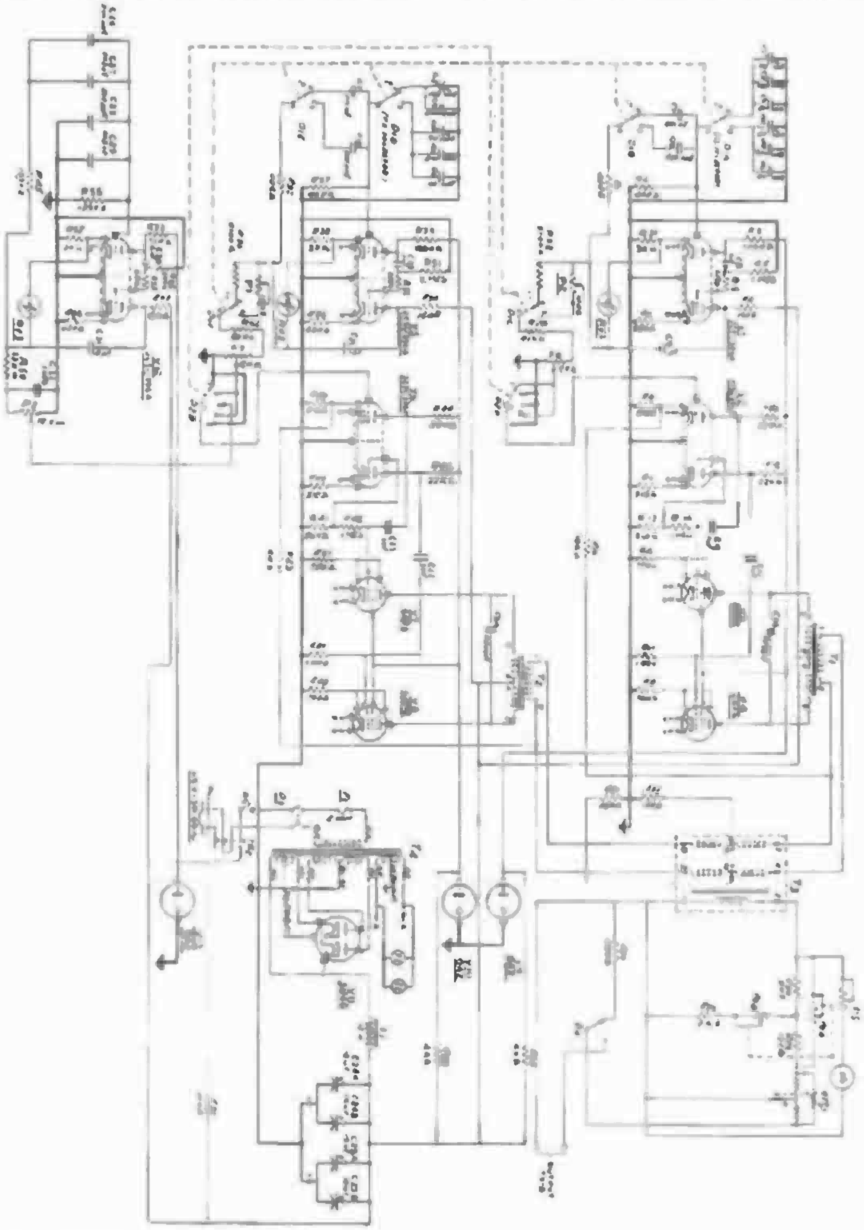


Fig. 18-230B. Schematic diagram of Westrex Model RA-1397 cross-modulation signal generator.

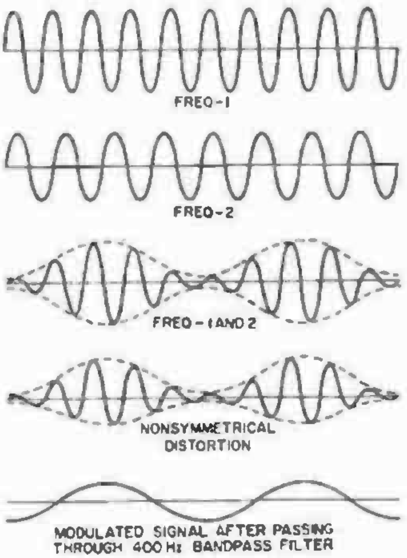


Fig. 18-230C. Waveforms encountered using the Westrex Model RA-1397 cross-modulation signal generator.

recording), new frequencies are generated which equal the difference frequency and multiples of the two principal frequencies (also the harmonics of each component frequency). Curve D illustrates a typical case of nonsymmetrical distortion. It should be noted that the peak amplitudes are greater on one side of the baseline than on the other. This gives rise to a low-frequency component shown at E (envelope shaped). This component is isolated for measurement by passing it through a 400-Hz read-out filter. The amplitude of the 400-Hz read-out signal is compared with a reference frequency of 400 Hz. The amplitude of the read-out frequency is measured and termed "cancellation," and its amplitude is expressed in decibels.

It should be observed that if the distortion is symmetrical with respect to the positive and negative halves of the signal, the cancellation is complete and no difference frequency distortion is present. Thus, a sensitive even-order distortion measurement is provided that does not indicate odd-order distortion.

A typical cross-modulation test made using the Westrex method of measurement is shown in Fig. 18-230D. The plot shows the cancellation for various combinations of print and negative density.

The schematic diagram given in Fig. 18-230B shows that the instrument consists of three RC oscillators, V1, V5, and V12, two amplifiers, a hybrid mixing circuit, and a power supply. Oscillators V1 and V5 consist of 2C51/396A dual-triode tubes, connected as resistance-capacity tuned oscillators, with a thermistor in the feedback loop for controlling the amplitude of oscillation. Oscillator V1 provides frequencies of 3600 and 8600 Hz. Oscillator V5 generates frequencies of 4000 and 9000 Hz. The modulating oscillator generates a 400-Hz frequency.

For 16-mm film, frequencies of 3600 and 4000 Hz are employed; for 35-mm film, 8600- and 9000-Hz frequencies are used. Individual controls for adjusting the oscillator amplitudes, output level control, and an output meter, are provided on the front panel, along with several other controls for circuit adjustment when first installed.

Relative levels of the two superimposed signals are initially adjusted for a 2.5-dB differential, which results in a 75-percent modulated envelope. The peak values of the two frequency signals and the 400-Hz reference signal are made equal in amplitude and applied to the recording system at a peak level 2 dB down from the 100-percent recording level of the channel.

The motion picture industry has standardized cross-modulation tests in order that the values of cancellation measured will have a definite meaning in the terms of distortion (cancellation). A cancellation of 30 dB is consid-

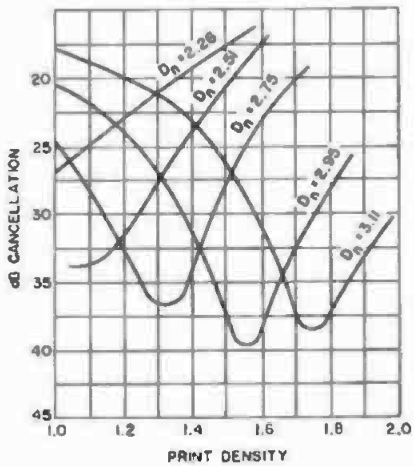


Fig. 18-230D. Cross-modulation distortion plot using Westrex cross-modulation signal generator.

ered to be the minimum acceptable value for good quality recording. The reader is referred to USASI (ASA) Standard PH22-52-1960 reaffirmed December 1967.

18.231 Show a block diagram of a cross-modulation oscillator connected to a recording channel.—As a rule, cross-modulation oscillators are patched directly into the recording amplifier that drives the film recorder. However, in some instances, the gain of the recording amplifier may not be sufficient, in which instance the cross-modulation oscillator may be patched into an amplifier ahead of the recording amplifier, or into one of the mixer inputs, as shown in Fig. 18-231. Care must be taken that all equalization and any filters between the mixer console and the recorder are patched out of the circuit. This does not include the light-valve equalizer (if such a device is being used). The cross-modulation oscillator is used as described in Question 18.232.

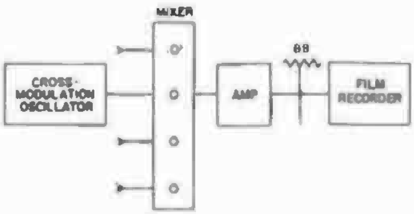


Fig. 18-231. The cross-modulation oscillator connected to the recording channel for recording a cross-modulation test.

18.232 Explain the theory of cross-modulation tests and how they are made.—Because of the characteristics of motion picture film, complete opacity and transmission are not possible. For high-quality recording and reproduction, it is essential for the print to have a high ratio of signal to noise and low distortion. To determine the correct negative and print densities and to maintain a uniform quality from day to day, it is necessary that tests be made frequently to assure the correct development of both the negative and the print, and to determine the correct negative density when an emulsion is changed The following discussion pertains to both light-valve and galvanometer light modulators recording variable-area sound track. For variable-density sound track, an intermodulation oscillator and analyzer are used as described in Question 18.274. Although the following examples are based on a print density of 1.30, the procedure is the same for any density.

For satisfactory reproduction, the image of the sound track on the print must be identical to the image formed on the negative by the optical system of the recorder. However, such conditions do not prevail because of the diffusion created by the light-sensitive emulsions, the wavelength of the exposure light, and the angle of incidence at which the exposure light strikes the film. These factors and others cause the image on the print to become larger or smaller than the image originally formed by the recorder optical system. This phenomenon is called image spread. Image spread is not important at frequencies below 1000 Hz, but at the higher frequencies it is all important and is the controlling factor in the final quality of reproduction. Because the image on the print will always be larger than the optical image that formed it, image spread is purposely introduced in the negative sound track for compensation.

The images on the print are reversed from those on the negative. The blacks of the negative become white on the print and the whites of the negative become black on the print. If image spread is induced in the negative, the white spaces between modulations on the print become smaller; thus, the print image is brought to the exact size of the original image formed by the recorder optical system. Because image spread at 1000 Hz is quite small, this frequency is used as a reference frequency for 35-mm film measurements. For 16-mm film, 400 Hz is used.

Quality control of sound tracks using the cross-modulation method of measurement consists of recording a high frequency modulated by a lower frequency. If too great an image spread is present in the print, the valleys between the modulations will be filled in. On the other band if too great an image spread is present in the negative, the valleys between the modulations of the print will be widened because of the reversal process between the negative and the print. Typical cross-modulation recordings are shown in Fig. 18-297.

When measuring a cross-modulation test, the reference frequency is first reproduced and used as an index for the balance of the test. Next, the high-frequency signal is reproduced to measure the high-frequency loss. Finally, the several cross-modulation tests are reproduced. The signal measured in a cross-modulation test will be the same as the modulating frequency used to modulate the carrier frequency when the tests were recorded.

During the reproduction of the cross-modulation frequencies, a bandpass filter is used to permit only the modulating frequency (400 Hz) to be measured, while rejecting the high-frequency carrier. The amplitude of the 400-Hz modulating signal, when measured, will be governed by the amount of image spread existing in the negative and print. When the image spread of the negative matches or cancels the image spread of the print, the amplitude of the 400 Hz will be at a minimum. This is referred to as the cancellation voltage.

The high-frequency film loss is measured by using either 9000 Hz or 6000 Hz. The difference in amplitude between the 1000-Hz reference frequency and the 9000- or 6000-Hz signal will be indicative of the high-frequency loss. High-frequency loss is compensated for in the film loss equalizer; however, a small variation of plus or minus 1 to 2 dB will be noticed from day to day. This must be expected, as processing, film emulsions, printing, and other factors all have their effects on the high-frequency response. If this loss is important, it may be compensated for during the rerecording operations or during the transfer from magnetic film to optical sound track. Generally, if the loss is within the limits mentioned, it may be ignored.

To illustrate the procedure used for making cross-modulation tests, it will be assumed a 35-mm variable-area studio-type recorder using either a light valve or galvanometer is to be put into service.

Before starting the cross-modulation tests, a series of exposure lamp tests must be made, using the same emulsion that is to be used for the cross-modulation tests. Start by adjusting the lamp current to 5.6 amperes and expose about 10 feet of wide track on the film. Increase the lamp current to 5.7 amperes and make a second exposure. Continue this routine up to 6.4 amperes, increasing the current in steps of 0.10 ampere. Request the process laboratory to develop this negative to a gamma of 2.5 or 3.0. After development, measure and log the lamp current versus density as these exposures will be referred to again later in the tests. If the recording channel is to be used for making cross-modulation tests, all filters and equalizers must be *patched out* to provide a flat frequency response within plus or minus 1 dB up to the frequency used for measuring the high-frequency loss.

Next, a tabulation sheet similar to that shown in Fig. 18-232 is prepared. On this sheet are entered data pertinent to the tests. At the left in column A are the modulating and carrier frequencies used for both 16-mm and 35-mm film tests. In column B is entered the percentage of modulation of the light modulator, which is generally 80 percent. Using 80-percent modulation of the light modulator precludes any chance of overmodulation and clipping of the peaks which would result in a distorted sound track. Photometer settings are entered in column C, if used.

The lamp current used for each cross-modulation test is entered in column D and the measured densities (wide-track portion) in column E. Print densities asked for and received back from the laboratory are entered in columns F to J. Additional space is provided for entering the emulsion, edge, and light modulator numbers, negative fog density, recorder number, and type sound track recorded. At the bottom of the page are spaces for entering the density tolerances for a minimum cancellation of 30 dB and the high-frequency loss.

Referring to previously logged lamp current versus density tabulation, select a lamp current for a density of about 2.50 to be used as the midpoint exposure for the cross-modulation tests. The selection of a density of this value aids in plotting the cancellation as it will likely occur between a density of 2.4 and 2.6. Enter this lamp current (for this illustration it will be assumed to be 6.0 amperes) in column D, lines 1, 3, and 9 of Fig. 18-232. Fill in lines 5 to 13 starting at a lamp current of 5.6 amperes and increasing the current in steps of 0.10 ampere up to and including 6.4 amperes. Line 2 and line 5 of column D are the same current (5.6 amp) as are lines 4 and 13 (6.4 amp).

The lamp currents in column D are those which will be used when recording the cross-modulation tests. If, however, the highest negative density is about 2.0 or slightly less, select as the midpoint of the tests a lamp current that will result in a density about 0.2 to 0.3 below the maximum density, and enter the other lamp currents in the manner described previously. It will be noted the lamp current for the 1000-Hz reference frequency and the second 9000-Hz high-frequency loss test are the same as for the midpoint of the 400/9000-Hz cross-modulation test (line 9).

The following procedure is that used when making cross-modulation tests using the RCA cross-modulation oscillator. Insert the optical system periscope in the recorder optical system and observe the waveforms of the cross-modulation oscillator in the following manner: (If the recorder does not use a periscope, the following tests may be

EMULSION DATA TEST—No. 212

Date 10/25/67 Type of Test X-MOD Emulsion No. 5375-103-14 Edge No. 59N5—55731

Engineer Gunselle, J.H. Laboratory Trans-Canada Films

Neg./Print TCF Recdr. No. PR-31 #1 Lv. ______ Gal. 1284

Fog 0.06 35MM 16MM B & W ~~Color.~~

PRINT CANCELLATION

	A	B	C	D	E	F	G	H	I	J
	Freq.	% Mod.	Photo	Amps.	Neg. Den.	Ask 1.20 / Rec. 1.18	1.30 / 1.29	1.40 / 1.37		
1	1,000	80%		6.0	2.57	0	0	0		
2	9,000			5.6	1.76	-3.0	-2.5	-2.5		
3	9,000			6.0	2.68	-3.0	-4.0	-4.5		
4	9,000			6.4	3.00	-4.5	-4.5	-5.0		
5	9/4			5.6	1.80	-29.0	-26.0	-24.0		
6	9/4			5.7	2.10	-38.0	-32.0	-27.0		
7	9/4			5.8	2.25	-45.0	-40.0	-33.0		
8	9/4			5.9	2.50	-28.0	-38.0	-42.0		
9	9/4			6.0	2.60	-27.0	-28.5	-32.0		
10	9/4			6.1	2.74	-24.0	-25.5	-27.5		
11	9/4			6.2	2.87	-21.5	-22.0	-22.5		
12	9/4			6.3	2.96	-19.5	-20.0	-21.0		
13	9/4			6.4	3.00	-18.0	-19.0	-19.5		

CONCLUSIONS

2.32 Neg. Den. for 1.30/1.~~20~~ Print

Tolerance for 30 dB Cancellation

Max. Neg. Density 2.38

Max. Print Density 1.35

Min. Neg. Density 2.11

Min. Print Density 1.25

9,000 Cycle Output -3dB

Fig. 18-232. Tabulation sheet for cross-modulation tests.

made by observing the deflection of the light modulator on the monitor card.)

1. Connect the output of the cross-modulation oscillator to either the recording amplifier or to one of the mixer inputs. Set the mixer control to the average or normal signal level for the channel and leave it set throughout the balance of the tests.
2. Set the cross-modulation oscillator to the 1000-Hz position and adjust the output control for this frequency for exactly 100-percent deflection of the light modulator as observed in the periscope or on the monitor card.
3. Set the cross-modulation oscillator to the 9000-Hz position and adjust the output at this frequency for 100-percent deflection of the light modulator.
4. Set the cross-modulation oscillator to the 400/9000-Hz position and adjust the output control for these frequencies to exactly 100-percent modulation of the light modulator.
5. Reduce the gain of the recording amplifier 2 dB. This will then result in a recording level of 80-percent modulation of the light modulator. After the above adjustments are completed, the controls are left set throughout the balance of the tests.
6. Adjust the exposure lamp current to 6.0 amperes.
7. Start the recorder running with the light modulator switch in the OFF position. When the recorder is up to speed, throw the light modulator switch to RECORD and expose 25 feet at 1000 Hz as a reference frequency.
8. Throw the light modulator switch to OFF position, including the bias.
9. Record 5 feet of wide track. This exposure will be used later for density measurements. (A wide-track exposure is made at the end of each cross-modulation recording.)
10. With the light modulator switch in the OFF position (including bias), switch the cross-modulation oscillator to 9000 Hz and change the lamp current to 5.6 amperes.
11. From the wide-track position (9) return to the normal recording position and record 10 feet at 9000 Hz.
12. Repeat operations (10 and (11), in essence, except for changing the lamp current to 6.0 amperes and then to 6.4 amperes.
13. Throw the light modulator switch to the OFF position, including bias. Set the cross-modulation oscillator to the 400/9000-Hz position. Record 10 feet of sound track for each lamp current in column D, starting at line 5 (5.6 amp) and continuing up to 6.4 amperes with a section of wide track between each test.
14. Send the exposed sound track to the process laboratory with instructions to process it to exactly the same value of gamma to which the lamp test was developed. Laboratories test their developing solutions daily to assure a uniform product from day to day. To arrive at a given gamma, the developing time of the machine is altered by changing the linear speed.
15. Request three prints at densities of 1.20, 1.30, and 1.40 At times, it might be advantageous to request prints at densities of 1.10 and 1.50, in addition to the above to extend the range of measurement.

Although the foregoing tests appear rather complicated and difficult to make, they must be made in the order given and while the recorder is in motion. The cross-modulation tests may be used with any type variable-area film recorder using either a light valve or a galvanometer. The recording of cross-modulation tests for push-pull recorders is described in Question 18.251. For 16-mm recorders, the procedure is exactly the same as for 35 mm, except that 6000 Hz is used for the carrier frequency and 400 Hz is used for the reference frequency. If optical reduction prints are used, the cross-modulation test must be made to include the final 16-mm print.

The cross-modulation test used by Westrex for variable-area recording differs somewhat from that used by RCA. Two frequencies of 8600 and 9000 Hz are combined in a hybrid coil to provide an envelope varying at a rate of 400 Hz (the difference frequency between the two frequencies).

The application of such a complex waveform to a recording system having nonlinear characteristics causes the generation of new frequencies which equal the difference between the two high frequencies, (400 Hz), also, multiples of the difference frequency, and harmonics of the two high frequencies (8600 and 9000 Hz).

In a recording system having non-symmetrical distortion, the peak excursions of one side of the sound track will have greater amplitude than the other. This condition causes the generation of a low-frequency component which is later isolated during the reproduction of the test track, by passing the recorded signal through a 400-Hz bandpass filter.

The amplitude of this low-frequency component is compared with a reference frequency of 400 Hz and is called the "cancellation." It is expressed in decibels.

The relative levels of the three signals (8600, 9000, and 400 Hz) have been standardized to permit the amount of cancellation to be stated in terms of the amplitude of the nonlinearity. The relative level of the 8600- and 9000-Hz signals is set to result in 75-percent modulation; that is, the 8600 Hz is set 2.5 dB lower in amplitude than the 9000-Hz signal amplitude.

If the distortion of the recorded test is symmetrical with respect to both halves of the signal, cancellation will be complete, and no difference frequency (400 Hz) will be measured. This provides only a test for even-order harmonic and does not indicate the degree of distortion for odd harmonics.

When the Westrex and RCA method of measuring cross modulation is applied to a given recording channel, the values of cross modulation measured will not be exactly the same for both methods, but only approximately the same. The minimum acceptable cancellation for the Westrex method is also 30 dB for both 16- and 35-mm sound tracks. The test is made using 80-percent modulation of the light modulator. For 35-mm sound tracks the frequencies are 8600 and 9000 Hz with a 400-Hz reference frequency. For 16-mm sound tracks the high frequencies are 3600 and 4000 Hz.

Measurement of the test is handled in the same manner as that described for the RCA cross-modulation test, requesting from the laboratory several different negative and print densities. The combination of negative and print densities that results in the greatest value of cancellation is the correct one, and should fall at a print density between 1.30 and 1.50.

Cross-modulation test equipment should be tested frequently as described in Question 23.185.

18.233 How are cross-modulation tests measured?—By filtering out the high-frequency component (carrier frequency) and measuring the amplitude of the lower frequency (fill-in). The greater the fill-in between modulations, the less cancellation obtained.

18.234 What are the basic components of a cross-modulation read-out set?—A 400-Hz bandpass filter for filtering out the high-frequency carrier and allowing only the 400-Hz component to be measured. A pad with a loss equal to the insertion loss of the filter is inserted by a key switch when the filter is removed from the circuit. This is necessary to prevent affecting the

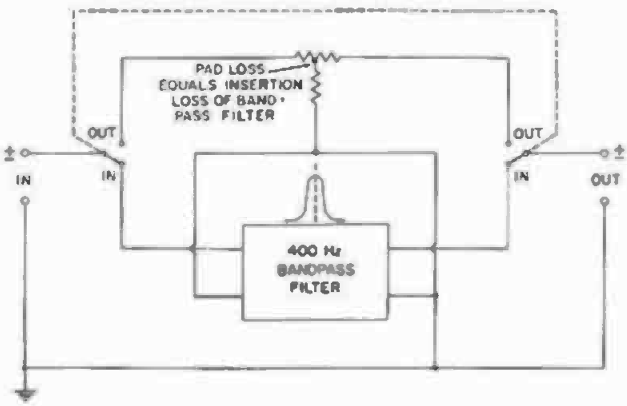


Fig. 18-234. Diagram for a cross-modulation test readout set for measuring tests made with either Westrex or RCA cross-modulation oscillators.

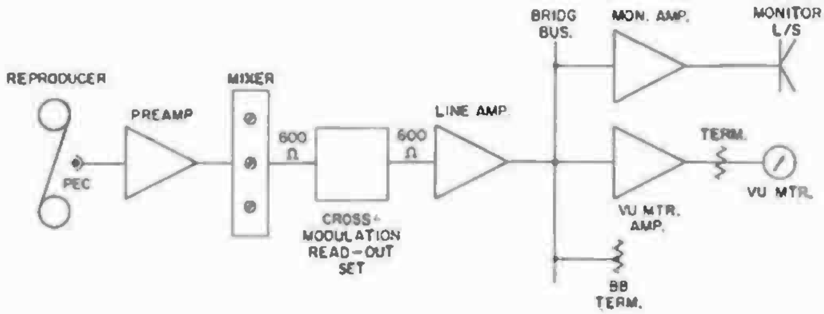


Fig. 18-235A. Using a cross-modulation readout set with a recording channel.

level of the single-frequency measurements which would result in erroneous readings. If the device is used in circuits that are balanced to ground, repeat coils should be connected in the input and output as the case may require. This is important as high-frequency leakage may result because of unbalance and give rise to false readings at 9000 Hz. A diagram of a typical cross-modulation test set is shown in Fig. 18-234. (See Question 5.30.)

18.235 Show a block diagram for connecting a cross-modulation read-out unit.—Because a considerable amount of amplification is required following a cross-modulation read-out unit, either the recording channel or a vacuum-tube voltmeter will be required for its operation. In the first method of connection, the read-out unit is connected in the recording channel as shown in Fig. 18-235A. Care must be exercised in order that all impedance matches are satisfied, and that the channel has a uniform frequency response up to and including the carrier frequency used for the cross-modulation measurements. A VU meter is connected across the output of the amplifier normally used to drive the recorder. This combination is used to measure the cancellation levels of the cross-modulation test. If any doubt exists as to the frequency response, a measurement is made from the reproducer to the output of the amplifier driving the VU meter, using a multifrequency test film. Correction factors (if required) can then be applied to the measured response at the carrier frequency. The measurement is made using the procedure given in Question 18.237.

A second method, shown in Fig. 18-235B, uses the recording amplifier connected at the output of the cross-modulation read-out unit. It is important that the amplifier used for driving the VU meter have sufficient gain and output to deflect the VU meter up to and including about plus 38 dBm (plus 40 dBm equals 10 watts). Additional gain may be obtained, if needed, by operating the amplifier without a termination, provided the frequency response is not affected and the amplifier is not driven into excessive distortion. The procedure is the same as for the first method discussed.

A third method, the preferred one, is shown in Fig. 18-235C. Here, a vacuum-tube voltmeter is connected at the output of the read-out unit. The output of the filter must be terminated in its normal load impedance because of the high impedance input of the vacuum-tube voltmeter. Using this method, the signal

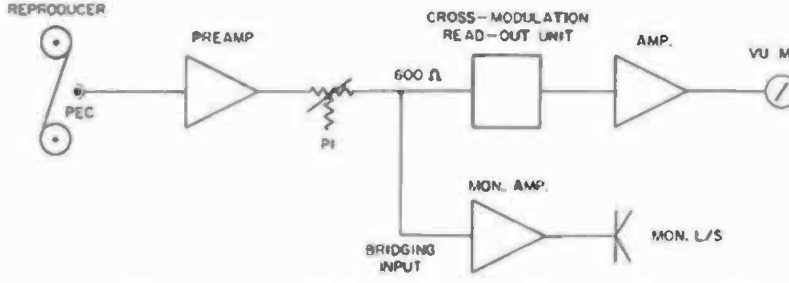


Fig. 18-235B. Using a cross-modulation readout set with an amplifier and VU meter.

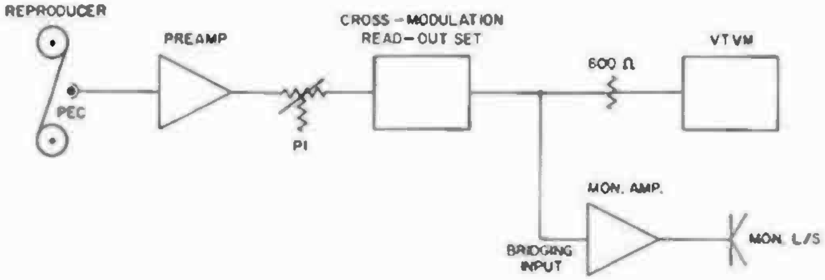


Fig. 18-235C. Using a cross-modulation readout set with a vacuum-tube voltmeter. This is the preferred method.

from the reproducer is adjusted for zero dBm output at 1000 Hz, and the cancellation read directly in relation to the 1000-Hz reference frequency.

18.236 Are special calibrations required for the connections shown in Fig. 18-235B?—Yes. A special scale is made (Fig. 18-236) for the amplifier used in combination with the VU meter in Figs. 18-235A and B. This scale is placed around the outer edge of the 34-dBm calibration of the VU meter attenuator scale. It will be observed that the minus 34-dBm calibration of the new scale is opposite the plus 4-dBm calibration of the normal meter scale. If a vacuum-tube voltmeter is employed as shown in Fig. 18-235C, the amplifier is not required.

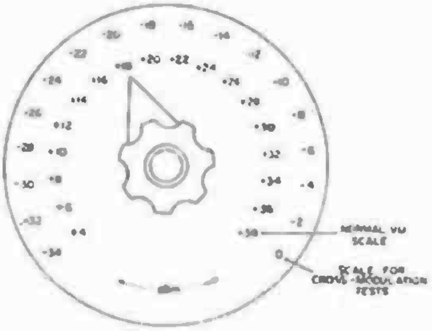


Fig. 18-236. A special VU meter attenuator scale for cross-modulation test measurements.

18.237 What is the procedure for measuring a cross-modulation test?—First, measure the visual density of the negative test tracks made as described in Question 18.232. Enter the densities in column E of the tabulation sheet (Fig. 18-232). Next, measure the print densities and enter them in columns F, G, and H. As the received densities will not always be the exact value ordered, enter both the requested and received densities at the top of the last three columns. This information will prove useful later when making corrections or estimates. The actual measurements are made as follows:

1. Set the switch on the filter to FILTER OUT. Thread up the 1.20 density print on the sound head. When the machine is up to speed, adjust the gain of the amplifiers and the pot P1 at 1000 Hz for an output level of zero on the new volume indicator scale. (This is opposite the plus 38-dBm calibration on the normal meter attenuator scale.) After setting the level, pot P1 is left unchanged for the balance of the measurements associated with the 1.20-density print. All output levels are read on the new scale of the volume indicator meter. Read the output levels of the three 9000-Hz recordings and enter them in column F. In reality, the output level is started at a plus 38 dBm and, as the levels drop off, the sensitivity of the meter is increased by removing loss from the meter attenuator. Starting at a high output level, using the inverse scale facilitates the reading of the meter and requires no mental corrections.
2. After reading the amplitude of the three 9000-Hz recordings, throw the filter switch to FILTER IN and read the amplitude of the nine 400/9000-Hz recordings. The filter removes the 9000-Hz signal and leaves only the 400-Hz signal. Read the levels by increasing the sensitivity of the volume indicator meter. Enter the levels in column F, as shown in Fig. 18-232. It will be noted as the cancellation increases, the level of the 400-Hz signal drops and, at a given density,

starts to return to a higher level. At the lower cancellation levels, it will be necessary to read down on the meter scale which is added to the meter attenuator. For example, if the attenuator scale reads minus 34 dBm and the meter indicates a minus 6 dB, the level is minus 40 dBm.

3. Measure the 1.30 and 1.40 densities in a similar manner. The results of these tests are then plotted as showu in Fig. 18-232.

18.238 How are cross-modulation tests plotted?—On cross-sectional paper, as shown in Fig. 18-238A. The graph is prepared by entering the print densities and cancellation in decibels at the left. A line is drawn across the paper at the 30-dB cancellation point P. It will be assumed for this example of plotting that the print density is to be held to 1.30, for reasons as explained in Questions 18.239 and 18.240. Under these conditions, the negative density will be varied to obtain a print density of 1.30. Therefore, the negative density will vary for different amounts of cancellation. The first curve plotted is for the 1.18 received print density, by taking the data from columns E and F of Fig. 18-232.

As an example, a print density of 1.18 and a negative density of 1.80 have a cancellation of 29 dB. Enter this value on the graph at the junction of the 29-dB line and a negative density of 1.80. Next, plot for a negative density of 2.10 and so on for all the negative densities in the 1.18 column. In this manner, the cancellation for all three print densities and negative densities are plotted. It will be noted that after a maximum cancellation has been reached, the curve rises again. The rising cancellation is plotted at the right and labeled for each density.

After plotting the three densities, an X is made where the curves cross the 30-dB cancellation line. Next, two lines D1 and D2 are drawn in at priut densities of 1.25 and 1.35. Lines Q and R are drawn in by pointing off the three densities 1.18, 1.29, and 1.37 above the X point on the three cancellation curves, as indicated by the dotted lines. This is done for both the decreasing and increasing cancellation curves. If the points for lines Q and R do not result in a straight line, an average value for the three is taken. A vertical line, S, is drawn downward from the point where lines D1 and Q meet. Also, line "T" from the point where D2 and R meet. Read the negative densities at the lower ends of lines S and T and compute their average (2.11 plus 2.55 divided by 2). This results in a density of 2.33. This is

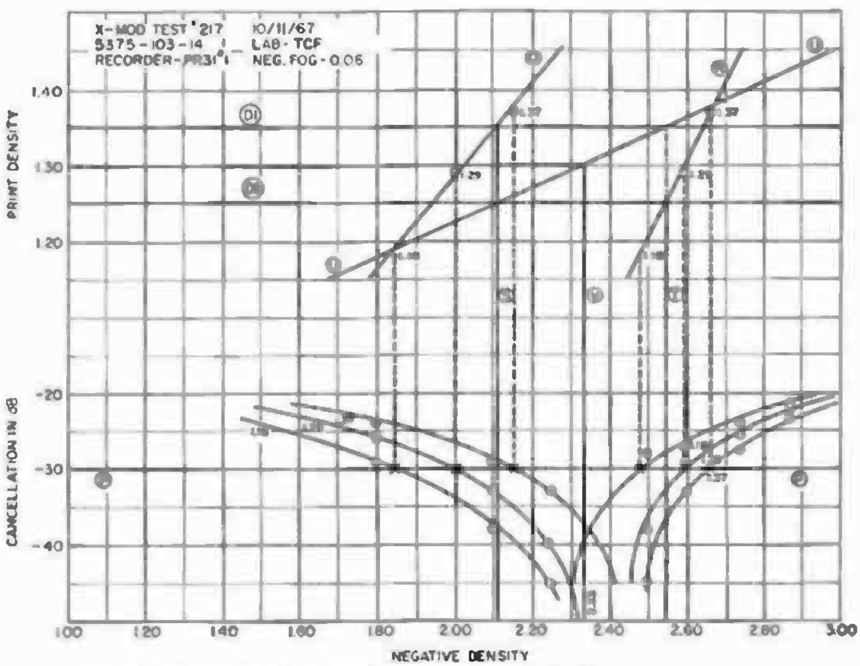


Fig. 18-238A. Method of plotting cross-modulation measurements.

the negative density required for a print density of 1.30. The cancellation of this negative and print density will be approximately 45 dB. It should be noted that the loss at 9000 Hz will be about 3.0 dB.

The diagonal line U is drawn from the junction of line D2 and line S to the junction of line D1 and line T. A vertical line V is drawn downward from the point where a print density of 1.30 crosses line U. This ends at a negative density of 2.33. Line U may be employed for selecting the correct print density when the negative density is incorrect, or vice versa. For example: What is the correct print density for a negative density of 2.60? Answer, 1.36. Or, what is the negative density for a print density of 1.23? Answer, 2.02.

In Fig. 18-238B are shown the results of an unsatisfactory cross-modulation test. It may be seen that the negative density can only be varied plus or minus 0.04 point from the optimum density before the cancellation becomes less than 30 dB. After plotting up the results and determining the optimum negative density for a given print density, the laboratory is instructed to develop the negative to within plus or minus 0.05 point of the specified density. The same tolerance is requested for the print. The spread between the curves plotted for a given negative and print density should not be less than plus or minus 0.22 from the optimum negative density. This will allow for variations in processing.

18.239 Why is the sound print density held constant and the negative density varied?—To facilitate the ordering of prints at a future time. All prints are made to a given density, thus permitting a better quality control.

18.240 Why is a density of 1.30 to 1.40 used for variable-area prints?—Because this range of print density permits the negative density to be varied within a practical range, and with good signal-to-noise ratio. Studios using variable-area recording systems (galvanometer or light valve) use print densities between 1.30 and 1.50, the average being 1.40.

18.241 What is the minimum acceptable cancellation for cross-modulation tests?—It has been determined from exhaustive listening tests that the minimum acceptable cross-modulation is 30 dB. This is true for both 16- and 35-mm film. However, good laboratory control should yield at least 36 to 40 dB, with 45 dB not uncommon. For direct-positive recording, the cancellation should be on the order of 36 dB or greater. (See Question 18.246.)

18.242 What are the causes of poor

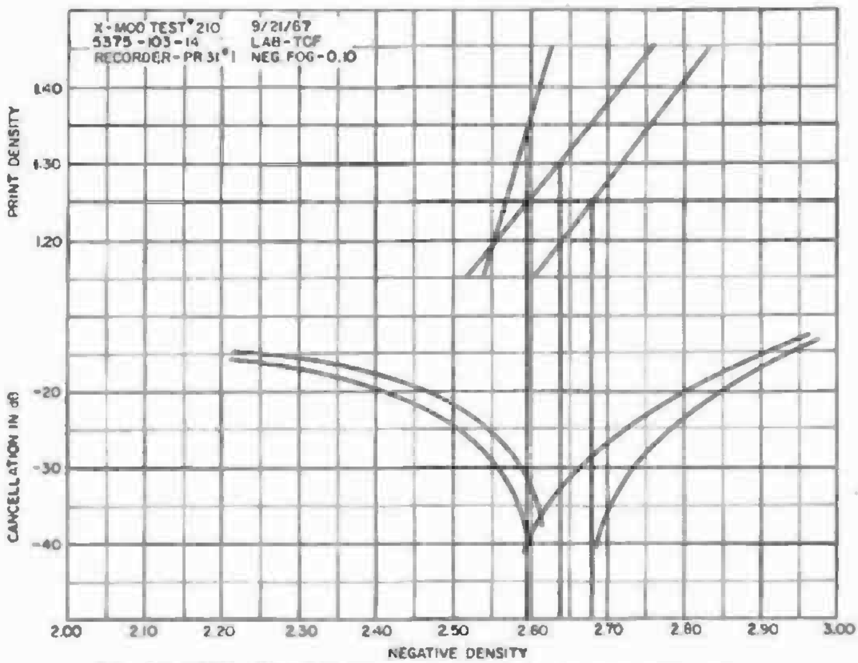


Fig. 18-238B. Unsatisfactory cross-modulation measurements.

cross-modulation cancellation?—The primary cause is image spread. If this is not the cause, there are many other items to be checked. The optical system of the recorder may be out of adjustment, negative or print developer not correct, poor contact between the negative and print in the printing machine, over- or underexposure of the negative or print, recorder or reproducer optical systems out of focus, high distortion in the 400-Hz oscillator of the cross-modulation oscillator, noise in the cross-modulation measuring circuit, excessive flutter in the recorder or reproducing system or printer, or the frequency response of the measuring circuit is inadequate. The frequency of the 400-Hz oscillator in the cross-modulation oscillator may not correspond to the center frequency of the bandpass filter used in the readout panel.

As may be seen from the foregoing, there are many factors which can contribute to poor cancellation; therefore, each and every piece of equipment from the original recorder must be functioning at its peak performance, if a maximum cancellation is to be achieved. This is the reason why a tight quality control must be established to obtain quality sound tracks.

18.243 If the density of a variable-area sound track exceeds the optimum value, how is the reproduction affected? —The high-frequency response is reduced and the high-frequency distortion is increased. This is often mistaken for excessive sibilance. If the density is too light, the signal-to-noise ratio is decreased and the background noise is increased.

18.244 Define the term "balanced density."—It is a term used to express a condition when the negative and print densities are such that the cross-modulation or intermodulation product is at a minimum, and there is little or no image growth or contraction.

18.245 When is a 6000-Hz carrier frequency used for 35-mm cross-modulation tests?—When the frequency response of the recording channel is limited to 6000 Hz; however, 9000 Hz is the frequency generally used.

18.246 Describe a cross-modulation compensator and its purpose.—A cross-modulation compensator is an amplifier developed by RCA for use with a recording galvanometer when recording direct-positive variable-area sound track. It is applicable to both 16-mm and 35-mm recording. This device will permit the production of higher densities with lower cross-modulation products. This is accomplished by predistorting the audio signal in such a manner that the cross-modulation components caused by image spread because of the high sound-track density are cancelled.

Because of the inherent characteristics of photographic materials, the attainment of complete opacity and trans-

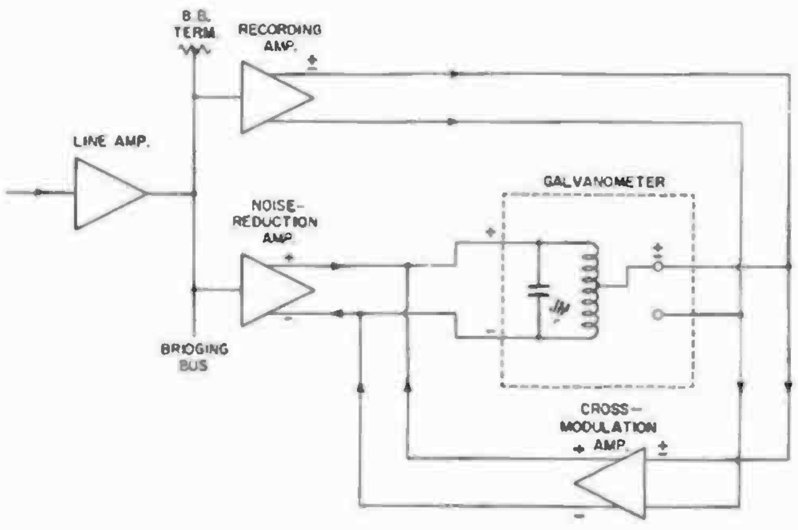


Fig. 18-246A. Block diagram for connecting a cross-modulation compensator amplifier to a recording channel for recording a direct-positive sound track.

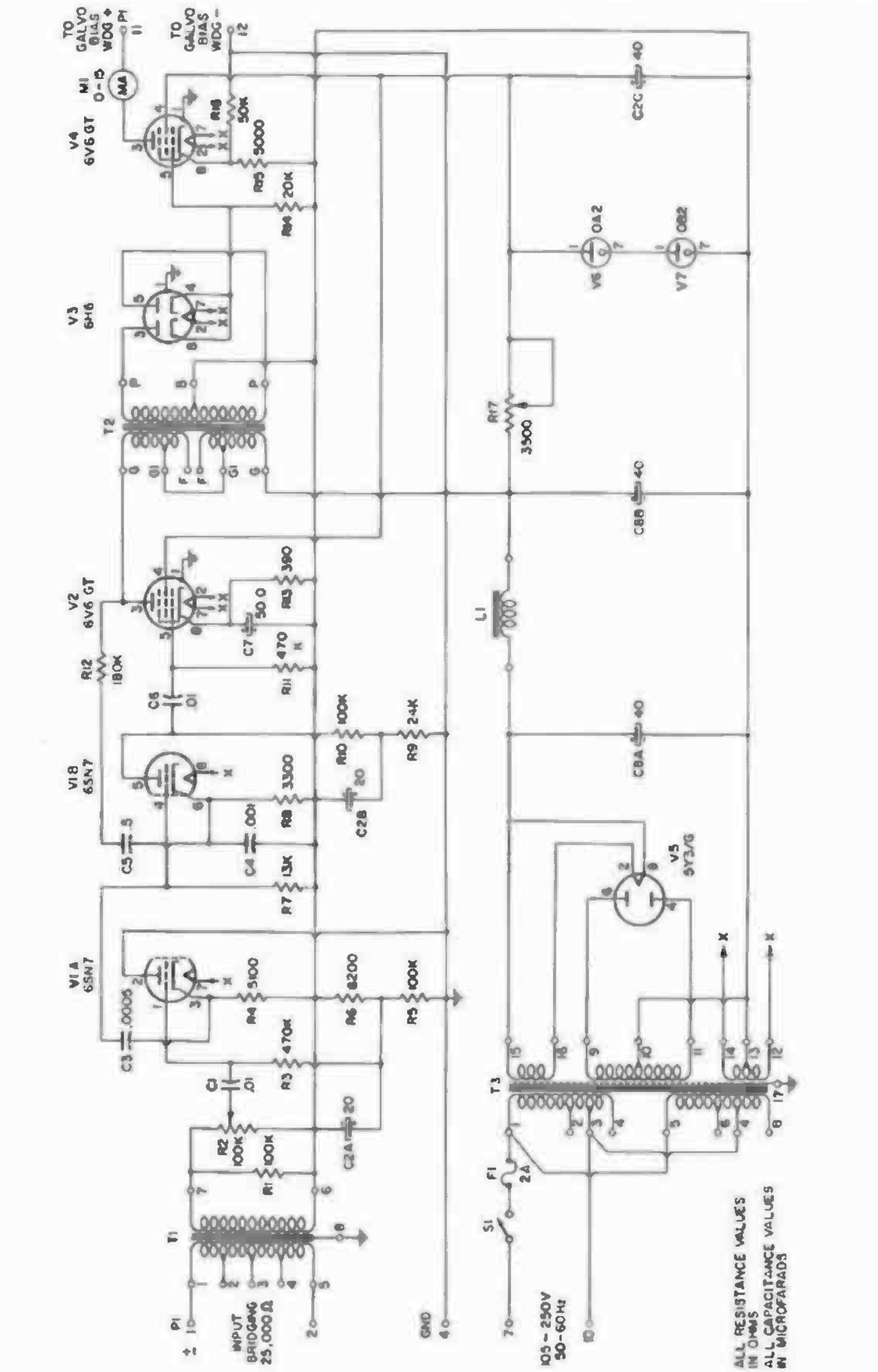


Fig. 18-246B. Schematic diagram for RCA MI-10285 cross-modulation compensator amplifier for recording direct-positive sound track.

parency is not possible. However, it is possible to produce a density contrast that is practical and will generally satisfy most requirements. The theory of cross-modulation tests is discussed in Question 18.232. By the use of the compensator, it is possible to produce direct-positive sound tracks and cancellation of 40 dB or more at a density of 1.90.

Direct-positive recording deals with a single recording consisting of a sound track on a positive print, which cannot be corrected for cross-modulation distortion, as when a negative and a print are used. This restricts the density to a low value of between 1.2 and 1.4. If the exposure is increased to obtain a higher density, excessive sibilant distortion is encountered because of the filling in of the recorded waveforms due to image spread at the higher density. The cross-modulation compensator corrects this distortion by predistorting the recorded signal. By altering the shape of the signal as it is being recorded, the shape of the optical image can be such as to counteract the distortion induced by image spread and the processing.

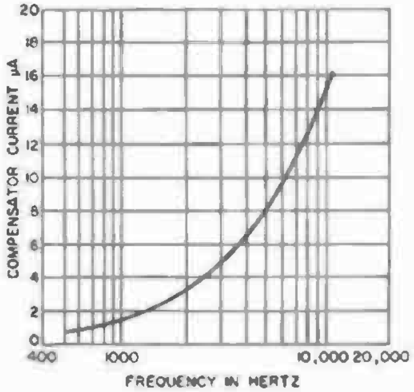


Fig. 18-246C. Frequency characteristics of RCA cross-modulation compensator amplifier for recording a direct-positive sound track.

If the original signal is passed through a filter, a compensating signal can be derived with a rising characteristic, then rectified and fed to the recording galvanometer as a bucking signal. If the phase and amplitude are properly corrected, optimum cancellation of the image spread or cross-modulation effects may be achieved.

A block diagram for the connection of a cross-modulation compensator amplifier is given in Fig. 18-246A. Except for the connection of the cross-modulation amplifier, the recording channel is normal. Cross-modulation amplifiers can be used only with a biased galvanometer and are not applicable to noise-reduction shutters. The output of the compensator unit is connected in parallel with the output of the noise-reduction amplifier (NRA) to buck the noise-reduction current. Normally, the noise-reduction current decreases with an increase of signal amplitude, while the compensating unit current increases. Therefore, the compensator current must be proportional to the signal amplitude.

A schematic diagram for a cross-modulation compensator is shown in Fig. 18-246B. The signal is applied to a bridging input of transformer T1, and

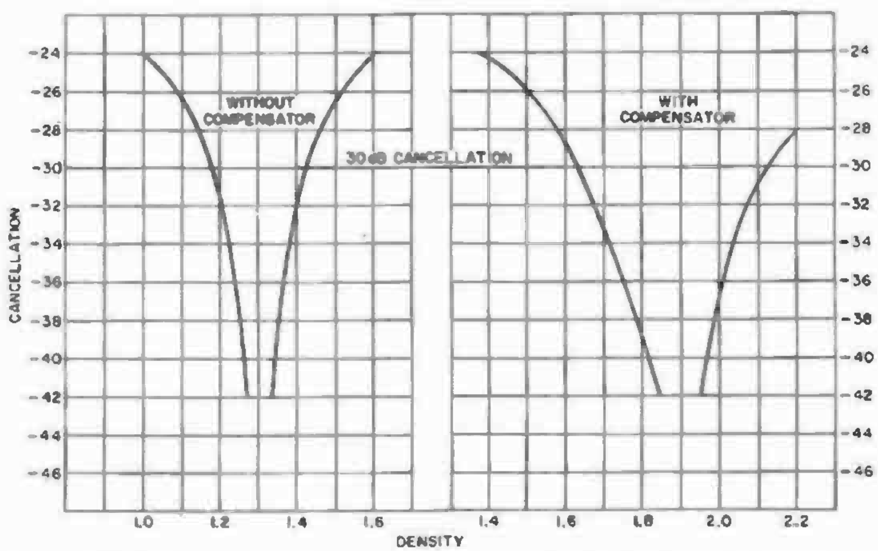


Fig. 18-246D. Cross-modulation cancellation versus density for direct-positive recording using a cross-modulation compensator amplifier.

terminated at the secondary by a gain control. The signal is then applied to the control grid of cathode follower V1A through a frequency-selective network which supplies the frequency correction. The slope of this correction is 6 dB per octave and attenuates frequencies below 500 Hz. The signal from the output of V1A is applied to the control grid of a two-stage amplifier, V1B and V2, employing negative feedback from the plate of V2 to the cathode of V1B. The feedback lowers the plate impedance of V2 and improves linearity and stabilizes the gain. Leaving V2, the signal is applied to transformer T2 and full-wave rectifier V3. The output of V3 is developed across R14 and applied, without filtering, to the control grid of V4, a dc amplifier. Because of its high output impedance V4 may be connected directly across the galvanometer bias coil without any loading effects occurring. The plate current of V4 for a zero signal is quite low due to the application of a high grid-bias voltage. Because the cathode of this tube is unbypassed, negative-current feedback is induced and the input to output linearity is maintained.

The rectifier in the compensator generates products consisting of sum and difference frequencies of both the fundamental and harmonics. Because the galvanometer does not respond to frequencies above 9000 Hz, the sum frequencies of the rectification products contribute nothing to the corrective signal. The difference frequencies of the fundamental from the rectifier products provide the corrective signal which cancels the cross-modulation components. To eliminate the effects of phase shift, filtering of the signal at the control grid of V4 has been omitted. The frequency characteristics are given in Fig. 18-246C.

The compensator is put into service by first making a series of lamp tests to determine the lamp current for the desired density, which for this discussion will be 1.60. After determining the correct lamp current, a series of cross-modulation recordings are made for different readings of the meter on the cross-modulation compensator, by keeping the percentage of modulation (galvanometer) constant and varying the input gain control on the compensator unit. The first recording is 1000 Hz for a reference frequency, then an 8000-Hz-carrier frequency for high-frequency attenuation, followed by an 8000-Hz-carrier frequency modulated 80 percent by 400 Hz. For these recordings, the galvanometer is deflected to 80 percent (down 2 dB from 100 percent).

After the film has been processed, the tests are measured in the usual manner using a cross-modulation readout unit. Tests with the greatest cancellation and minimum distortion are selected, and the compensator current noted. Assuming that the test indicates a current of 5 milliamperes, an 8000-Hz sine-wave signal is applied to the channel and the gain of the recording channel and the compensator are adjusted for an 80-percent deflection of the galvanometer. (The 8000-Hz signal is below the resonant frequency of the galvanometer.)

When adjusting the compensating amplifier and recording channel as just described above, the compensator adjustment is independent of the recording channel frequency characteristics, since any change in the characteristics is automatically reflected to the compensator unit because of its position in the recording circuits.

After the tests have been made, dialogue tests are recorded at the exact compensator current selected, then at 1 and 2 milliamperes above and below the selected value, using sibilant sentences, as given in Question 2.96. It may be found, after listening to the tests, that the compensator current should be increased or decreased slightly from the value of 5 milliamperes.

Typical cross-modulation measurements are plotted in Fig. 18-246D. When the compensator current has once been established, it will require no further attention, unless the laboratory processing methods are changed.

18.247 What are the effects of cross-modulation distortion?—The sound seems to have excessive sibilance which is not the case, but is high-frequency distortion. Cross-modulation distortion is caused by improper negative and print densities which affect the cancellation of cross-modulation distortion components. Cross-modulation distortion is more disturbing to the ear than harmonic distortion.

The proper negative and print densities may be determined by one of the two methods described in Questions 18.128 and 18.232.

18.248 What is the percent transmission when the negative and print densities for variable-area sound tracks are correct?—Including the fog density, the percentage of transmission is 50 percent. This condition is determined by the cross-modulation test.

18.249 If the exposure lamp in a film recorder is replaced, is it necessary to make a complete family of new cross-modulation tests?—Not if previous cross-modulation test data are available. An exposure lamp test is made and compared with the old and new densities for a group of given lamp currents. If the comparison is favorable, a lamp current corresponding to the old lamp is selected and recording is continued in the usual manner. If a photometer is used for monitoring the exposure lamp, the same readings for a given density are used.

18.250 How are exposure lamps for film recorders selected?—Exposure lamps for an optical film recorder must be carefully selected. Many times new lamps designed for recorder use are not entirely satisfactory. Several types of lamps are available for recording use which employ straight and curved filaments and have either bayonet or prefocused bases.

The filament should be carefully inspected as to its placement in the envelope relative to the optical system. If the filament is a curved one, the coils should be even and the arc smooth, with even spacing between the coils. If a line is drawn from the base of the lamp upward through the center of the filament, the filament should be at right angles to the line and the line should pass through the filament. If the filament is curved, the curved side is turned toward the optical system.

Most recorders are provided with an eccentric adjustment in the lamp socket for positioning the lamp horizontally and a vertical adjustment for centering the filament on the center axis of the optical system. The best way to install the lamp is to observe the filament through the optical system, if possible, by looking back into the optical system with a periscope and having the lamp burning just above a dull red color.

After installing a new lamp, it should be allowed to burn at its rated current for at least two hours before attempting to make any tests involving the recording of film. If the recorder is equipped with a photometer, the deflection for a given exposure current is noted and the lamp adjusted by the photometer rather than by current alone. After the lamp has been in operation for a few hours, it starts to blacken inside the envelope as a result of carbon deposits thrown off by the filament. This causes a reduction of intensity for a given current. The photometer indicates the light intensity and may be used for resetting the current for a given intensity, where the current alone would not be accurate.

When an exposure lamp is replaced, it is necessary to make a series of exposures at several different lamp currents, as explained in Question 18.249. The use of a photometer for monitoring the intensity of the exposure lamp will result in better control of the exposure, and eliminate the need for lamp tests when the lamp must be replaced in an emergency.

Lamps that are used for recording and reproduction are generally rated at about 75 percent of their burn-out current. A lamp rated at 10 volts, 5 amperes will burn out at about 6.5 amperes, while a lamp that is rated at 10 volts, 7.5 amperes, will burn out at about 10 amperes.

With a 10-volt, 5-amp lamp operating as rated, a color temperature of about 2950K (Kelvin) will be developed. If the lamp is operated at 6 amps, a color temperature of approximately 3200K will be developed. The color temperature of a 10-volt, 7.5-amp lamp, operating as rated, will develop a color temperature of about 2925K. With the same lamp operating at 9 amps, a color temperature of around 2975K will be developed.

Prefocussed recorder lamps employing curved filaments are manufactured in two types. It is possible for the filament to be turned 180 degrees from the normal for a given recorder lamp socket. Therefore, the filament curve must be checked in relation to the base before installation.

18.251 How are cross-modulation tests recorded on a push-pull recorder?—In a manner similar to that described for a standard recorder (Question 18.232), except only one side of the push-pull sound track is used. This is accomplished by racking over the galvanometer to one side and exposing only one-half of the track. If, after the prints are measured and excessive sibilance exists, yet the cancellation is satisfactory, the density of the exposed side of the track should be checked for nonuniformity.

If the density appears to be satisfactory, the trouble may be due to slippage in the printer. It is generally more satisfactory to measure the density of the exposed sound track before attempting to make a cross-modulation test.

Measurements of the density between the two sides of the exposure should also be made, because when the recorder is used in its normal manner (push-pull), the density variation between the two sides of the sound track must be nearly the same. If the density variation is excessive, severe distortion will result.

18.252 Why is it desirable to use push-pull sound tracks for recording rather than standard sound tracks?—Because of the greater signal-to-noise ratio, cancellation of even-order harmonics, and relieving of the necessity for blooping splices. Also, noise-reduction equipment is generally not required. (See Questions 18.209, 18.210, and 18.211.)

18.253 What is a galvanometer crossover test?—A test made on a variable-area push-pull film recorder to determine where the sound tracks are 180 degrees out of phase with respect to each other and if the azimuth adjustment is correct.

18.254 How is a crossover test made?—Thread up the recorder and proceed in the following manner:

1. Set galvanometer to the standby position (as shown at A of Fig. 18-254). Apply 1000 Hz to the galvanometer and deflect it to 10 percent-modulation (20 dB down from 100-percent modulation).
2. Galvanometer off.
3. Move galvanometer to the left far enough to cut off one sound track as shown at B in Fig. 18-254. The amount of movement is checked beforehand by observing the movement with a periscope in the optical system.
4. Record 10 feet of 1000-Hz track.
5. Galvanometer off.
6. Move galvanometer to standby position, (shown at A in Fig. 18-254).
7. Record 10 feet of 1000-Hz track.
8. Galvanometer off.
9. Move galvanometer to right until one track is cut off as shown at C in Fig. 18-254.
10. Record 10 feet of 1000-Hz track.
11. Develop negative and make one print.

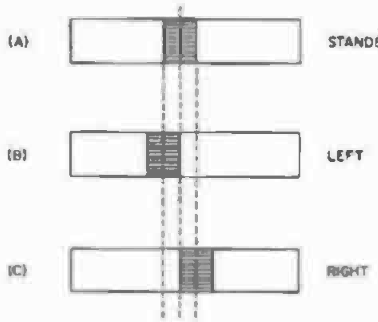


Fig. 18-254. Galvanometer crossover-test images as seen on the monitor card of a variable-area push-pull film recorder.

18.255 How is a crossover test measured?—The print is run on a normal sound head and measured by observing the output level from the three recorded positions on a VU meter in the following manner. Set the output level from the right-hand track to a convenient point on the VU meter scale, say zero. Tabulate the output readings as shown below. Assume the right-hand track reads zero, the standby position —1.5 dB, and the left-hand track —3.0 dB.

Crossover Right	*Crossover Standby*	*Crossover Left*
0	—1.5 dB	—3.0 dB

The average values of the right and left tracks are subtracted from the standby output reading. Example: right side reads 0.0 dB; left side reads —3.0 dB.

$$\frac{-3.0}{2} = -1.5 \text{ dB}$$

$$-1.5 \text{ dB} - (-1.5 \text{ dB}) = 0$$

Tolerance for the crossover test is —1.0 dB to +0.5 dB. A typical crossover test is shown in Fig. 18-255. The two sides of the track should not vary in output level more than plus or minus 0.10 dB.

18.256 Why is a lower exposure lamp current used for 16-mm than for 35-mm film?—The linear speed of 16-mm film is 40 percent slower than

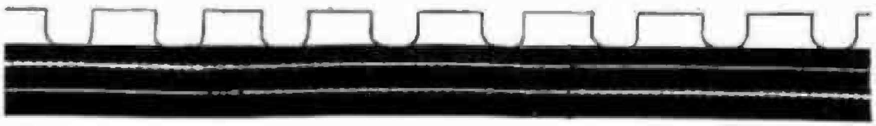


Fig. 18-255. A typical push-pull crossover test for a push-pull variable-area photographic film recorder (print).

35-mm film; therefore, for the same emulsion or a given density, the lamp current must be reduced to equal the exposure of the film running at 90 fpm.

18.257 In what direction is the filament of an exposure lamp placed?—If the filament is of the curved type, the curved side is placed facing toward the film. This is quite important as less light and possible shadows may be indicated with the filament turned in the opposite direction. In some instances, prefocused-base lamps may not fit in the exposure lamp socket with the curved side of the filament toward the film; therefore, lamps should be carefully inspected before installing them in the recorder.

18.258 What is the average negative base-fog density for variable-area sound track?—From 0.28 to 0.30, depending on the time of development. This figure includes the film-base density, and is referred to as the base-fog density.

18.259 What is the transmission through the opaque area for a print density of 1.40?—Four percent. This is the amount of light transmitted through the opaque portion of the film. (See Question 25.160.)

18.260 How are the optimum negative and print densities determined for a single-system recorder-camera?—By listening tests, as described in Question 18.128.

18.261 How is the optimum density determined for dupe negatives?—A typical example follows. Assume that an original negative is developed to a density of 2.35, and that a print made from this negative has a density of 1.40 and measures 42-dB cancellation. The negative consists of 1000 Hz for a reference frequency, 9000 Hz for high-frequency loss, and a 400/9000-Hz cross-modulation test.

This print is sent back to the laboratory and dupe negatives with densities of 1.60, 1.70, and 1.80 are made. A print having a density of 1.40 is made for each dupe negative and the cross-modulation is measured. Cross-modulation measurements indicate 40-dB cancellation for a print density of 1.60, with a high-frequency loss of 1 dB. Prints which will prove quite satisfactory may then be made at a density of 1.40.

18.262 What is an antihalation base film?—A film with a gray dye incorporated in the base to prevent halation around the image. This is extremely important for photographic film sound recording and is discussed further in Question 18.150.

18.263 What solution is used for cleaning sound track?—Several commercial solutions are available on the market. If optical sound track is to be used for rerecording, the track should be cleaned in advance of its use.

18.264 What are the average sound-track densities for Ektachrome reversal print film?—

Track	*Unmodulated density of positive print*	*Unmodulated track density of color print*
Variable-area	2.0 to 2.2	0.22 to 0.25 (clear area)
Variable-density	0.6 to 0.7	0.65 to 0.75

The sound track is printed from a positive sound track. (See Questions 18.287 and 18.315.)

18.265 What is the zero-shift method of determining negative and print densities for variable-area recording?—A method of determining the optimum densities for both negative and print by recording a high frequency (6000 Hz for 16-mm film and 9000 Hz for 35-mm film) at 80-percent modulation. A series of negatives is made using different lamp currents, with a section of unmodulated track between. The densities are then read and tabulated. Prints are made and their densities read on a special densitometer consisting of two photocells in a balanced circuit. Minimum distortion is achieved when the clear and opaque areas are equal, or when a condition of 50-percent transmission is obtained.

18.266 What other factors rather than image spread affect the high-frequency reproduction?—If overload takes place in the higher frequencies, the effect is often mistaken for excessive sibilance. This effect has been noted when the original sound tracks were recorded on magnetic tape which had been overloaded by driving the tape into saturation.

18.267 Describe the sound track image most commonly used for variable-area recording. — Present-day systems, using a light valve or a biased galvanometer, employ the double bilateral image, as shown in Fig. 18-299A. In the older systems, the single bilateral image was used.

18.268 What sound track image is most commonly used for variable-density recording?—The most commonly used image is that shown in Fig. 18-306A.

18.269 How are the negative and print densities determined for variable-density recording?—By the intermodulation distortion test method. Intermodulation tests are made in a manner similar to that described in Question 18.232, that is, by making a group of negatives and prints and then measuring the intermodulation distortion of the prints. The combination of densities having the lowest distortion is the correct combination of densities for the negative and print. In the absence of an intermodulation oscillator and analyzer, sibilant tests may be used as described for variable-area recording. (See Question 18.128.)

18.270 What is the relationship between the transmission and density of a variable-density sound track?—The negative is exposed on the straight line portion of the H and D curve where a linear relationship exists between the log of exposure and density. (See Question 18.121.)

18.271 What is the average visual density for a variable-density negative? —Approximately 0.45 to 0.60, as determined by an intermodulation test.

18.272 What is the average visual density for a variable-density print?—Approximately 0.6 to 0.8, as determined by an intermodulation test. (See Question 18.315.)

18.273 What is the effect when a variable-density sound track is overmodulated? — When a variable-density sound track is overmodulated, it is exposed along the sloping portion of the exposure curve. Thus, the distortion increases gradually. When a variable-area sound track is overloaded, the modulation peaks are clipped, as shown in Fig. 18-293. The distortion rises very rapidly.

18.274 Describe briefly an intermodulation analyzer.—Basically, the instrument consists of two separate units, a signal generator and an analyzer unit. The signal generator contains two stabilized oscillators with several fixed frequencies which may be used in combination. The second unit is the analyzer section consisting of an input attenuator, high-pass filter, amplifier, rectifier, low-pass filter, and a vacuum-tube voltmeter calibrated in percent intermodulation. When the instrument is used, the outputs of the oscillators are combined in a hybrid coil and an attenuator panel, then applied to the input of the recording channel. Sound track negatives are recorded at several densities from which prints are made at several densities. The prints are played back and the intermodulation distortion measured with the analyzer section. The percent intermodulation for the different density combinations is plotted and analyzed. The combination of negative and print density resulting in the lowest intermodulation distortion are the correct densities. (Intermodulation analyzers are described in detail in Question 22.129.)

18.275 Describe how intermodulation distortion is plotted to determine the correct negative density.—By plotting the intermodulation distortion for negative density against a given print density, as shown in Fig. 18-275. It will be noted the lowest intermodulation is obtained for a negative density of 0.45, using a print density of 0.75. The curve is rather broad in the region of the optimum densities. Under a condition of balanced densities, the intermodulation distortion can, as a rule, be held to less than 5 percent for standard track, and 3 percent for push-pull sound track.

18.276 How is intermodulation data plotted for print densisty?—It is plotted similar to Fig. 18-275, negative density versus print density (reverse of Fig. 18.275). With these two curves, the permissible variation in density above and

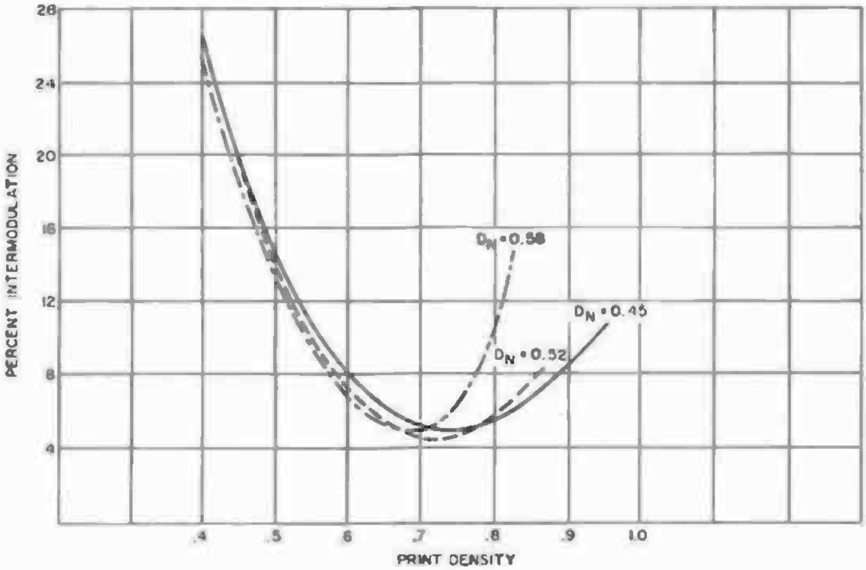


Fig. 18-275. Intermodulation tests for variable-density recording. The above plot indicates the lowest distortion is attained when the negative and print densities are 0.45 and 0.75 respectively.

below the optimum values can be readily determined.

18.277 Are intermodulation tests made with or without noise-reduction bias?—They may be made either way. However, they are generally made without bias. A slightly lower density is indicated when the measurements are made using noise-reduction bias.

18.278 At what percentage of modulation are intermodulation tests made?—At 2 dB and 10 dB below 100 percent-modulation of the light valve. A typical IM sound track is shown in Fig. 18-278.

18.279 What frequencies are used for intermodulation tests?—According to the USASI (ASA) Standard PH22.51-1961, 60 Hz and 2000 Hz.

18.280 In what ratio are the intermodulation frequencies mixed?—The accepted practice is to use a ratio of 4:1 with the higher frequency 12 dB lower in amplitude than the low frequency. This is called the SMPTE (Society of Motion Picture and Television Engineers) method.

18.281 Are intermodulation distortion measurements applicable to variable-area recording systems?—The results, as a rule, are not too satisfactory. Cross-modulation measurements are used for variable-area sound track distortion, as described in Questions 18.225 to 18.238 inclusive.

18.282 Describe additional tests that may be made with an intermodulation analyzer.—A test for spurious variations may be made by omitting the low-frequency signal and recording only the high-frequency signal. Observing the signal through the analyzer section will indicate intermodulation distortion from sources other than the nonlinearity of the exposure print-density characteristics. Any modulation of the high-frequency signal at a low-frequency rate, such as might result from poor printer contact, can also be mea-

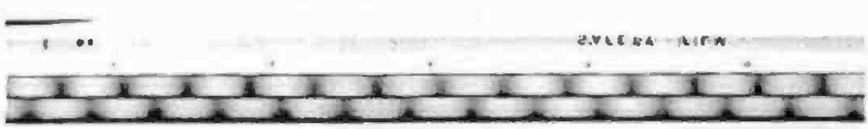


Fig. 18-278. A class-A push-pull variable-density sound track (print) with intermodulation test frequencies of 60 and 2000 Hz, mixed in a ratio of 4:1. Record level, 4 dB below 100-percent modulation.

sured with the analyzer. The reproducer used for playing back the intermodulation print tests should be tested to assure that the phototube circuit is not introducing distortion.

18.283 How are the optimum negative and print densities determined in the absence of a cross-modulation oscillator or an intermodulation analyzer?— A series of sibilant sentences are recorded; for instance, "Sister Suzie sells seashells by the seashore," or "Sister Suzie is sewing shirts for soldiers." The various print densities are listened to

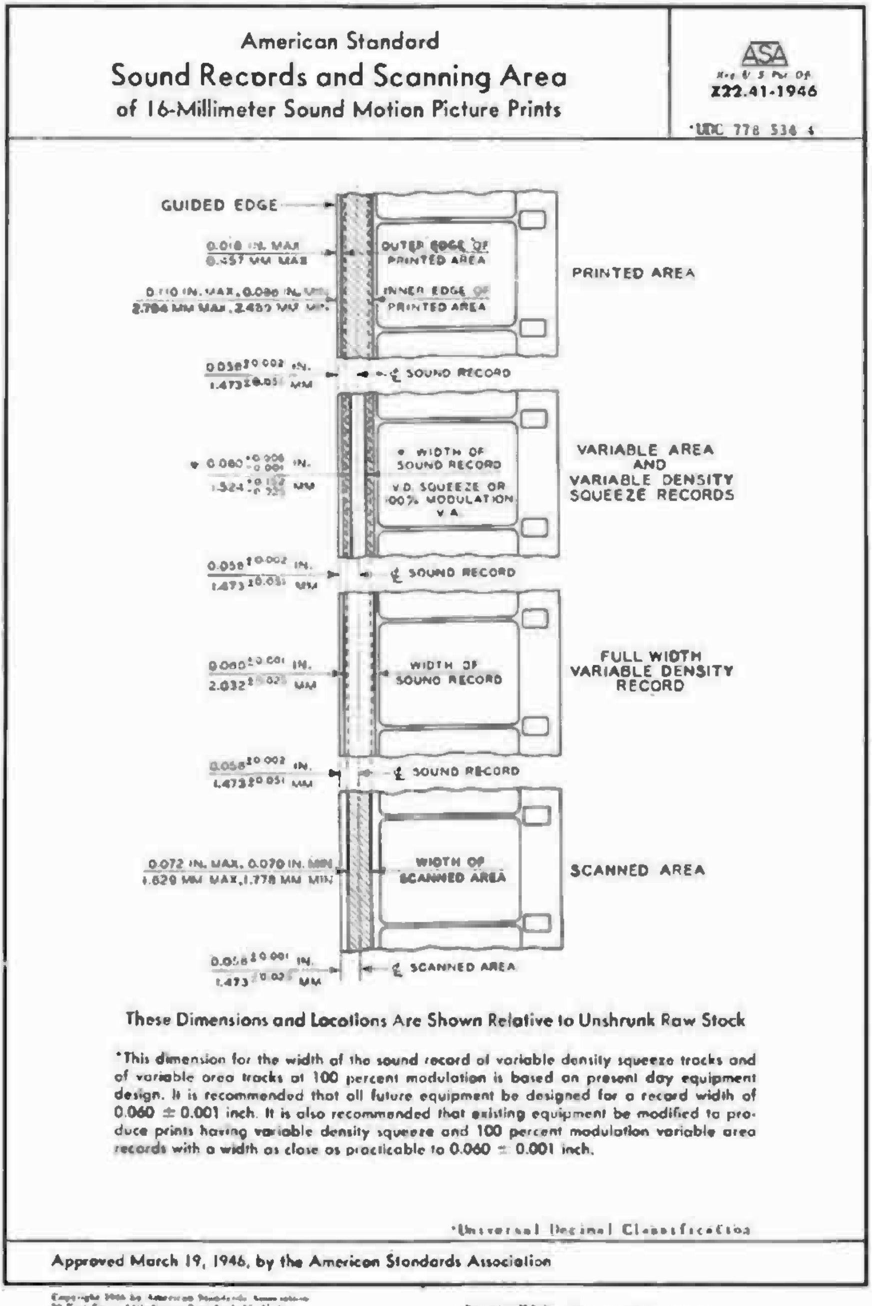


Fig. 18-284. Present USASI (ASA) PH22-41-1957 Standard for 16-mm optical film-track placement. This Standard is now under review by the SMPTE.

USA standard

Approved April 26, 1967

USAS
PH22.40–1967
Revision of
PH22.40-1957
UDC 778.534.4

Sponsor
Society of Motion Picture
and Television Engineers, Inc.

Dimensions of

Photographic Sound Record on 35mm Motion-Picture Prints

1. Scope

1.1 This standard specifies the location and dimensions of variable area and variable density sound records on 35mm motion-picture prints.

1.2 This standard specifies the area scanned in the sound reproducer.

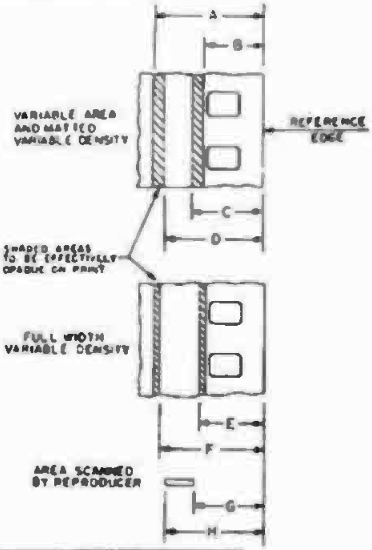
2. Dimensions

2.1 The dimensions and location of the sound record shall be as specified in the figure and table.

2.2 The sound record on the film shall be displaced from the center of the corresponding picture by a distance of 21 frames ± ½ frame in the direction of film travel during normal projection.

3. Related Standards

3.1 Prints made in conformance with this standard are intended to be used in accordance with USA Standard 35mm Photographic Sound Motion-Picture Film, Usage in Projector, PH22.3-1961.



Dimensions	Inches	Millimeters
A	0.308 nom	7.82 nom
B	0.192 nom	4.88 nom
C	0.205 ± 0.001	5.21 ± 0.03
D	0.281 ± 0.001	7.14 ± 0.03
E	0.193 $^{+0.004}_{-0.000}$	4.90 $^{+0.10}_{-0.00}$
F	0.293 $^{+0.000}_{-0.004}$	7.44 $^{+0.00}_{-0.10}$
G	0.202 ± 0.001	5.13 ± 0.03
H	0.286 ± 0.001	7.26 ± 0.03

3.2 Dimensions A and B, describing the printed area of the sound record, are established by USA Standard Dimensions of Exposed Areas for Picture and Photographic Sound on 35mm Motion-Picture Prints Made on Continuous Contact Printers, PH22.111-1965, and are shown in the table as nominal values for reference only.

Appendix

(This Appendix is not a part of USA Standard Dimensions of Photographic Sound Record on 35mm Motion-Picture Prints, PH22.40-1967, but is included to facilitate its use.)

This standard specifies that the photographic sound record will be advanced with respect to the picture by 21 frames when a composite print is produced. Consequently, when sound and corresponding picture should be synchronized for an observer close to the projected picture, or if a situation not introducing an acoustic delay at the time of projection is desired, the scanning point of the sound record must be positioned at the 21st frame ahead of the corresponding picture frame.

In the average theater, however, it is necessary to emit the sound pulses before the corresponding picture frame is positioned in the aperture. Since sound travels approximately 1100 ft per second or about 50 ft per frame during the normal projection rate of 24 frames per second, the projectionist can place the sound and picture in synchronization in the theater where he wishes by varying the length of the threading path in the projector.

For example, if the positioning of frame 21 at the scanning point brings the corresponding picture and sound to the screen and the speaker at the same instant, then positioning frame 20 at the scanning point would give synchronism at about 50 ft from the screen, 19 frames would give synchronism at 100 ft, etc.

Fig. 18-285. Standard sound-track placement dimensions for 35-mm photographic film recording.

and the combination of negative and print density with the least sibilance is then selected for use. (See Questions 18.128, 18.232, and 18.274.)

18.284 What is the standard for track placement on 16-mm optical film? —This Standard is at the present time under review by the SMPTE; however, the new Standard will be similar to the existing PH22-41-1957 (Fig. 18-284). See page 994.

18.285 What is the standard for sound-track placement on 35-mm optical film?—The USASI (ASA) Standard PH22.40-1967 is given in Fig. 18-285, and is a revision of PH22.40-1957. See page 995.

18.286 What type sound track is used with Ektachrome reversal print film?—A positive sound-track image is required. During the processing, a sulphide developer is applied to the sound-track area, resulting in a reversal sulfide track. The sound-track image may be either variable-area or variable-density. Optimum densities are determined by the cross-modulation or intermodulation tests discussed in Questions 18.232 and 18.274.

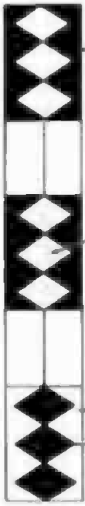


Fig. 18-287. Ektachrome print-reversal process.

18.287 Describe the reversal film process.—Ektachrome print is a reversal film. It is exposed as a negative, but after processing, it becomes a positive as shown in Fig. 18-287. Three steps take place in the reversal process. In the first step the exposed silver halides are developed to a negative silver. In the second step, the remaining unexposed halides are converted to silver sulfide. In the third and final step, the white image of step one becomes the black silver-sulfide image—a complete reversal from the original exposure. Ektachrome print film is designed to yield a nonreversal silver sound track, rather than a dye-sulfide sound track. Therefore, the track must be exposed from a negative. Such negatives may be either variable-density or variable-area. The correct density is determined by a series of intermodulation or cross-modulations tests, discussed in Questions 18.232 and 18.274.

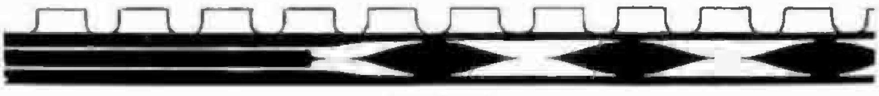
18.288 Show the appearance of a 40-Hz modulated bilateral (duplex) sound track.—Such a track is shown in Fig. 18-288. The 40-Hz modulation is shown coming from the bias lines into full modulation. This particular track is slightly modulated over 100 percent, as can be seen by the loss of the bias lines between the waveforms.

18.289 Show a 9000-Hz bilateral variable-area sound track modulated 100 percent.—See Fig. 18-289.

18.290—Show the appearance of the bias lines in a bilateral (duplex) sound track.—See Fig. 18-290. The bias lines in the original sound track were 4 mils wide.

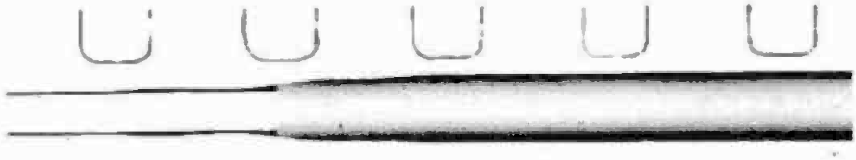
18.291 Show the effect of improperly centering a galvanometer. — The

(a) *Negative.*

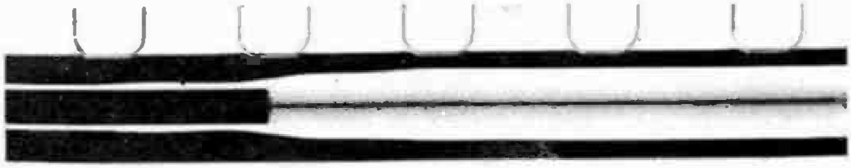


(b) *Print.*

Fig. 18-288. Variable-area, bilateral (duplex) 35-mm sound tracks with a frequency of 40 Hz.



(a) *Negative.*



(b) *Print.*

Fig. 18-289. Bias lines for a 35-mm variable-area bilateral (duplex) sound track.

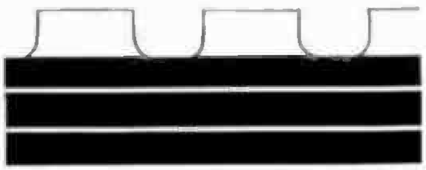
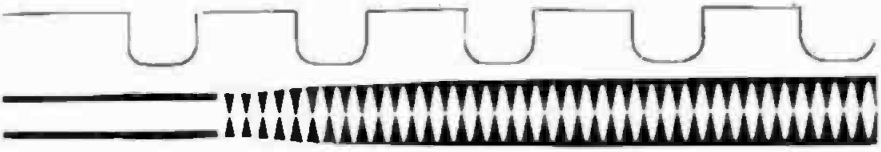


Fig. 18-290. Bias lines for a 35-mm, variable-area, duplex sound track (print).

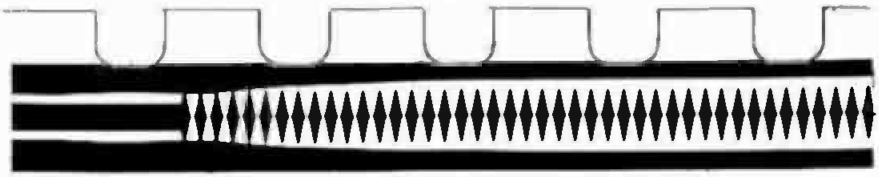
negative image in Fig. 18-291 shows that the peaks of the waveforms (black) in the center of the sound-track area are almost touching, while the valleys still have some distance to go before reaching the outer edges of the envelope. This condition is the result of not centering the galvanometer properly when it is at rest (before recording). Under these conditions, if the galvanometer is permitted to deflect to its 100-percent mark, the peaks at the center will touch and the bias line will be lost. This is the same as overmodulating (deflecting over 100 percent) with the galvanometer properly centered. This may be seen more clearly by referring to the image print in Fig. 18-291. Here, the image (black) has a separation between each modulation and no bias line between. The effect on a complex modulation is shown in Fig. 18-293 (an overmodulated sound track). The importance of properly calibrating and centering the galvanometer cannot be overstressed.

18.292 ***Show a heavily modulated bilateral (duplex) variable-area sound track.***—Such a sound track is shown in Fig. 18-292.

18.293 ***Show an overmodulated variable-area sound track.***—See Fig. 18-293. It will be noted that the peaks of the modulations are severely clipped; also, the bias line completely disappears in

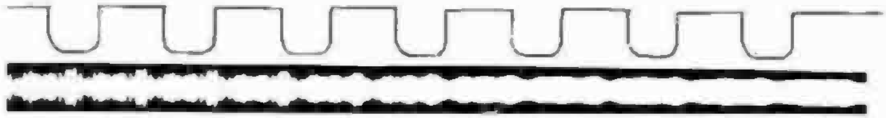


(a) *Negative.*

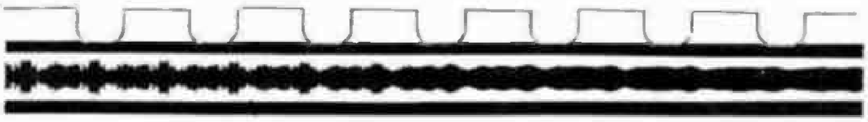


(b) *Print.*

Fig. 18-291. Galvanometer improperly centered—frequency, 1000 Hz.



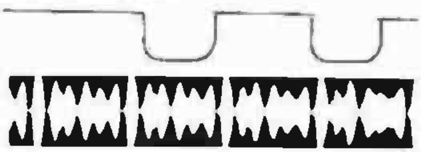
(a) *Negative.*



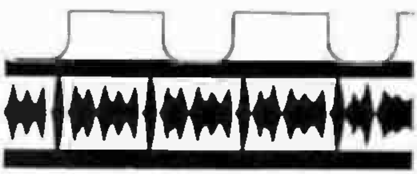
(b) *Print.*

Fig. 18-292. Heavily-modulated bilateral or duplex variable-area, 35-mm sound track.

the heaviest modulated portions. Clipping of the peaks, as shown, results in harmonic distortion. Although the modulation is high in Fig. 18-292, the peaks are not clipped.



(a) *Negative.*



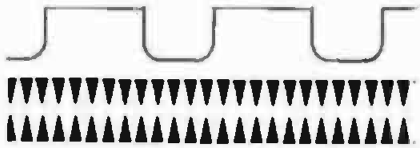
(b) *Print.*

Fig. 18-293. Variable-area, duplex, 35-mm overmodulated sound tracks showing the loss of bias lines and the clipping of the modulation peaks.

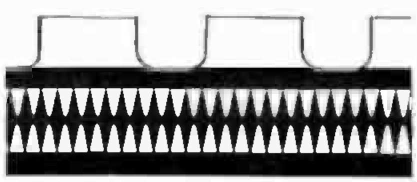
18.294 Show a 1000-Hz variable-area sound track modulated 80 percent, with too little noise-reduction margin.—See Fig. 18-294. Note that the peaks are clipped.

18.295 Show a 1000-Hz variable-area sound track with 3 dB of noise-reduction margin.—See Fig. 18-295. No clipping of the peaks is apparent.

18.296 Show a 6000-Hz variable-area sound track print which has been underexposed, normally exposed, and overexposed.—See Fig. 18-296. Print is overexposed. Negative density is correct.



(a) *Negative.*



(b) *Print.*

Fig. 18-294. Variable-area, duplex, 35-mm sound tracks, 1000 Hz, modulated 80%, with too little noise-reduction margin.

18.297 Show a 6000/400-Hz cross-modulated variable-area sound track with low-, normal-, and high-density prints.—See Fig. 18-297. It will be noted in (a) the average transmission is decreased, while in (e) it has increased. In (c) the average transmission is 50 percent or correct. The 400-Hz com-

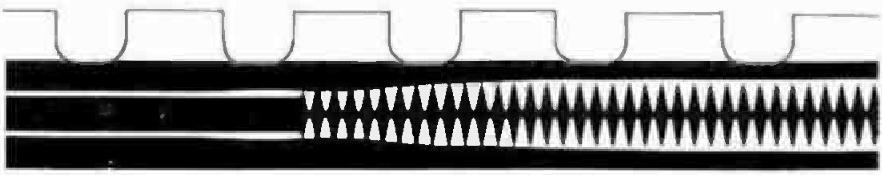
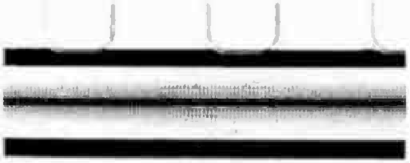
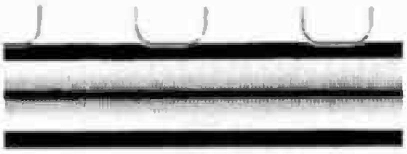


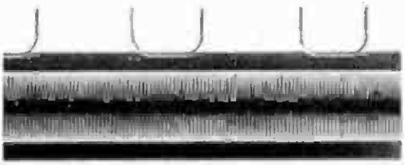
Fig. 18-295. Variable-area, duplex, 35-mm sound tracks, 1000 Hz, modulated 80% with a 3-dB noise reduction (print).



(a) *Underexposed.*



(b) *Correct exposure.*



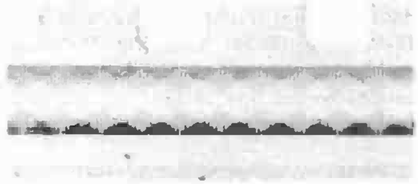
(c) *Overexposed.*

Fig. 18-296. Variable-area, duplex, 35-mm sound tracks—6000 Hz (print).

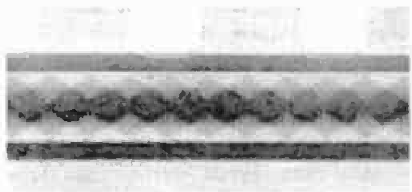
ponent in both (a) and (e) can be measured. In (c) it is at a minimum.

18.298 Show a class-B, variable-area sound-track print with modulation and bias lines.—See Fig. 18-298. In the class-B system of recording, the positive half of each cycle is recorded on one track, with the negative half on the other track. The bias lines may be seen between the modulation. The frequencies shown are dialogue.

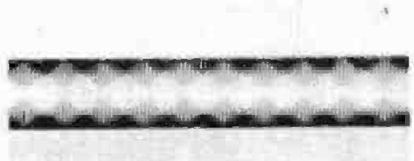
18.299 Describe a variable-area, double-bilateral sound track.—It will be noted that this sound track (Fig. 18-299A) consists of two identical bilateral tracks recorded side by side. Sound tracks of the double-bilateral type are less affected by uneven slit illumination and azimuth adjustment in a reproducer sound head. This is because the tracks are narrower than a single bilateral track. A slight increase in the signal-to-noise ratio is also obtained. A double-bilateral track must not be mistaken for a push-pull sound track, as double-bilateral tracks are in phase, while push-pull tracks are 180 degrees out of phase. Variable-area push-pull sound tracks are illustrated in Figs. 18-298 and 18-309. Push-pull variable-



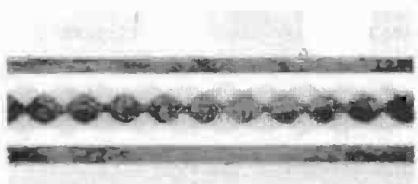
(a) *Negative (low density).*



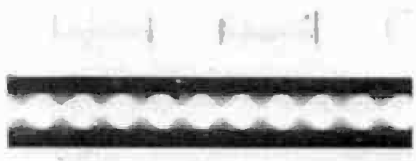
(b) *Print of (a).*



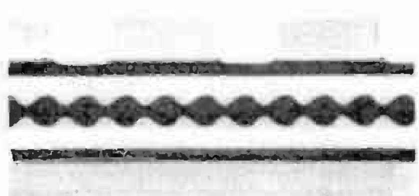
(c) *Negative (normal density).*



(d) *Print of (c).*



(e) *Negative (high density).*



(f) *Print of (e).*

Fig. 18-297. Variable-area, duplex 35-mm standard tracks — 6000/400 Hz cross-modulation test.

density sound tracks are shown in Figs. 18-278 and 18-306. (See Fig. 18-302B.)

18.300 When Is a high-density print used?—Before the advent of magnetic recording, considerable difficulty was encountered when rerecording (dubbing) because of the rather low signal-to-noise ratio. To increase the signal-to-noise ratio special high-density dubbing prints called trans-tracks (transfer tracks) were used, having a density of 2.15 or greater. Although these tracks are no longer used, their appearance is of interest. Fig. 18-300 shows a cross-modulation negative and a high- and low-density print made from this same negative. The developers used were then called P-1 and P-2.

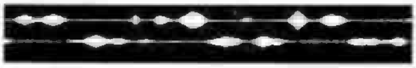


Fig. 18-298. A class-B modulated variable-area, 35-mm sound track (print).

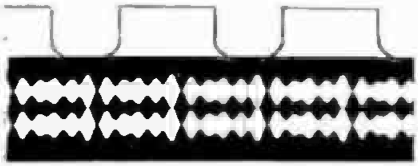
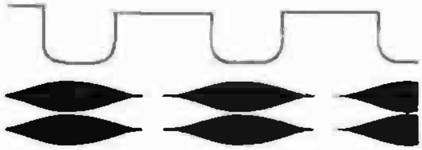
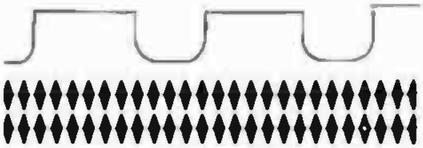


Fig. 18-299A. A double-bilateral modulated, variable-area sound track (print).

18.301 Show a 9000-Hz in-and-out-of-focus test.—See Fig. 18-301 Focus adjustments and tests are discussed in Questions 18.52 and 18.53.



(a) *60 Hz.*

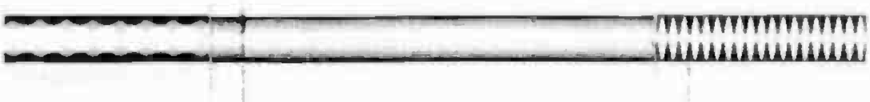


(b) *1000 Hz.*

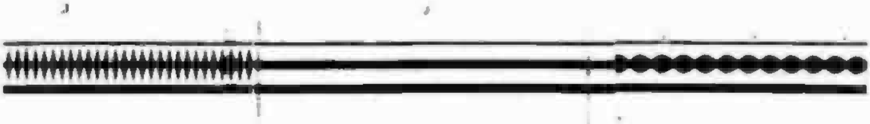
Fig. 18-299B. Double bilateral sound tracks (negative).

18.302 Show a noise-reduction amplifier opening test.—In Fig. 18-302A is shown an opening test for a bilateral (duplex) sound track using noise-reduction shutters. The test frequency is 1000 Hz. The track shown in Fig. 18-302B is for a biased galvanometer, using a double-bilateral sound track configuration. Both illustrations are negative (black) images.

18.303 Show a 1000-Hz noise-reduction closing time test.—A closing time test for a double-bilateral recorder using a biased galvanometer is given in Fig. 18-303A. The normal closing time is 4 inches of film for 90 percent-closure. The track shown in Fig. 18-



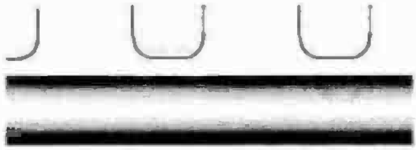
(a) *Negative, density 2.54.*



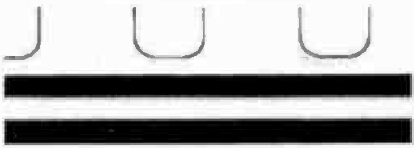
(b) *Print, developed in P-1 developer, print density 1.40.*

(c) *Print, developed in P-2 developer, print density 2.15.*

Fig. 18-300. Cross-modulation test negative with high- and low-density prints. Frequency 6000 Hz, modulated by 400 Hz.



(a) Correct focus.



(b) Out of focus.

Fig. 18-301. A 9000 Hz negative focus test.

303B is of a duplex recorder using noise-reduction shutters. The normal closing time is 3.5 inches of film, starting from the last modulation to a point where the bias line returns to its normal dimensions.

18.304 Show the appearance of a 16-mm direct-positive variable-area sound track.—Such a sound track is shown in Fig. 18-304.

18.305 Show a standard variable-density sound track.—Two standard variable-density sound tracks, one with music modulations, the other with 4000-Hz signal, are shown in Fig. 18-305.

18.306 Show a 200-mil variable-density push-pull sound track.—Such a sound track is shown in Fig. 18-306. The modulations of the two halves of the track are 180 degrees out of phase with respect to each other.

18.307 Describe a magoptical release print.—A magoptical print is a release print having a combination of magnetic and photographic sound tracks. The magnetic sound tracks are placed in their normal position on the release print, with the optical sound

Fig. 18-304. A 16-mm variable-area bilateral, black and white, direct-positive sound track.

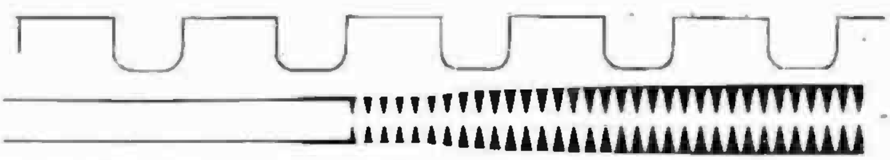


Fig. 18-302A. A 35-mm, duplex variable-area noise-reduction shutter opening test.

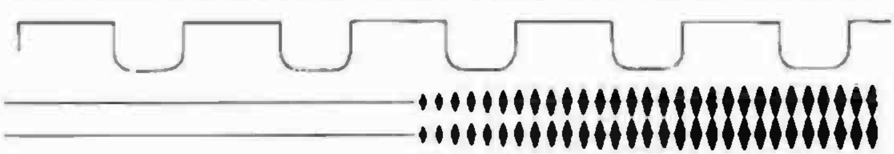
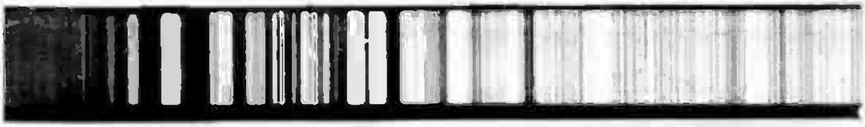


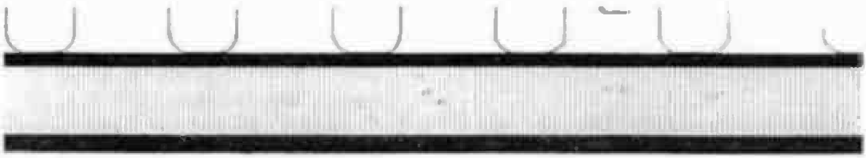
Fig. 18-302B. A 35-mm, double-bilateral, variable-area, biased galvanometer, noise-reduction opening time test.

Fig. 18-303A. A 35-mm, double-bilateral, variable-area, biased galvanometer closing time test.

Fig. 18-303B. Closing time test for a variable-area bilateral or duplex noise-reduction shutter.



(a) *Music, modulated 100%, 100 mils wide.*



(b) *4000 Hz, modulated 80%.*

Fig. 18-305. Variable-density 35-mm sound track prints.

track also in its normal position, although somewhat narrower than usual. Using this method of making release prints, the same print may be played in theaters having either magnetic or optical sound heads. For 16-mm prints, a different method is used, as explained in Questions 18.314 and 17.184.

18.308 What is a supersonic noise-reduction amplifier?—A system developed in 1948 by the Westrex Corp., for direct-positive recording, whereby a high-frequency signal was applied to the light valve ribbons along with the audio signal for variable-density recording. This signal changes the waveform in such a manner that when combined with the toe characteristic of the film, the waveform would result in a linear relationship between the audio signal in the light valve and the transmission characteristic of the film. This method accomplished in somewhat the same manner, results which the cross-modulation compensator, manufactured by RCA, achieved with variable-area film.

A frequency of 24,000 Hz is applied to the ribbons of a biplanar light valve. The valve ribbons are continuously modulated to an amplitude of 200 percent. This high-frequency bias signal is applied in conjunction with the normal audio signal.

An elementary block diagram for a supersonic bias recording system is shown in Fig. 18-308. The light valve is driven from its normal amplifier. The high-frequency bias is induced in series with the center leg of the light valve. It will be noted particularly that a noise-reduction amplifier is not used with this system; only the supersonic bias oscillator is used.

18.309 Show the appearance of a 200-mil variable-area direct-positive push-pull sound track.—The images shown in Fig. 18-309 were made using a

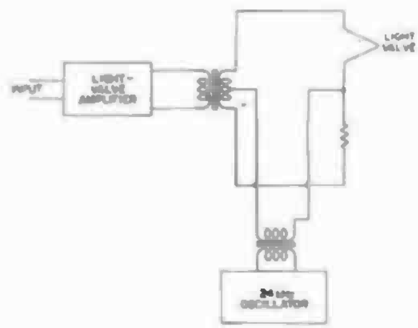


Fig. 18-308. Supersonic bias system used by Westrex for direct-positive variable-density recording.

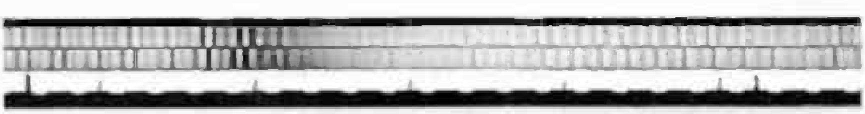


Fig. 18-306. A 35-mm variable-density 200 mil, push-pull sound track (print). Modulated 100%.

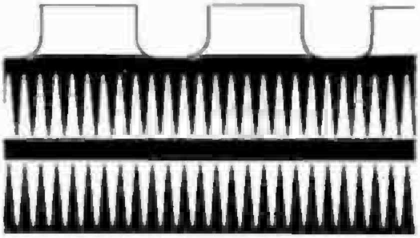


Fig. 18-309. A 200 mil, variable-area, direct-positive, push-pull sound track.

light valve. It will be noted that the peaks (white) of the modulation are 180 degrees out of phase with respect to each other.

18.310 What is latent image keeping and how are such tests made?—The term "latent image keeping" is the change in the exposure of the film between the time of exposure and the time of development. No definite rules can be applied as to the handling of the film between the time of exposure and development, except that it should be handled in the manner prescribed for such film. Because of the many variable factors involved, any one may cause the exposure to increase or decrease. Factors affecting the exposure are the type of emulsion, degree of exposure, temperature, humidity, and the time lapse between exposure and processing.

Latent image keeping tests are made by taking a series of lamp tests exposed at different time intervals and developing them all at the same time. The density of each test is then plotted as time versus density. It is of extreme importance when making such tests, that the exposure lamp be set to exactly the same intensity and that the same emulsion be used for all tests. Suggested time intervals are 36, 24, 12, 8, and 6 hours, exposed in that order.

18.311 Give a cross-sectional view of color film construction.—A cross-sectional view of an Ektrachrome color print is given in Fig. 18-311. It can be observed that the stock has three layers, which consists of blue, green, and red. (See Question 18.287.)

18.312 What Is the cause of density drift in a recorder, although the photometer readings are normal?—This is generally due to the improper air circulation around the exposure lamp and the recording optical system. This can be overcome by the use of a small fan, such as a Muffin fan, mounted to pull the air out of the exposure lamp compartment, rather than blow the air in. If the recorder case will permit, a square hole for the fan is cut in the rear of the case and the fan is mounted to draw out the air. The fan should be connected by a relay across the dc exciter lamp voltage so that the fan starts when the exciter lamp is energized.

18.313 What Is a trans-track?—A special variable-area high-density print used for rerecording before the advent of magnetic recording. Now obsolete.

18.314 Describe a magoptical release print using an infrared exciter lamp for recording the optical sound track.—Magoptical sound prints can be made which utilize a discovery of George Lewin in 1955. He covered the optical sound track with a full-width magnetic stripe. The procedure for making such prints is: After the print has been processed, the sound track, which may be either variable-area or variable-density, is fully covered with a magnetic stripe, and the sound is recorded in the usual manner. To reproduce the optical sound track, an infrared exciter lamp is used, rather than the white light, because the magnetic stripe is *transparent* to the infrared light. A germanium photodiode, which is sensitive to infrared light, is used to reproduce the sound from the optical track. In this system, either one of the sound tracks may be reproduced individually or simultaneously. The reader is referred to the references. (See Questions 18.307 and 17.184.)

18.315 Describe a simple device for marking takes while looping.—A simple device that may be used for marking takes while looping dialogue is shown in Fig. 18-337. It consists of a 500-Hz

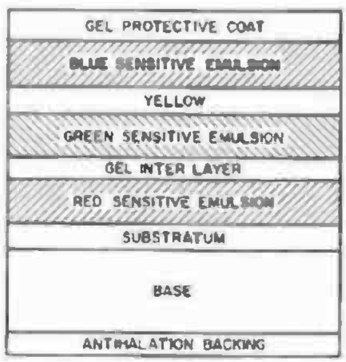


Fig. 18-311. Color layers of Ektachrome reversal print film.

high-frequency buzzer and a repeat coil. A 10,000-ohm resistor is connected in series with one side of the coil to provide a high-impedance output which may be bridged across a bridging bus, or any other convenient part of the recording system. When the system is up to speed, a single tone is recorded ahead of each take. If the take is satisfactory, two tones are recorded at the end of the take. Later, these tones are used by the editorial department for identifying the selected takes.

18.316 Can alternating current be used on a recorder exposure lamp?—Yes, but it may induce a 60-Hz modnlation on the sound track. This is possible if the lamp filament is of a light construction. Alternating current should not be used if a dc supply is available. Exciter lamps in 16-mm projectors use ac, but of a high freqnency generated by an oscillator. This subject is discussed in Question 19.166.

18.317 What are the characteristics of a class-AB recorder?—The recording mask is so shaped that, at low percentages of modulation, the system records as a class-A recorder. Above a given level, the system acts as a class-B recorder. This system was used a number of years ago for variable-area recording in some types of newsreel cameras. A noise reduction system is not required. Now obsolete.

18.318 Do push-pull recording systems require a noise-reduction system?—Push-pull recording systems such as the variable-area class-AB and class-B are inherently noiseless recording systems, requiring neither bias-current shutters or a noise-reduction amplifier. However, the variable-area class-A push-pull recording system does require a noise-reduction system, somewhat similar to that of the duplex system.

The Westrex Corp. three-ribbon light valve described in Question 18.15 also requires a noise reduction-amplifier.

18.319 How long may exposed film be held before development?—It should not be held longer than 24 hours. An exposure test must be included for use by the processing laboratory. This test is developed before processing the sound track to assure a correct development time.

18.320 What is a step tablet?—It is a piece of film with accurately graduated exposures which range from a diffused density of 0.05 to 3.00 in 21 steps, and is used for the calibration of densitometers. Each tablet is individually calibrated.

18.321 Can negative sound tracks be used for reproduction?—Yes. However, they are likely to be quite distorted and noisy because of the low signal-to-noise ratio and image spread. The sound track on a negative is black, with a large amount of the sound-track area clear, which reduces the signal-to-noise ratio. Because this is a negative, the action of the noise-reduction system, insofar as the reproduction is concerned, is in reverse.

18.322 Why is a positive sound track used for release prints?—The electrical output from a photocell when used with optically recorded sound tracks is dependent on the amount of light falling on the photocell anode and the frequency of interruption of that light.

When a positive sound track is played back, the greatest output signal from the photocell is obtained as the clear areas pass before the scanning beam. During periods of low or no modulation, the signal-to-noise ratio of the print is high because the areas around the sound track image are opaque, due to the noise-reduction action during the recording of the original negative.

18.323 Why is a negative sound track used for adjusting and measuring the cancellation of a push-pull sound head?—Prints are not used because of printer slippage and variations in the processing which will cause variations in the adjustment of the cancellation.

A typical example is a sound head which, when measured with a negative sonnd track, indicated a cancellation of 40 dB. Using a print made from the same negative, the cancellation measured 14 dB. (See Questions 19.97, 19.98, and 19.99.)

18.324 Show the construction of a laboratory motion picture printer.— In Fig. 18-324 is shown a Bell and Howell Model-C additive color printer. This machine is capable of printing 16-, 35-, 35/32-, and 65/70-mm picture with sound track, running at either 60 or 180 fpm. The operation may be controlled by the use of punched tape.

18.325 What is a wild shot?—A sound track made without the benefit

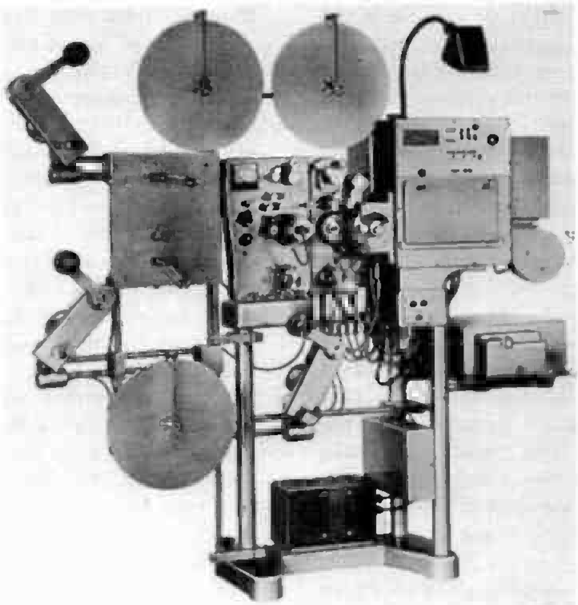


Fig. 18-324. Bell and Howell Model-C additive color printer.

of synchronous driving system between the camera and the sound, or the camera is driven alone at an increased or decreased frames-per-second speed.

When making running shots, such as a running horse or fast-moving vehicle, the camera is mounted on a camera car and the camera motor driven from batteries. The frames per second are set using a tachometer. Very often, a wild shot involving sound is made using a ¼-inch tape recorder which is battery driven. (See Question 3.72.)

18.326 What is printer distortion?—Distortion of the sound track created by flutter and slippage in the film trausport system of the printer mechanism.

18.327 What effect does printer flutter have on the sound track reproduction?—The same effect as a sound head with flutter. Piano music takes on the characteristics of a harp, resulting in distorted reproduction.

The flutter of a printer in good condition will generally be on the order of 0.12 percent or less. Excessive flutter also affects the printing of cross-modulation tests. (See Question 18.242.)

18.328 How may the flutter of a printer be measured?—By recording a 3000-Hz negative on a recorder known to have low flutter, and then making a print from this negative on the printer in question. The flutter of the print and negative is then measured on a sound head of low flutter. The measured difference in flutter between the negative and print is an indication of the printer flutter. (Flutter bridges are discussed in Question 22.41, and flutter measurements are discussed in Question 23.149.)

18.329 What is an exposure test?—It is a full-width exposure of the sound track without modulation, put on the end of each roll of recorded sound track. It is generally about 20 feet long, and it is the exact same exposure as the sound track. It is torn off by the laboratory and run through first to determine the time of development for the required density. At the end of the program material, a punch mark is put in the film to assist the laboratory in finding the inner end. In a variable-area recording system, the exposure test is made by tilting the galvanometer so that the light image covers the full slit. For a light valve, the bias is removed. For a direct positive, the procedure is reversed; that is, the bias is closed down to near the closure point.

When a biased galvanometer is recording a standard track, it is tilted to completely illuminate the slit. For a direct-positive track, the procedure is again reversed; that is, the galvanometer is tilted until the slit is completely dark. The importance of these tests cannot be overemphasized, as it is the only means the laboratory has for determining the correct density. It is the policy in some studios to first record a spot-check cross-modulation test for 20 feet at the end of the program on each roll, and then use the exposure test. The cross-modulation test is then run along with the dailies, and is then measured. This test can be used as a quality control. For the light-valve variable-density recording, the procedure is the same.

18.330 Describe the facilities required in a transfer channel for magnetic tape, film, and photographic sound track.—Generally, one of the busiest sections of a sound department is the one that deals with transfer channels. It is here that the production sound track from the shooting of the previous day is transferred, along with music and sound effects.

If the studio employs ¼-inch sync-pulse magnetic tape recorders or 17.5-mm magnetic film for production recording, these tracks must be transferred to 35-mm magnetic film for editorial purposes. After a rerecording session, the master sound track must be transferred to a photographic sound-track negative for printing on the release print, or magnetic submasters must be made for transferring to release prints, using magnetic sound tracks. (See Questions 17.179 and 17.202.) The ¼-inch sync-pulse equipment must also be capable of transferring recorded tapes, using Rangertone, Ryder Echelon, and Nagra sync-pulses, and equipped to record a sync pulse on a tape when required.

The equipment of a transfer channel must be quite flexible and capable of being patched up quickly. When necessary, it must be usable as back-up equipment for other parts of the department. A typical transfer channel in one of the Canadian motion picture studios, consisting of eight racks of equipment, is shown in Figs. 18-330A and B.

In rack 1 at the far end are mounted: film-loss equalizers, a machine-and-monitoring-selection panel, high- and low-pass filters, cross-modulation readout panel, and a small patch panel. (See Question 18.234.) Rack 2 houses two line amplifiers, limiter-compressor amplifier and balancing panel, patch panel, attenuators, and two monitor amplifiers. Rack 3 contains an RCA ¼-inch sync-pulse recorder-reproducer, with a sync-pulse synchronizer and resolver used for the transfer of daily production sound track. A magoptical machine is shown in rack 4. This machine will reproduce 17.5-mm magnetic film at 45 fpm and 35-mm film at 90 fpm, as well as standard or push-pull optical sound track. It can also be used to record either 17.5- or 35-mm magnetic film.

A three-track magnetic recorder-reproducer is shown in rack 5. This machine is used to transfer dubbing masters to photographic sound track. At the extreme right in rack 6 is a 16-mm

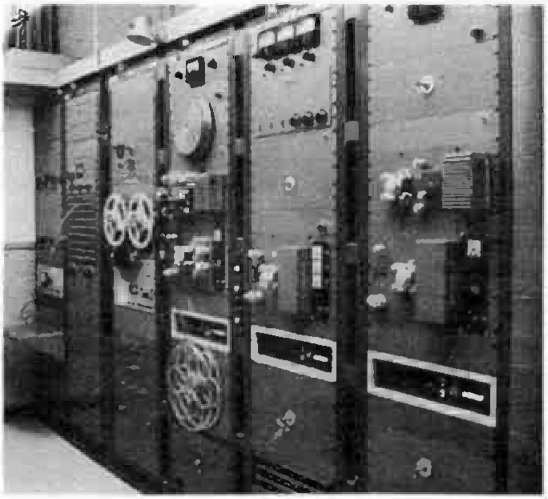


Fig. 18-330A. Recording and reproducing equipment in a transfer channel. Transfers may be made to magnetic or photographic film.

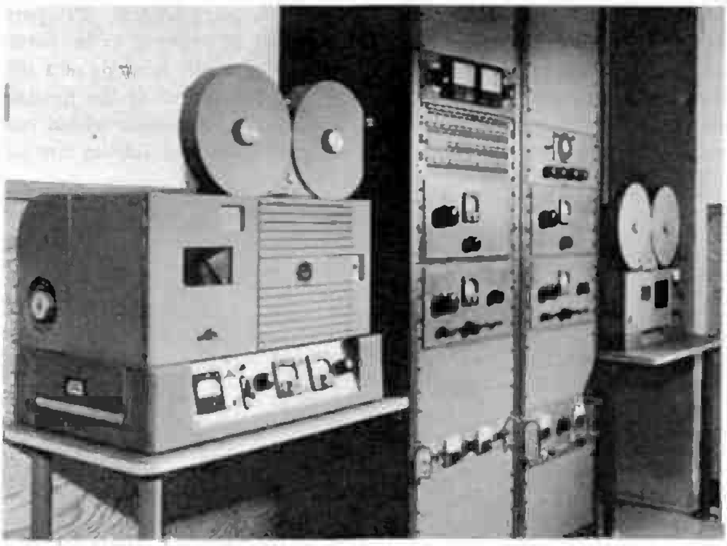


Fig. 18-330B. Photographic film recorders and control equipment.

magoptical machine capable of recording and reproducing 16-mm magnetic film running at 36 fpm. This machine is also capable of reproducing 16-mm optical sound track. All the machines shown and those in the machine room and the projection booth are capable of being selsyn interlocked, when necessary, with the transfer room.

On the other side of the room (Fig. 18-330B) are two RCA photographic film recorders and two racks of control equipment. These two racks, 7 and 8, contain two RCA recording and noise-reduction amplifiers, and two Kepco constant-voltage power supplies for the recorder exposure lamps. Included also are two VU meters for indicating the magnetic and optical bus levels, patch panel, and a cross-modulation oscillator. Turntables and nonsync tape machines (not shown) are included as auxiliary equipment.

The motors of all machines are dual-purpose; that is, they may be operated selsyn interlock or as a straight synchronous motor. Two selsyn buses are provided at each machine, with the distributors being remotely controlled by a 28-Vdc control system near the center of the racks. Intercommunication is provided throughout the department.

All magnetic reproducers are supplied high and low voltage from a group of Kepco constant-voltage power supplies in a power room some 75 feet. away. Interconnecting wiring and wiring to other parts of the department are carried by a 6-cell metal gutter above the racks. Such a gutter is discussed in Question 24.49. The block diagram for this installation is given and discussed in Question 18.331.

18.331 Give the block diagram for the transfer channel discussed in Question 18.330.—Starting from the upper left in Fig. 18-331A, and proceeding downward are shown racks, 5, 4, 6, and 3. At rack 5 is a three-track 35-mm recorder-reproducer used for transferring the dubbing-master sound tracks to photographic sound track. The output at the top is taken from a combining network which combines the three sound tracks into a composite sound track, and is fed to a machine monitor switch S1. This switch is used to select the monitor circuits of the several machines. The signal from S1 passes through monitor gain control P1, then to switch S2 and to a monitor amplifier and loudspeaker. In the direct position of switch S2, monitoring is taken directly from the magnetic recording bridging bus, through bridging transformer 1, gain control P2, then to the monitor amplifier. The individual sound tracks from the machine are brought out separately to jacks in the patch panel.

Switch S3 selects the signal to be recorded and applies it to the channel gain control P3, and from this control to the input of line amplifier 1, which drives the magnetic bridging bus. From the bus, the signal is taken through bridging transformer 2, an attenuator, and to the input of a limiting amplifier. The limiting amplifier is so adjusted that 1 dB of limiting is obtained for 90-percent deflection of the recording galvanometer on the photographic recorders. Ceiling control P4 permits the amount of limiting to be varied over a range of 4 dB, in steps of 1 dB.

At the output of the limiting amplifier is an attenuator which reduces the signal level and applies the signal to a

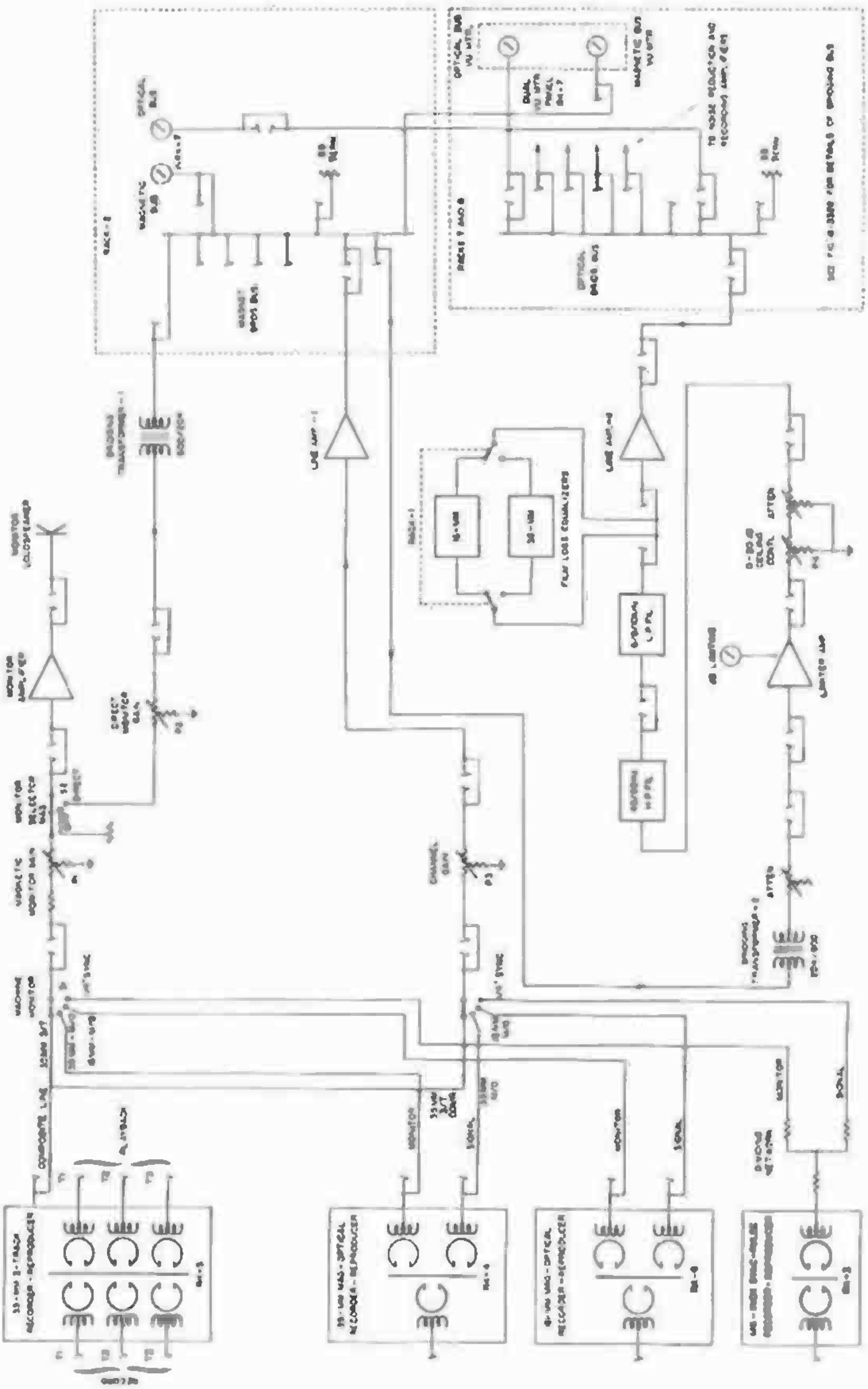


Fig. 18-331A. Block diagram for the transfer channel discussed in Question 18.331.

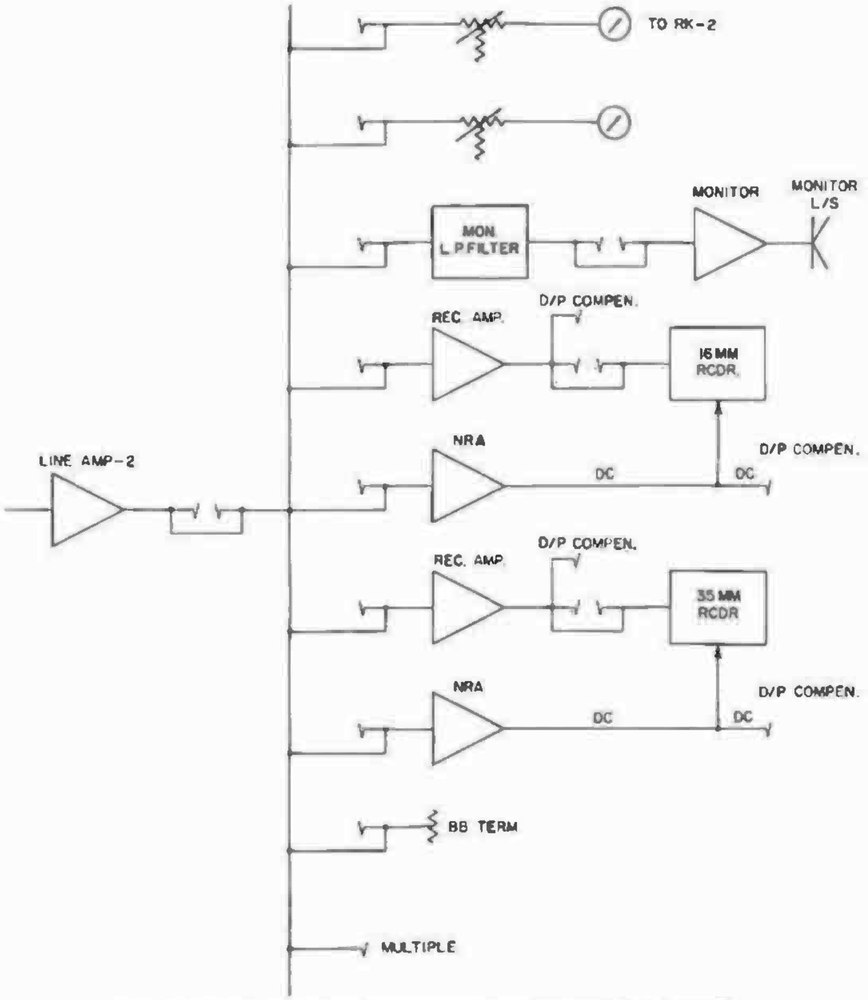


Fig. 18-331B. Bridging-bus connections for transfer channel.

high- and low-pass filter, film loss equalizer, line amplifier, and to the optical recording bridging bus. Across the bus are fed the noise reduction and recording amplifiers for the optical recorders, also two VU meters that appear in racks 1 and 7 (Fig. 18-330B). Two VU meters are also connected across the magnetic bridging bus and appear in the same racks. A meter for indicating the amount of dB limiting (not shown) also is mounted in rack 1. The details of the optical bridging bus are shown in Fig. 18-331B.

For transfer of 16-mm magnetic sound track to an optical sound track, an 80-Hz high-pass filter and a 6000-Hz low-pass filter are used. For 35-mm transfers, the 45-Hz high-pass and 8000 Hz low-pass filters are employed. The combination of equipment permits any type transfer to be made, either magnetic or photographic. The same output lines are used for the magoptical machines; the type reproduction is selected at the machines.

Only the basic design has been given for this installation. Operating levels have been purposely left out, as each installation is set up to meet local conditions and equipment. The bridging bus with its various pieces of equipment is shown in Fig. 18-331B.

The frequency response of such an installation in the flat position should be within plus or minus 1 dB, at 30 to 12,000 Hz, and the optical recorders should be well within the manufacturer's specifications. The monitoring system should have wide-frequency characteristics with low distortion. As a rule, the transfer channel does not attempt to make any corrections to an existing sound track; its purpose is to make, as nearly as is possible, a one-to-one copy of the original sound track.

The recording and reproducing levels must be checked before any transfers are made, and if the material carries a head tone, the level is individually adjusted for each sound track to be transferred. The signal-to-noise ratio, frequency response, and distortion must be checked daily. Equipment for such installations is discused throughout the various sections.

18.332 What is an editorial sync mark?—It is a sync mark used by the editorial department to synchronize the sound tracks with the picture for rerecording.

Because each sound track is separate from the picture during rerecording and the rerecording sound heads and projector are interlocked electrically, the machines all start from the same point. Therefore, a start mark is placed on the sound track so that the sound starts with the action. This is illustrated in Fig. 18-332.

After the picture has been rerecorded, the master sound track is transferred to an optical negative and synchronized with the picture for printing. The sound is then moved ahead by the prescribed amount for the particular film in use.

18.333 What is a mercer clip?—A plastic or metal clip the size of one picture frame and having four fingers at each end that may be slipped into the sprocket holes. It is used by film editors

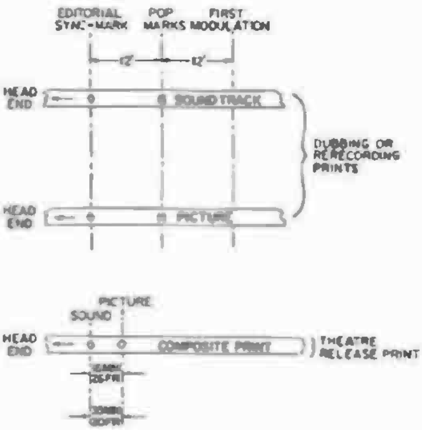


Fig. 18-332. Comparison of editorial sync marks and that used on a composite release print.

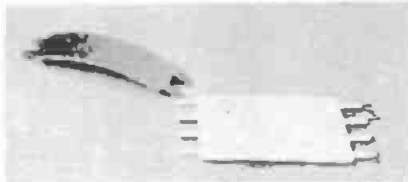


Fig. 18-333. Mercer clips for making temporary splices in a film-processing laboratory.

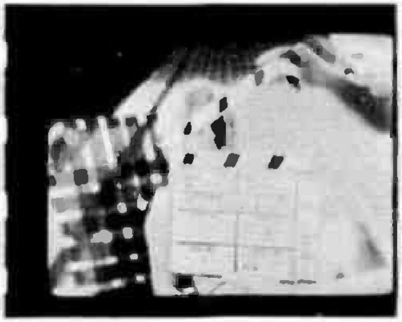
and laboratories to mark a section of picture or sound track to be printed. The clips are placed at the start and at the end of the desired section. A clip of this type is shown in Fig. 18-337.

18.334 What are clap sticks and how are they used?—Two flat sticks hinged at one end, with a camera scene slate attached. When the camera and sound are up to speed, the assistant cameraman holds the slate before the camera to photograph the scene number. If it is a sound shot, the assistant cameraman announces the production, scene number, and take number, then claps the sticks together making a loud sharp sound. This sound is rerecorded in sync with the picture and is later used by the editorial department to synchronize the picture and sound track by matching them in an editing machine.

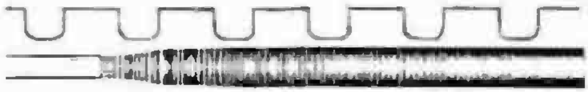
A single frame of film is shown in Fig. 18-334 with the clap sticks in front of the camera. The appearance of the recorded waveform is shown at part (b). Because of its steep wavefront, it can be easily identified by sight or sound. In some types of clap sticks, a microswitch is incorporated which actuates a buzzer whose sound is recorded rather than the sound of the sticks. The waveform of the buzzer is shown at part (c), and like the sticks, has a steep wavefront.

18.335 What is a pop mark?—A cue signal cut into the head end of a sound track by the editorial department to warn the rerecording mixer that the first modulations are about to start. For optical sound track, several scratches are made across the track in the distance of one frame. When these scratches pass the phototube of the reproducer, they cause a popping sound because of the steep waveform of the scratches.

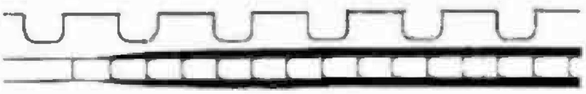
For magnetic film, a single frame containing 400 or 1000 Hz is cut into the



(a) Picture.



(b) Clap-sticks sound track.



(c) Buzzer.

Fig. 18-334. Clap sticks, and clap-stick modulations.

track or recorded by a hand device that impresses a saturated signal which also results in a distinctive sound.

18.336 What are the gear ratios normally used for synchronous film recorders?—For portable recorders using 16-mm magnetic film and running at a linear speed of 36 fpm, the motor turns at 1800 rpm. This speed is reduced through a 25:1 gear reduction to a 20-tooth film sprocket turning at 72 rpm. The motor is on the order of 1/75 hp and of the permanent split-capacitor type, with a torque rating of approximately 3 inch/pounds.

For 35-mm film, running at a linear speed of 90 fpm, the gear reduction is 10:1, using a 32-tooth sprocket turning at 180 rpm.

Studio recording equipment generally employ dual-purpose motors designed to run at 1200 rpm, so that they may be used as both synchronous motors or selsyn interlock. Various types of motors are discussed in Section 3.

18.337 Define the terms "Sepmag," "Commag," and "Compt."—These are terms used in Europe to designate different type of film operation.

Sepmag—Sound on a separate magnetic film.

Commag—Magnetic stripe on picture base.

Compt—Photographic sound track on 16-mm film base.

18.338 Describe the basic principles of a variable-area optical system for sound recording.—Such an optical system is shown in Fig. 18-338A and is similar to that used in the RCA PR-31 film recorder, discussed in Question 18.26. At the upper center is an exposure lamp filament A directed to a condenser lens B. In front of this lens is placed an image or optical mask (See Question 18.77). The light leaving the mask strikes an objective lens E where it is focused on the recording galvanometer mirror F. The light is reflected from this mirror to another condenser lens G. On the opposite side is placed a mechanical slit H which forms the scanning beam I at the film N. A sharp image of the slit is obtained by an objective lens J near the film. A small portion of the light reflected from the galvanometer mirror F is reflected on a curved mirror K thence to a flat mirror L and from there to a monitor card M at the top of the schematic. The reflected image of the galvanometer mirror permits the recordist to observe the percentage modulation of the galvanometer while recording. Various calibrations are provided on the monitor card to show the basic setting, margin,

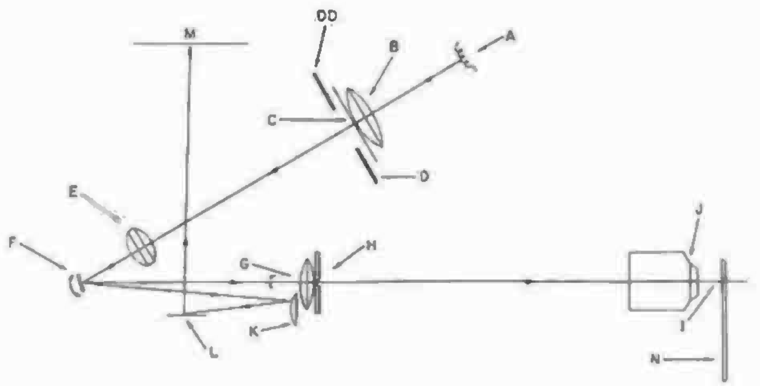


Fig. 18-338A. An optical system for recording variable-area sound tracks similar to that used by RCA.

and the operation of the noise reduction system.

A schematic diagram of a Westrex Corp. optical system used in their RA-1 1231 photographic recorder is given in Fig. 18-338B, and is of interest since a large number of these recorders are still in service. This recorder employs an RA-1438 modulator unit (Fig. 18-338C), with the optical train given in Fig. 18-338B, which may be used for recording either negative or direct positive sound tracks.

Referring to Fig. 18-338B, in the negative sound track position the exposure lamp beam A is projected through a light tunnel B which turns the light 90 degrees and directs it through an objective lens C then through the light valve ribbons D. The light then passes through a light valve objective lens E and through a 45-degree angle mirror F which is used only when the exposure lamp is in the direct-positive position. Leaving mirror F the light is reflected at a 45-degree angle by mirror G to another objective lens J at the film K on the surface of the recording drum L.

When the exposure lamp is changed to the direct-positive position (downward) the light passes through condenser lens H, to the slit mirror F and is reflected upward through the light valve objective lens E onto the light valve ribbon surfaces at D and back again through the objective lens and

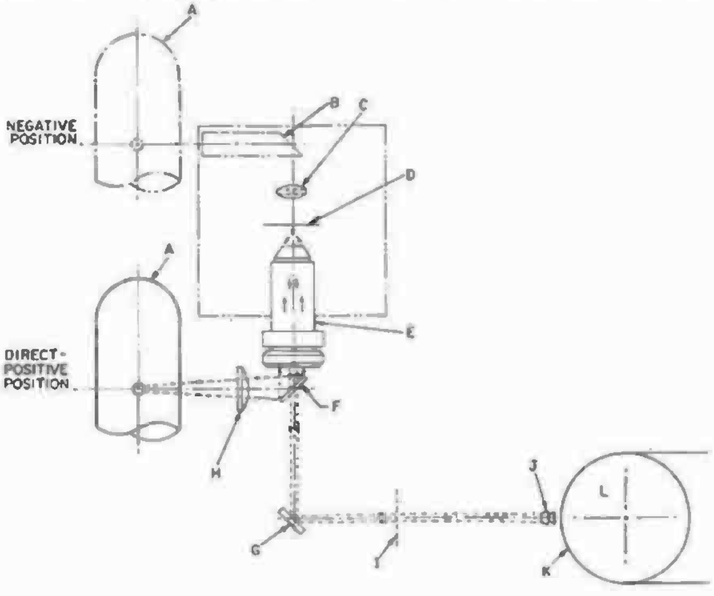


Fig. 18-338B. The optical schematic of the Westrex RA-1438 variable-area modulator unit used in the RA-1231 photographic film recorder.

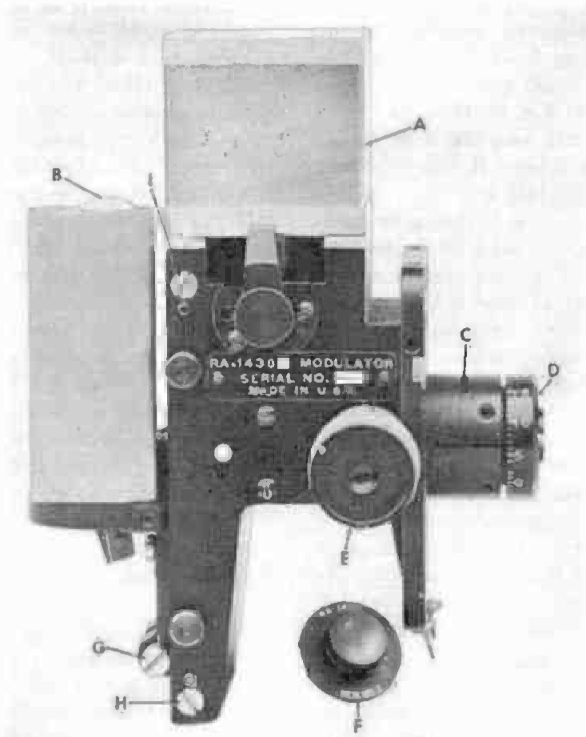


Fig. 18-338C. The Westrex RA-1438 modulator used in the RA-1231 photographic film recorders for recording negative and direct-positive sound tracks.

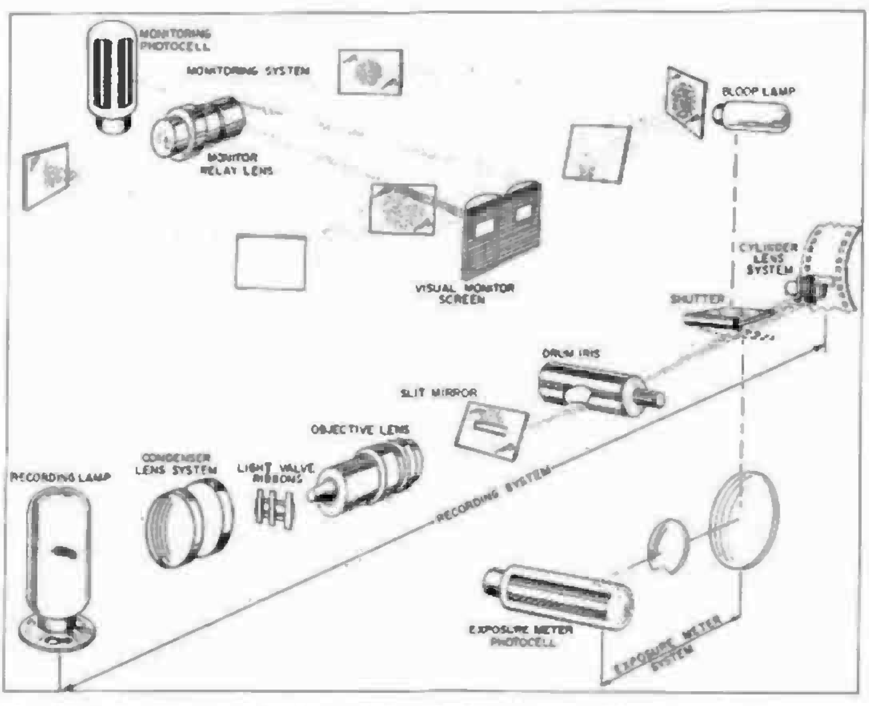


Fig. 18-338D. The optical schematic for the Westrex variable-area recording system using a light valve.

mirror F. The 45-degree mirror G reflects the light to the remainder of the optical system as described for negative sound track recording.

Returning to Fig. 18-333C, the principal components are: the light valve unit A, exposure lamp B, objective lens C, and its focus ring D.

At E is shown an eyepiece or periscope used for checking the noise-reduction current settings, and observing the percentage of modulation of the light valve ribbons. This eye piece must be removed when recording as it is inserted in the path of the light on its way to the objective lens. A calibrated scale in the eye piece permits the percentage of modulation of the light valve to be measured in milliamperes.

Part F is a cover to close the opening in the modulator when the periscope is not being used. Screw G is a tilting adjustment for the exposure lamp, and H an adjustment screw for adjusting the position of the exposure lamp in the direct-positive recording position. Adjustment screw I positions the exposure lamp for recording negative sound track. The exposure lamp in the illustration shown, is in position for direct-positive recording.

The schematic diagram for the optical system is given in Fig. 18-338B. A later type optical train used in the RA-1581A photographic recorder is given in Fig. 18-26B.

18.339 Describe a footage-counter system for rerecording stages.—Over the years, several different designs of footage counters have been used. The first systems were designed around the use of a Strowger stepping switch, used in automatic telephone exchanges. The switch was actuated by a commutator on the projection machine, and energized lamps painted with numerals lit up in various combinations to indicate the footage. Small mechanical counters mounted at the mixer console were the next to be used. These also operated from a commutator on the projection machine, or in some designs were driven by a selsyn interlock motor from the motor system. Another system employed a mechanical counter driven by a selsyn motor in combination with an optical projection system that projected the numerals on the upper or lower portion of the screen.

A later-type counter uses a selsyn motor to drive a group of printed-circuit switches which energize a group of lights on a readout unit placed either above or below the screen. The readout unit contains a group of plastic lenses on which numerals are painted. The image of the numeral is projected on a small plastic screen about eight inches in height. The readout unit is small enough to be mounted at the bottom or top edge of the picture screen. A reset button at the console actuates a motor that returns the counting switch to zero. When the counter is at rest, a time-delay circuit holds the lamp on the run position for one minute. If the system is not turned over within this time, the lamps are automatically cut off, thus increasing their life.

18.340 Describe a small studio combination magnetic and photographic film-recording channel.—The basic block diagram for a small studio installation employing both magnetic and photographic recorders is given in Fig. 18-340. This system can be used for recording music, dialogue, and looping, and then for rerecording to a photographic sound track, while recording a magnetic sound track for protection and playback.

At the left is a mixer console, with four inputs and a fifth input for reverberating the sound tracks. The four inputs may be from a microphone, turntable, or magnetic or film-reproducing devices. If a microphone is used for dialogue, an additional preamplifier will be required for each microphone, as the system shown is designed for an input level of between minus 10 and 0 dBm. An equalizer is shown in each input, although only one or two can be used. In this instance, a pad will be required to compensate for the insertion loss of the equalizer if it is removed. Leaving the equalizers, the signals are fed through the preamplifier and then to a four-position mixer-combining network with the four mixer controls.

At the output of the mixer is a hybrid coil feeding a compressor-limiter amplifier. The lower winding of this coil is fed from the output of a reverberation unit, or chamber, through a separate control at the mixer panel. The output of the compressor-limiter amplifier passes through a separate ceiling control, then to a 40/80-Hz high-pass filter, and then to a 6000/8000-Hz low-pass filter, which feeds a 600-ohm bridging bus. From the bus, the signal is fed through a monitor amplifier low-pass filter and a power amplifier, to the loudspeaker behind the screen. The VU meter is also fed from the bridging bus.

The bridging bus also feeds a magnetic film or tape recorder for protection tracks, a film-loss equalizer for the

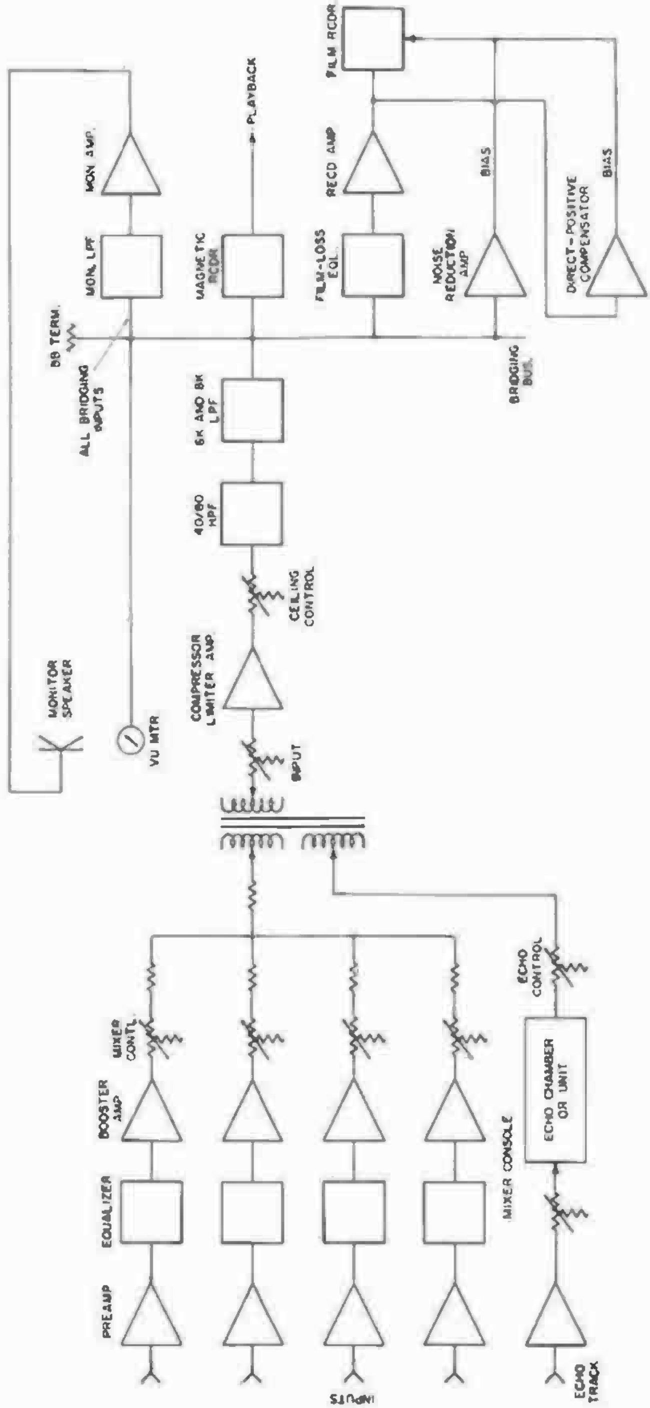


Fig. 18-340. Basic block diagram for a small studio installation employing both magnetic and photographic recorders.

film recorder, recording amplifier, film recorder, and noise-reduction amplifier. Many other combinations of equipment are possible, and are discussed in Section 9. The other devices are discussed throughout the various sections of this book.

18.341 Describe the procedure for checking out a rerecording channel.—Consistent recording quality requires that certain tests and measurements of the equipment be made daily. These tests consist of measuring the overall gain, distortion, signal-to-noise ratio, compression, monitoring level, and 100-percent modulation of the recorders. If photographic recorders are being used, the noise-reduction amplifiers and the margin must be checked. It is the policy in most recording establishments to make certain spot checks before each recording session. If the equipment has been off for any time, a reasonable amount of warm-up time must be allowed, particularly for exposure lamps in photographic recorders. It has been found from experience that equipment will give its best service if it is left on throughout the day. Generally, equipment turned on and off will require greater maintenance. A list of the most important tests and measurements are given below. They represent no particular equipment and will be altered by local conditions.

A. Send a 400-Hz signal corresponding to the output level of the reproducer into a mixer input. The mixer control is set to its normal operating point (generally 15 to 20 dB loss) and left there. Key out any equalization in the input circuit and set the compressor to *compression off*. Assuming that the system has been operating normally, the VU meter across the bridging bus should indicate its normal level (which is usually plus 10 or plus 14 dBm). The bus level should be within plus or minus 0.25 dB of its normal operating level (sine wave).

B. Key in the compressor. The level at the bridging bus should drop a given amount, depending on the compression ratio in use.

C. Measure the monitoring level. This should be within plus or minus 0.5 dB of the normal level.

D. Measure the signal-to-noise ratio across the bridging bus. It should be at least 60 dBm. If the bus is operated at plus 10 dBm, the signal-to-noise ratio is then 70 dB.

E. The distortion of the system with and without compression is then measured. The distortion at the normal bridging bus level without compression should not exceed 0.50 percent, with compression 1 percent.

F. A spot check of the frequency response may be made by sweeping the oscillator across the bandwidth of the console, taking into consideration any filters in the circuit.

G. The deflection of the light modulator in the photographic recorder is noted, and the noise-reduction amplifier is checked.

H. The 100-percent modulation of the magnetic recorders is noted.

I. Thread loops consisting of a 400-Hz or 1000-Hz signal on all the reproducers. For optical reproducers, the percent modulation should be 80 percent; for magnetic reproducers, 100 percent. The output level of each reproducer is adjusted to produce the same bridging bus level as the oscillator, with the mixer control set to the same loss as given in A.

J. The VU meter may now be set for a lead of 6 to 8 dB to prevent unseen peaks from overloading the system.

The various items used in a rerecording channel are dealt with in more detail in their appropriate sections, and their operations are fully explained.

18.342 Give a brief description of sound trucks.—Sound trucks are designed using two different methods. In the first method, the truck is designed to house portable equipment. In the second design, the truck houses permanently installed equipment, and it is much larger in size with more facilities provided.

In the first design, the truck is principally a carrier, with racks and fixtures built in for holding the portable equipment during transportation. For bad weather or nonset operation, the mixer is mounted on the right side of the drivers seat and the recorder is on a table in the interior, with the camera control and batteries. Usually a rack is mounted on either the side or the top of the truck for carrying the microphone boom.

The second design incorporates a modified studio installation. The truck contains a battery system which drives an alternator that generates single and three-phase power for operating the sound equipment and driving the cameras. A portable mixing panel with intercommunication is set on the stage, and is connected to the sound truck by cables. The permanently installed equipment is shock-mounted in racks which also contain test equipment and other devices necessary for its operation. In the studio, these trucks are parked outside a stage or may be used for locations. However, most heavy sound trucks have now been replaced with light portable equipment, as described throughout the various sections.

18.343 Describe how the monitoring level for a rerecording stage is determined.—Each rerecording stage has a particular monitoring level that is best for the size of the room, the picture, and the distance the mixing console is placed from the screen. There is a definite psychological relationship between these factors. If the picture is small for the distance and the monitoring level is low, the mixer will have a tendency to record everything at too high a level. If possible, the picture size should be slightly oversize for the room and the monitoring level kept on the high side. This keeps the sound perspective in relation to the picture in balance.

After the monitoring level has once been determined for a stage, it should be kept within 0.5 dB by measuring the electrical monitor level daily. If the stage is large, temperature and humidity will also have their effects.

For recording 16-mm pictures, the monitoring level should be reduced 3 to 4 dB below that normally used for recording 35-mm pictures. The reason for this is that dialogue for 16-mm pictures is generally recorded with an average level of 70 to 85 percent, whereas for 35-mm, it is recorded at an average level of 40 to 60 percent. The monitoring level, if left set as for 35-mm recording, would be too high. Thus, the mixer has a tendency to pull everything down in level. Reducing the monitoring level by 4 dB will result in a proper balance for 16-mm between the dialogue, sound effects, and music.

It is assumed that before the monitoring level is finally set, the monitor low-pass filter (if used) has the proper cutoff, and the low-end characteristic of the stage has been properly adjusted for its reverberation characteristic. This may be ascertained by the use of an Academy Test Reel (see Question 19.62). The reproducing characteristic should never be over-bassed, or the high end made "hot." This only results in false reproduction, and the final recorded program material will not be suitable for either theater or television reproduction. The final listening quality of a stage should be natural and free from distortion, with a high rate of intelligibility. When rerecording, if the original dialogue has a bassy sound, then a low end rolloff should be used (as discussed in Questions 6.80 and 6.122).

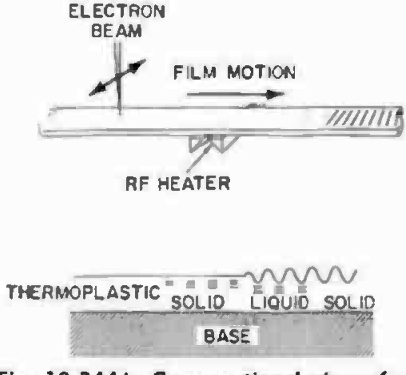


Fig. 18-344A. Cross-sectional view of a thermoplastic (TPR) tape for recording optical images.

18.344 Describe a thermoplastic recording system.—Thermoplastic recording (TPR) is a development of General Electric, invented by Dr. William E. Glenn of the General Electric Research Laboratory. The recording medium is a thermoplastic film, with a base similar to regular motion picture film, and requires no processing. It is said to combine the immediate playback features and the versatility of magnetic film, or tape, with the storage capacity of photography.

Electrons are used to convert information to be recorded, including visual images, into microscopic wrinkles in the surface of the film. The ripples are formed by means of a minute, modulated electronic beam which scans the surface of the film, and charges the surface in a pattern comparable to the image to be recorded. The recorder has an electrical input similar to a magnetic recorder, and an image output similar to photographic film. The image can, if

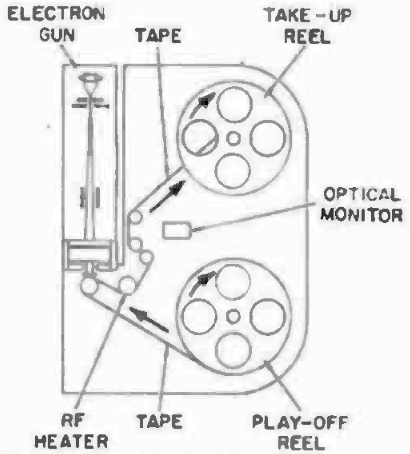


Fig. 18-344B. Basic plan of the thermoplastic recorder.

necessary, be converted to electrical impulses.

The basic principle for recording the image is shown in Fig. 18-344A, a cross-section of the theromplastic film. On top of the base is a transparent conducting coating, and on top of this is a thin coating of thermoplastic material. This material will melt when subjected to fairly high temperature. The surface of the thermoplastic is charged with an electron beam in a pattern that corresponds to the pattern of ripples from the image. As the film moves on, a current induced in the transparent conducting coating heats the film, and the thermoplastic coating melts. The charges are attracted to the conducting coating and depress the surface of the thermoplastic. After the surface has been deformed by the charges, the film is permitted to cool, which freezes the ripple pattern into place in less than 1/100 second. The recorded information may be erased by heating the thermoplastic again to a higher temperature, to permit the

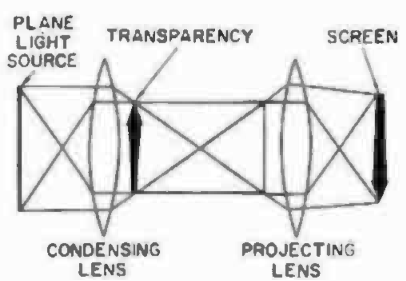


Fig. 18-344C. Conventional motion picture projector optical system.

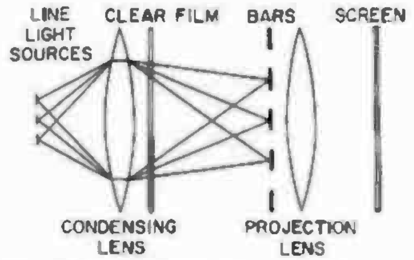


Fig. 18-344D. Projection optical system for a TRP projection system.

charges to leak away. Surface tension smooths the surface back to its original shape, and the film then becomes reusable.

The basic design for a thermoplastic recorder is shown in Fig. 18-344B. The film is recorded in a vacuum (0.1 micron pressure) since the electron gun of the recording mechanism must be in a vacuum. This is not difficult because with modern equipment, the required vacuum may be achieved in about one minute. The signal source for this particular recorder is simply the intermediate frequency taken from a black and white television set with about a 1-volt level. For color recording, another signal of the same voltage level is added to an additional electrode.

To illustrate the principles of projecting the picture, the basic principle of a standard motion picture projector is shown in Fig. 18-344C. It employs a plane light source, condensing lens, and a projection lens. The light source is imaged on the projection lens by means of the condensing lens; in turn, the projection lens images the slide to the projected screen.

If this plan is modified to use a line light source and image on a set of bars in front of the projection lens (Fig. 18-344D), no light passes through the

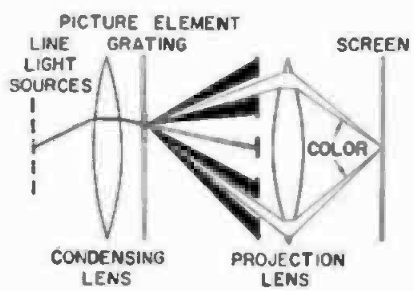


Fig. 18-344E. Optical system for a TRP system projecting a color film.

screen. If, however, there are ripples on the surface of the film at any particular point, these will scatter light through the bars. Thus, the bars act somewhat like a shutter, and permit the light to pass where ripples occur. As the light passes through the projection lens, it images the ripple on the screen as a white spot. A clear slide with ripples on the surface will appear as a black and white picture when projected on the screen.

The system can be modified to produce color pictures, as shown in Fig. 18-344E. Each picture element has in it a set of ripples that form a small diffraction grating. The light that is diffracted by this grating forms a spectrum on each side of the central beam. The slots are made small enough so that only one color from the spectrum passes through the projection lens. Since the projection lens can receive only one color of the spectrum, the spot that appears on the screen will appear as a single color. For the spot to appear as another color, the spacing of the grating is changed, allowing a different part of the spectrum to pass through the grating. To produce a color that is formed by a superimposition of two or more colors, the gratings are superimposed and superimposition of the color results.

To illustrate its storage capacity, a TPR system could, in principle, record 24 volumes of the *Encyclopaedia Britannica* on a reel the size of a spool of thread, taking only one minute to record each volume. The recorder uses a 16-mm film base with 8-mm perforations, with an image size for 8-mm film.

18.345 Show a typical mixers log sheet for production recording.—A production mixers log is shown in Fig. 18-345. The time, scene, and take number are entered for each take, along with the crew names and other pertinent data.

18.346 Describe a microscope suitable for optical sound-track inspection.—Such a microscope is shown in Fig. 18-346. The microscope sample stage is

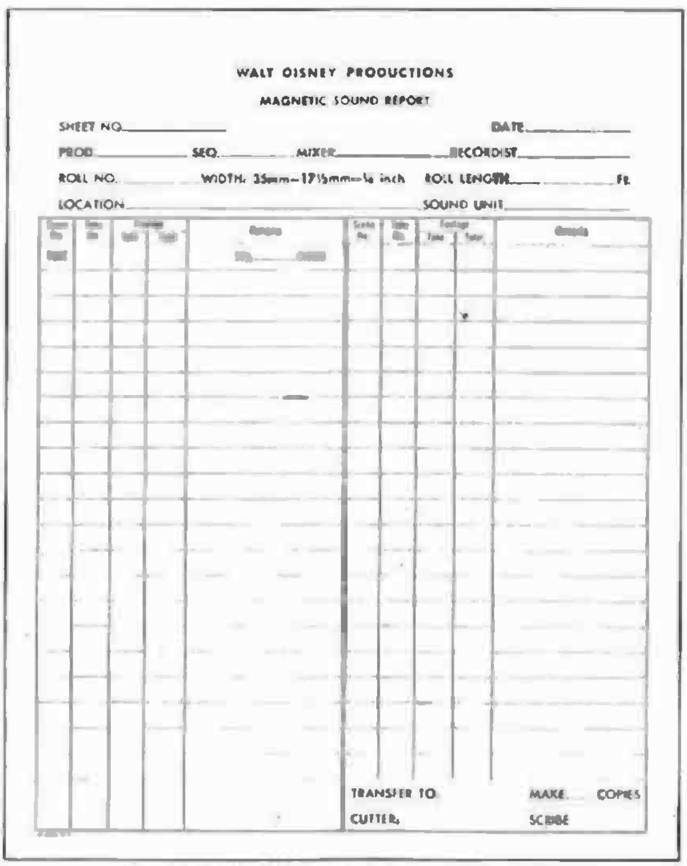

WALT DISNEY PRODUCTIONS
MAGNETIC SOUND REPORT
SHEET NO. DATE
PROD SEQ. MIXER RECORDIST
ROLL NO. WIDTH: 35mm—17½mm—¼ inch ROLL LENGTH Ft.
LOCATION SOUND UNIT
TRANSFER TO MAKE COPIES
CUTTER SCRIBE

Fig. 18-345. Production mixers log sheet.

modified and fitted with two guide rollers and bars to hold either 16- or 35-mm film in place. The film may be mounted on rewinds if desired and passed under the objective lens of the microscope. The rollers move inward for 16-mm film inspection. The eye piece should be fitted with a reticule calibrated zero to 100 mils, with the capability of being read to within 0.5 mil. A magnification of 10 is generally sufficient for most purposes.

18.347 Describe an optical comparitor.—A comparitor is an optical device for magnifying samples placed on a stage. It is similar to a microscope except that the image of the object is rear-projected on a screen about 12 inches in diameter. Transparent scales with many different type calibrations may be placed over the viewing screen. Micrometer adjustments on the sample stage permit the measurements of the sample in two planes in 1/10,000 inch.

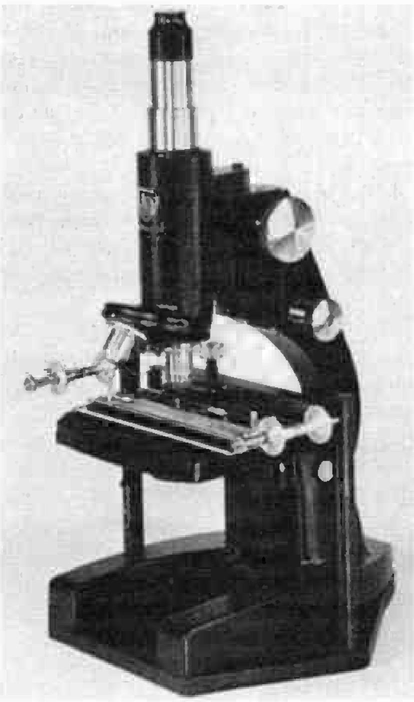


Fig. 18-346. Microscope for inspecting and measuring optical sound track.

Such devices are quite useful where a large amount of optical film has to be inspected, as the operator is not required to use an eye piece, but only to view the sample on a screen. Comparitors can be used for inspecting the sharpness of the sound track, placement, modulation amplitudes, and many other measnrements. They are quite useful for inspecting magnetic recording and reproducing heads, particularly if the gap is less than 1 mil in height. Objects may be magnified 10 to 100 times, depending on the particular objective lens employed.

18.348 Describe a solid-state modulator for recording on photographic film.—A Russian scientist, Lossov, demonstrated in 1922, that optical film could be exposed by the electroluminescent properties of silicon carbide. Only recently with semiconductor devices available, this phenomenon has been applied to sound recording on photographic film. A silicon-carbide diode emits either a yellow or blue light, depending on the impurities used to form the junction. This material is high on the hardness scale. It will withstand temperatures up to 1000°C, and will operate at near room temperatures. Its life is limited only by its external equipment, snch as leads and contacts.

Diodes capable of emitting light are termed "light-emitting diodes" (LED)

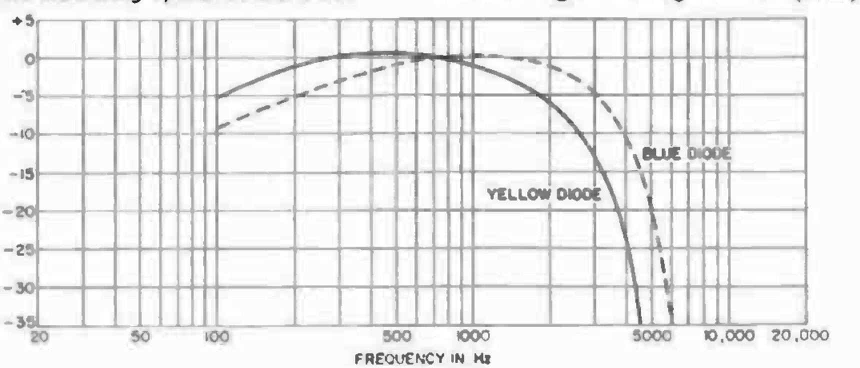


Fig. 18-348A. Frequency response for silicon carbide electroluminescent diode. Recorded on Kodachrome IIA color film.

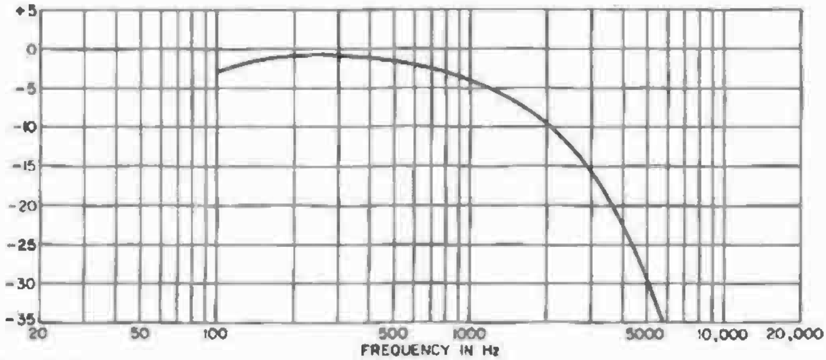


Fig. 18-348B. Frequency response for silicon-carbide electroluminescent diode. Recorded on black and white Plus-X reversal film.

and change electrical energy into light energy without passing through a thermal stage, as do incandescent lamps.

For 16-mm recording, the dimensions of the diode are $0.5 \times 1 \times 2$ cm, with the 1×2-cm face parallel with the light-generating junction. It is thought that the light is emitted from a region 10^{-4} cm thick, about two light wavelengths in width. While blue silicon carbide is somewhat faster, the yellow diode is essentially constant over the audio spectrum, being down less than 6 dB at 50 kHz. Frequencies of 100 to 6000 Hz have been achieved on 16-mm film with suitable equalization (diode dimensions curtail the range). The light is produced by the forward bias at the p-n junction which is about 1×2 mm for 16-mm recording.

The frequency-response measurements given in Figs. 18-348A and B were recorded using constant-amplitude frequencies applied to the input of the diode. The loss of high-frequency response can be overcome to a great extent by the use of equalization in the recording amplifier circuits. Such devices are under development for optical sound-track recording on both 16- and 8-mm photographic film.

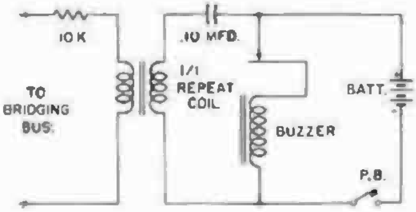


Fig. 18-349. Marking device for indicating selected takes when looping dialogue.

The diode is held in close proximity to the film (not in actual contact). No electro-optical parts are required, and the recording is of variable-density type. The operating point is at 10-percent transmission for black and white, and 70-percent transmission for color at 9000A. Signal-to-noise ratio is about 30 to 40 dB for black and white, and 20 dB for color.

Power specified is 15V at 50 mA for 16 mm. For 8-mm recording, the diode diameters are smaller; thus, less power is required. The diode light output is slightly sublineal with current. The THD is on the order of 4.9 percent at 40 percent modulation. For further discussion of this subject, the reader is referred to the reference.

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Section 19

Motion Picture Projection Equipment

Heinrich Hertz discovered photoemission in 1887, but until Dr. Lee DeForest invented the "Audion" in 1907, photoemission could not be electronically amplified and reproduced. Lauste and Ruhmer were experimenting with photographic sound recording around the turn of the century. Edison, Gaumont, Whitman, Madelar, and others worked to develop talking pictures, but most projection involved phonographic devices. More than eighty different systems, using both film and disc, were on the market around 1928. No standardization existed, which must have resulted in confusion in the projection booth.

With the varied-type productions of today, this confusion could still occur, unless the projectionist is well equipped and has a store of knowledge to meet the problems. He must be somewhat of a showman, and he must be familiar with lenses, apertures, aspect ratios, synchronization, and various type sound tracks. Also, he must understand the proper use and care of his equipment. In the motion picture studio, the projectionist is a key figure who works closely with the sound department. He operates specialized equipment adapted for dubbing, looping, and the running of dailies.

Maintenance is generally the responsibility of the sound department. This department has at its disposal such test films as Academy test reels, frequency films, and track for checking alignment and flutter, etc.

In addition to the types of motion picture projection equipment previously mentioned, this section also deals with 8- and 16-mm projectors, combination 35-, 55-,
tioned, this section also deals with 8- and 16-mm projectors, combination 35-, 55-,
general information. The Standards given and quoted in this section are those in present use. However, several are at present under review and may be changed in the near future.

19.1 What are actinic rays?—Light ray colors to which photographic emulsions respond, such as red, green, blue, violet, and ultraviolet.

19.2 What is a lambert?—A unit of brightness.

19.3 What is a lumen?—The unit of luminous flux representing a definite rate of light emission.

19.4 What is a lux?—The practical unit of illumination.

19.5 Explain candle and candela, and their associated terms.—One candle power is the amount of light given off by one candle over one square foot of surface, when located one foot from that surface. Other terms associated with the measurement of light are given.

$$\text{Foot-candles} = \frac{\text{lumens}}{\text{area}} = \frac{\text{foot-lamberts}}{\text{reflectivity}}$$

$$\text{Lumens} = \text{foot-candles} \times \text{area} = \frac{\text{foot-lamberts} \times \text{area}}{\text{reflectivity}}$$

$$\text{Foot lamberts} = \frac{\text{lumens} \times \text{reflectivity}}{\text{area}} = \text{foot-candles} \times \text{reflectivity}$$

$$\text{Foot-lambert} = 3.426 \text{ nits}$$

$$\text{nit} = 0.2919 \text{ foot-lambert}$$

The *nit* is an international unit of luminance equal to 1 candle/sq meter. Motion picture screen luminance in the United States is generally measured in foot-lamberts, although in international usage, the preferred unit is the nit.

Luminous intensity or candle power is a measure of light source which describes its luminous flux per unit solid angle in a particular direction. For many years, the standard measure of luminous intensity has been the international candle, established by the National Bureau of Standards. The light from a group of carbon filament lamps was used as the basis for this standard measure. In 1948, the International Commission on Illumination agreed on the introduction of a new standard of luminous intensity, *candela*, to distinguish it from the term *candle*. The candela is defined by the radiation from a black-body radiator operating at the solidification temperature of platinum. One candela is the luminous intensity of $\frac{1}{60}$th square centimeter of such a radiator. The effective change in the candle is on the order of tenths of a percent.

The term *luminous flux* is used to denote the time-rate flow of light energy, that characteristic of radiant energy which produces visual sensation. The unit of luminous flux is the *lumen*, which is the flux emitted in unit solid angle by a uniform point source of one candela. Such a source produces a total luminous flux of 4π lumens. Illumination is the density of luminous flux incident on a surface. The common unit of illumination is the *foot-candle*.

Black-body radiation can be explained as follows. As a body is raised in temperature, it first emits radiation primarily in the invisible infrared region. Then as the temperature is increased, the radiation shifts to the shorter wavelengths in the visible spectrum. If the radiating body is one that can be technically called black, its behavior can be accurately described by the laws of radiation. A black body is one which *absorbs* all incident radiation. This radiation is neither transmitted nor reflected.

19.6 How does the intensity of light vary with distance?—It varies inversely as the square of the distance.

19.7 What are the basic types of lenses?—Plano-convex, double-convex, concave-convex (sometimes called a meniscus), plano-concave, and double-concave. Lenses are divided into two groups, positive and negative. A positive lens converges the light rays and must be thicker at the center than at the outer edges. A negative lens diverges the light and must be thicker at the borders than at the center. Basic lenses are shown in Fig. 19-7.

19.8 What is a wide-angle lens?—One in which the field of view covers an area of 70 to 100 degrees.

19.9 What is the focal length of a lens?—The distance from the focal point to the center of the lens.

19.10 What is the f number of a lens?—A number used to indicate the speed of a lens. The number is equal to the focal length divided by the effective aperture. The f value deals with the relationship of the diameter to the focal length. A lens with an f1 rating means that for every inch of focal length there will be one inch of diameter. Therefore, an f1, 6-inch lens will have a diameter of 6 inches. An f2, 6-inch lens will have a 3-inch diameter, or 2 inches of focal length for every inch of diameter.

19.11 What is a lens "T" stop calibration?—An individual calibration of a lens relative to its light passing capabilities. A "T" stop calibration takes into account the light lost by reflection, the number of elements in the optical train, the light lost by absorption as it passes through the lens elements, and manufacturing tolerances. As a rule, the lens elements are coated, which reduces the light loss from about four to one percent.

In the f system of lens calibration, each f value is computed mathematically from the physical dimensions of the lens system. Lenses calibrated by the "T" stop method use a calibrating light which is passed through the lens, and the loss of light for different diaphragm settings is measured by means of a photocell. Thus, a "T" stop calibration is an individual calibration of

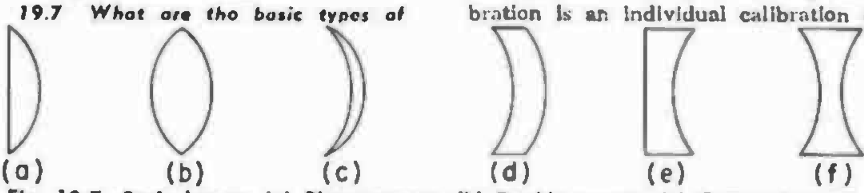


Fig. 19-7. Basic lenses. (a) Plano-convex. (b) Double-convex. (c) Converging concavo-convex. (d) Diverging concavo-convex. (e) Plano-concave. (f) Double-concave.

the true light-passing capabilities of the lens. As a rule, lenses used in the motion picture industry are calibrated at regular intervals or after they have been serviced. The term "T" is short for transmission.

19.12 What is meant by the speed of a lens?—It is the diameter relative to the equivalent focus (abbr. ef). Projection lenses vary in diameter from 1½ inches to 5 inches. The diameter has no bearing on the picture size; however, the smaller the lens diameter, the more sharply defined will be the picture, but with less light. The larger the lens, the greater the amount of light, but with less definition.

Projection lenses are made in three sizes: quarter, half, and full size. A quarter-size lens means that the diameter is one-fourth the equivalent focus. For example, a quarter-size 8-inch lens has a diameter of two inches.

19.13 What is the optical axis of a lens?—The one path (through an optical system) which does not change the direction of light rays transmitted by that lens.

19.14 What is chromatic aberration in a lens?—Color fringes when white light is passed through a lens caused by the light rays of different colors being bent different amounts. This effect is also called light dispersion.

19.15 What is spherical aberration in a lens?—A deviation of the light rays as a result of different zones of the lens having different focal lengths. The rays of light form images at various points along the optical axis, resulting in a blurred image.

19.16 Does the curvature of a lens affect the focal length?—Yes. The greater the curvature the shorter will be the focal length, because of the greater angle of incidence and consequent greater refraction of the light rays.

19.17 What is a meniscus lens?—One with a concave and a convex surface. (See Question 19.7.)

19.18 What are Newton's rings?—Colored rings due to light interference. They may be seen about the contact area of a convex lens with a plane surface, or they may be caused by two lenses differing in curvature.

19.19 What does the term anamorphosis mean?—It is a term applied to a lens system used in the photographing and reproduction of wide-screen motion pictures. The image is photographed using a predistorted optical image, which elongates the image in the vertical plane. The image is restored to its normal appearance by the use of the lens system in the projector. This particular lens system has a reverse characteristic of the lens system used to photograph the image, and it was developed by a French physicist, Dr. Henri Chretien. This system, using an aspect ratio of 2.35:1, was first used by 20th Century-Fox, in the production of *The Robe*, in 1953. This system is now used by most motion picture studios, with aspect ratios up to 2.94:1.

In the original CinemaScope system, the camera lens sees a picture twice as wide as the conventional lens does, but it compresses the image horizontally by 50 percent, in order to eliminate the need for a film twice as wide. When projected, the projector lens system spreads the image to the desired aspect ratio. The image ratio or screen is generally on the order of 2.35:1, or greater. (See Questions 19.83, 19.123, and 19.126.)

19.20 What is a coated lens?—A lens in which the elements are coated with a solution of magnesium fluoride to reduce reflections from the surface of the glass and to improve the transmission of light. A coated lens will transmit approximately 30 percent more light than an uncoated one. As an example, uncoated glass reflects about 6.5 percent of the light falling upon it. When the lens is coated, this figure drops to about 0.30 percent. A six-element uncoated lens may transmit 73.51 percent of the light, but when coated, this same lens will transmit 94.15 percent of the light. An eight-element uncoated lens may transmit 66.34 percent of the light, but when coated, the transmission is increased to 92.27 percent.

Coated lenses may be identified by shining a light on their surfaces and noting the color of the reflected light. If the lens is coated, the reflected light will appear light blue in color.

19.21 What is the recommended method for cleaning projection lenses?—Modern projection lenses must be cleaned very carefully. Although they may be coated and present a fairly hard surface, the coating is microscopically thin and may be scratched very easily. Solvents such as alcohol and petroleum ether should be used very sparingly, as these solvents may attack the optical cement or the lens-mount lacquer. A lens should be cleaned with a lint-free cotton cloth and a camel's hair brush. Heavy scum may be removed with a cleaning solution of mild pure soap in water. Precaution should be taken that the solution contains no caustic soda. Isopropyl alcohol has been found to be a good cleaner. Several commercial lens cleaners, sold under various trade names, are also available.

19.22 ***How is a projection lens using multiple elements constructed?***—In Fig. 19-22 is shown a six-element lens used in a motion picture projector manufactured by the Kollmorgen Optical Corporation. The various optical elements may be clearly seen.

19.23 ***What is an iris?***—An adjustable lens diaphragm.

19.24 ***What is the frame rate for silent motion picture projection?***—

35-mm, 24 frames per second
16-mm, 18 frames per second
8-mm, 18 frames per second

Originally, the frame rate for both 8- and 16-mm silent projection was 16 frames per second. However, to eliminate flicker, the frame rate has been increased to 18 frames per second. A three-blade shutter is used, and this results in a flicker rate of 54 Hz.

19.25 ***What are the standard number of frames per foot of film?***

35-mm, 16 frames per foot
16-mm, 40 frames per foot
8-mm, 72 frames per foot (nonprofessional)
8-mm, 80 frames per foot (Super 8-mm professional)

(See Questions 19.26, 19.31, and 18.50.)

19.26 ***What is the projection frame-rate per second for 8-mm film?***—For single film, the rate is 18 frames per second and for Super 8-mm, it is 24 frames per second. The sound is recorded on a magnetic stripe, using a frequency response within the limits of the Standard given in Fig. 17-170. Before the issuance of the present Standard, the linear speed was 16 frames per second. However, because of the improvements in projector design and illumination (with a noticeable reduction of picture flutter on the screen), the linear speed has been increased to 18 frames per second (nonprofessional), and 24 frames per second for Super 8-mm film (professional). A three-bladed shutter is employed.

Although the linear speed for professional sound projection is 24 frames per second, in the nonprofessional field, using a single system or sound camera with prestriped film, the sound and picture are run at 18 frames per second to conserve film. If the picture and sound are run at 24 frames per second, the product is termed "Super 8-mm film." (See Questions 19.25, 19.31, and 18.50.)

19.27 ***Can optical sound track be recorded on 8-mm film?***—Yes. Considerable work has been done in this direction. However, it is the general practice at the present time to magnetically

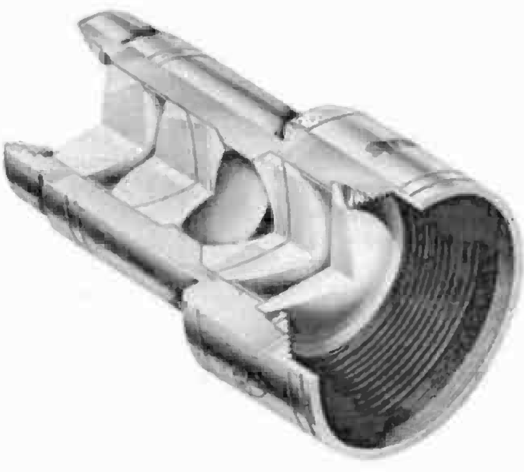


Fig. 19-22. Kollmorgen Super Snoplite six-element projection lens.

stripe the print. The magnetic frequency response is given in Fig. 17-170, as set forth in USASI (ASA) Standard PH22.134-1963. (See Question 18.348.)

19.28 Give the number of perforations per frame of picture.—

8-mm, one hole above and below the frame, on the frame line, one edge only.

16-mm single perforation, one hole above and below the frame, on the frame line, one edge only.

16-mm double perforation, one hole above and below the frame, on the frame line, both edges.

35-mm, four holes each side of the frame splitting the frame line.

55-mm, 6 perforations.

70-mm, 5 and 6 perforations.

As a rule, double perforated 16-mm film is used only in the amateur photographic field.

19.29 Why is the sound track advanced or retarded in relation to the picture?—To compensate for the physical displacement of the photocell or magnetic head, relative to the picture head. When magnetic sound tracks are used, the sound track is retarded, or behind the picture. Optical sound tracks are advanced relative to the picture. In the first instance, it is termed, *pull-up*, and in the second, *pull-down*, with reference to the picture aperture. For 35-mm and 16-mm projection, this subject is discussed in Question 19.44. For 8-mm projection, an optical sound track is advanced 22 frames, and for a magnetic track, it is advanced 18 frames. This is the present recommended practice, since no Standard has been published. (See Question 19.30.)

19.30 What is the sound track displacement relative to the picture?—The actual displacement of the sound track will depend on whether the track is magnetic or optical. For reproducing 35-mm optical sound track, the phototube and its associated equipment is placed in a sound head below the picture head, and the magnetic reproducing head is housed in a penthouse above the picture head. For standard optical sound track reproduction, the sound start mark is 20 frames ahead of the picture aperture, while the 35-mm magnetic sound start mark is 28 frames behind the picture. The only reason for this displacement is purely one of mechanical design. For 70-mm film, the sound start mark is 24 frames behind the picture.

In studio 35-mm projection rooms where only rough cuts of picture consisting of a single sound track and picture are run, the machines are fitted with a magnetic sound head placed in approximately the same position as the phototube. Most projectors used for dubbing are set up in this same manner. Therefore, the sound start mark is set the same for both types of sound track —20 frames ahead of the picture. If the projectors are fitted with preview magazines, the sound start mark is also 20 frames ahead of the picture.

The sound start mark for 16-mm film is placed 26 frames ahead of the picture for both optical and magnetic reproduction. These start marks relative to the picture start mark are shown in Fig. 19-44. A projector with both a sound head and penthouse is shown in Fig. 19-80A.

In large theaters where the projection throw is greater than 100 feet, it may be necessary to advance or retard the sound track a few sprocket holes to bring the sound into proper synchronization at audience distances of 100 feet or more from the screen. This may be accomplished for optical sound track by threading the sound start mark at the 19th, rather than the 20th frame. In the average theater it is necessary to emit the sound before the corresponding picture frame is projected, since sound travels at approximately 1100 fpm or about 50 ft per frame for a normal projection rate of 24 fpm. The projectionist can place the sound in synchronization in the theater by varying the length of the threading path between the projector and the sound head.

For release prints using magnetic sound tracks and a penthouse, the start mark is moved upward. Installations using separate sound heads running interlock with the projectors will move downward. To summarize:

Film and Sound Track	*Displacement Relative to Picture*
35-mm optical	20 frames ahead
35-mm magnetic	28 frames behind
16-mm optical	26 frames ahead
16-mm magnetic	26 frames ahead
8-mm magnetic	18 frames ahead
8-mm optical	22 frames ahead
70-mm magnetic	24 frames behind

The displacement for 8-mm film is the recommended practice at this time, as no standard has yet been published.

19.31 At what linear speeds are sound motion pictures projected?—Sound motion pictures must be reproduced at exactly the same linear speed at which they were recorded. Standard linear speeds in the industry are:

35-mm, 90 feet per minute, or 18 inches per second,
16-mm, 36 feet per minute, or 7.2 inches per second,
8-mm, 18 feet per minute, or 3.6 inches per second (nonprofessional),
8-mm, 20 feet per minute, or 4 inch per second (Super 8-mm professional).

For home use or nonprofessional use, the slower speed is used to conserve film. (See Questions 19.25, 19.26, and 18.50.)

19.32 What is projection synchronism?—The physical distance the start marks for sound and picture are separated. (See Questions 19.29 and 19.30.)

19.33 How do projection light systems compare to direct sunlight?—The intensity of direct sunlight is approximately 10,000 foot-candles. In Fig. 19-33 is a graphical comparison for four sources of light—sunlight, incandescent, the conventional arc light, and xenon gas lamps. It will be observed that the incandescent lamp is the poorest of all, while the arc lamp falls off rapidly in the shorter wavelengths. The xenon lamp approaches that of the sun in the shorter wavelengths. The selection of a proper light source for screening rooms is an important factor, particularly for color prints.

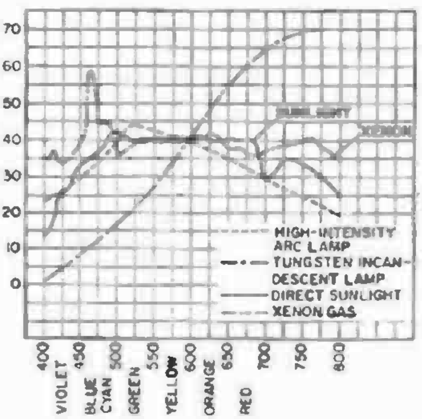


Fig. 19-33. Relative intensity of tungsten incandescent lamp and high-intensity arc lamp compared to direct sunlight.

It should be mentioned that the arc lamp and the xenon lamps are of extremely high intensity, and should never be directly viewed without eye protection.

19.34 How may the candle power of an incandescent lamp be estimated?—Properly heated, about 27 candle power per square millimeter of heated surface.

19.35 What is the average brilliancy in candle power of an arc light used for motion picture projection?—For low-intensity arcs using 15 to 45 amperes, approximately 150 candle power per square millimeter of heated surface. High-intensity arcs using 50 to 150 amperes have a brilliancy in the order of 700 candle power per square millimeter of heated surface.

19.36 How are the polarities of the carbons determined in a projection arc light?—If operated from direct current, the rear carbon is always negative. As the carbons burn down, a crater is formed in the positive carbon. An image of the crater is reflected by a concave mirror at the rear of the negative carbon on the picture aperture. Because the negative carbon protrudes through a hole in the reflecting mirror, some loss of light will take place.

19.37 Why is it desirable to use direct current with an arc light?—Principally because the light is steadier than if operated from alternating current. Also, if ac is used, special carbons must be employed to produce a white light. Arc lights operated from alternating current polarize the carbons in one direction more than the other. Because the carbons are alternately positive and negative, craters are formed in both carbons, resulting in an unsteady light with less brillance.

19.38 What type rectifiers are recommended for arc lights?—Vacuum tube, silicon-diode rectifiers, or a motor generator. The silicon-diode rectifier is preferred.

19.39 What is the purpose of a motor feed in an arc light?—Because the carbons burn slowly at the ends, they require constant attention to maintain a steady light intensity. They must be moved toward each other at a steady rate comparable to their burning rate.

A drive motor, operated from a relay connected across the carbons, moves the carbons at a predetermined rate; thus, the light intensity remains constant. (In high-intensity arc lights, the negative carbon is rotated to obtain an even burning surface.) The motor relay is controlled by the voltage drop across the carbons. As the carbon burns away, the voltage drop increases, the relay closes, and the motor starts to close the gap between the carbons. When the carbons reach a certain point, the voltage drop across the relay is quite low, the relay drops out, and the motor stops. Thus, the distance between the carbons and the light output is kept fairly constant.

19.40 ***Does all the light from the arc reach the screen?***—No; about 5 percent of the light striking the lens is reflected, with another one-half percent lost because of the density of the lens system. The aperture at the film reduces the light approximately 50 percent. Thus it is evident that less than 50 percent of the light actually reaches the screen. If dirt is permitted to accumulate on the lens and porthole glass, an additional loss is induced. It is not uncommon to find that only 30 percent of the light leaving the arc arrives at the screen. If the projection system is used for projecting wide-screen pictures, it will be necessary to increase the amperage of the arc to two or more times that normally used for a standard Academy aperture ratio.

As a typical example, a review room normally using 42 amperes is required to increase the current to 75 amperes to obtain a satisfactory picture when projecting a wide-screen picture. The increase in current is governed by the ratio of the picture, the type of screen, the efficiency of the lens system, and the lamphouse reflection system.

19.41 ***What is a crater?***—The depression burned in the positive carbon when direct current is used to operate an arc light.

19.42 ***What is a douser?***—An automatic control set in motion by the projectionist when cutting over from one projector to another. The douser cuts off the light from one machine, opens the other, turns on the sound-head exciter lamps, and transfers the sound circuits from one machine to the other.

19.43 ***How is projector changeover indicated?***—By small opaque dots which appear in the upper right portion of the picture. The changeover operation consists of two steps: the motor cue and the actual changeover. When the first group of dots are seen, the motor of the projector is started. When the second group of dots are seen, the changeover switch is actuated. This action operates the douser that cuts off the light of the machine which is running, and opens the douser shutter on the machine being cut over. Assuming that the up-to-speed time of the projectors is known by the projectionist, the changeover can be made smoothly, and the audience does not have to be aware of the changeover. Motor cue and changeover cue marks can be seen on the standard leader (Fig. 19-44).

19.44 ***Describe an SMPTE universal picture leader.***—The leader is a length of film attached to the head end of a motion picture. It contains sound and picture start marks, cutover cues, and other information required by the projectionist. The various markings are placed at standard distances and may be used in any standard projector either for 16-mm or 35-mm projection. Prints that are suitable for making negatives of this standard leader are available from the SMPTE (Fig. 19-44).

The SMPTE recommended practice RP-25-1968 states that for 16-mm and 35-mm synchronization information during the editing and rerecording periods, a single frame of 1000 Hz modulated 80 percent can be inserted to coincide with single frame number-2 of the universal leader when aligned for editorial sync. The same procedure is used for both magnetic and optical sound track. (See Question 18.332.)

19.45 ***What is the aspect ratio of a picture?***—The ratio of the width to height. The standard ratio is 1.34:1 and is often referred to as the Academy Standard because this ratio was originally adopted by the Academy of Motion Picture Arts and Sciences. (See Question 19.83.)

The width of the screen is generally based on a ratio of three times the distance from the rear row of seats to the screen. Thus, a screen of 24 feet would be approximately 75 feet from the last row of seats. However, for wide-screen presentation this ratio does not always apply. For dubbing stages (rerecord-

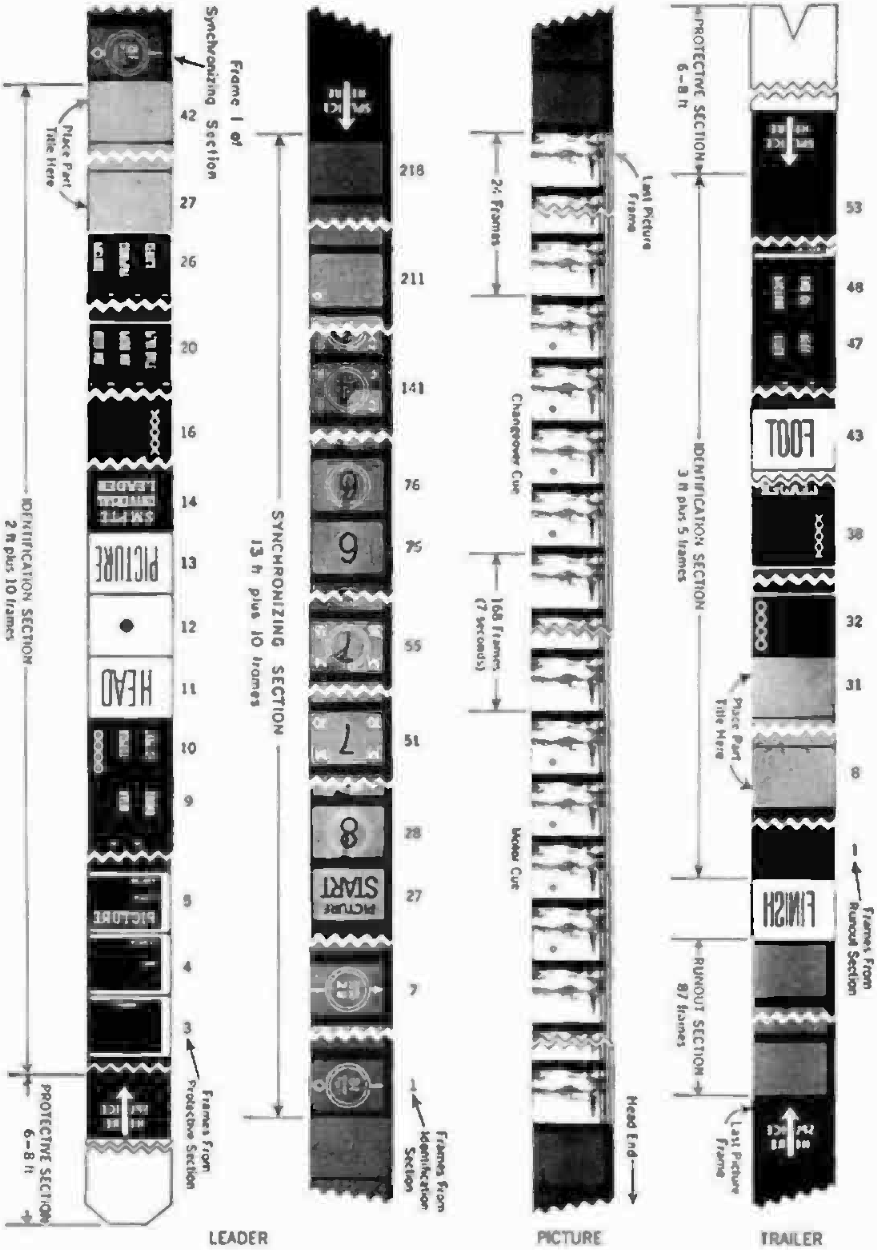


Fig. 19-44. USASI (ASA) Standard PH22.55-1966 leader for composite print containing both picture and sound track, as viewed from the light source in a projector. For 16-mm film, the sound track is on the left edge of the film.

ing), the screen is generally larger than it would be for a theater of equivalent dimensions.

19.46 What are the aspect ratios for wide-screen projection? — Wide-screen projection has used many different ratios since its introduction by Lyman H. Howe. He used this for his travelogues several years ago. Presently, the ratios are many and varied and range from 1.66:1 to 2.94:1. A tabulation of the ratios in use today is given in Question 19.83.

19.47 What is the relationship between the size of the picture and the amount of light required?—Halving the focal length of the lens increases the picture area four times. This means that

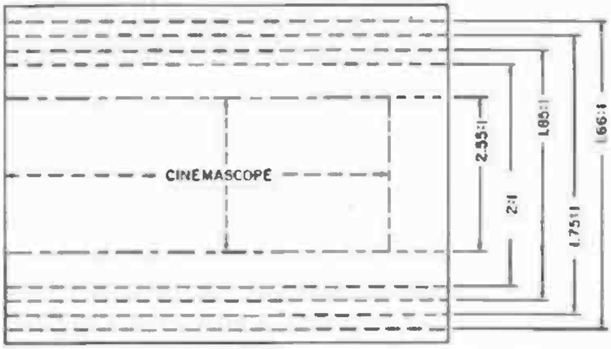


Fig. 19-46. Aspect ratios used for wide-screen motion picture projection.

the light intensity must be increased four times to obtain a picture having a brightness equal to the original. A typical example would be replacing a lens of 5-inch focal length with a 2.5-inch lens.

19.48 What does cropping a picture mean?—To block off the top and bottom by changing the ratio of the aperture plate in the projector to obtain a given ratio for wide-screen projection. Cropping is illustrated in Fig. 19-48.

19.49 What does the term "throw" mean?—It is the distance from the center of the projector lens to the center of the screen. This is sometimes called the front focus.

19.50 How can the picture size for a given lens size be calculated?—The picture size equals,

$$\frac{\text{Throw} \times \text{Width of aperture}}{\text{Equiv. focal length of lens}}$$

19.51 For a lens of given size, how is the distance from the screen to the projector calculated?—The throw eqnals,

$$\frac{\text{Width of screen} \times \text{Eqnivalent focus}}{\text{Width of projector aperture}}$$

19.52 How is the proper size lens calculated for a given size picture?

$$\frac{\text{Throw} \times \text{Width of aperture}}{\text{Width of picture}}$$

19.53 What is the standard for motion picture screen luminance?—The Standard for 16-mm motion picture screen projection luminance is given in USASI (ASA) PH22.100-1967, and for 35-mm indoor projection in USASI (ASA) PH22.133-1965. However, in the absence of such standards, the following general conditions should prevail.

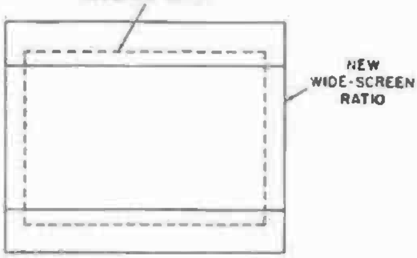


Fig. 19-48. A standard Academy aperture cropped by means of a cropping aperture in the projector.

Studio review rooms are used for screening the studio product; therefore, they must meet certain standards of the industry if the final product is to be satisfactory. Also, review rooms are used by the producer, the director, the cast, the camera, sound, and lighting crew, and the editorial staff for reviewing their work. This requires that the light and sound levels be standard (in most instances they are checked daily).

The measurement of screen luminance is made daily. The projection machine is in complete operation, but there is no film in the aperture. The photometer used for measuring screen luminance is to have the sensitivity of an average observer, as specified by the International Commission on Illumination. Such meters are available from theater equipment supply houses. The acceptance angle of the meter must be small and must accept the reflected light from the screen area no larger than a circle whose diameter is 10 percent of the screen width.

For both 16- and 35-mm projection the luminance at the center of the screen must be within 16 foot-lamberts, plus or minus 2 foot-lamberts, or 55 nits plus or minus 7 nits, and is to be uniform over the area observed. (See Question 19.5.) Luminance at a distance of 5 percent of the screen width from the side edges of the screen on a horizontal axis is to be 80 percent of the center luminance, plus or minus 10 percent of the center luminance. Because of other differences between the Standard for 16- and 35-mm projection, the reader is referred to the Standards.

Studio review rooms generally are designed to seat about 25 to 40 people. Therefore, the standard viewing area is to be within 15 degrees of each side to the center of the screen, in both the horizontal and vertical planes, and within three picture heights plus or minus 1. To meet these conditions, the projection machine must project the picture close to the center axis of the screen in both vertical and horizontal planes to reduce the effect of keystoning to a negligible amount. Keystoning in the horizontal plane must be kept to an absolute minimum. (See Question 19.54.)

The Standard for indoor theater screen luminance is given in USASI (ASA) PH22.124-1961. For drive-in theaters, the Standard is given in the SMPTE recommended practice 12, July, 1962.

The term "nit" may be converted to screen luminance foot-lamberts as follows:

1 nit = 0.2919 foot-lamberts

1 foot-lambert = 3.426 nits

It should be kept in mind that although the Standards specify that screen brightness (luminance) is measured with no film in the projector aperture, the average luminance will be considerably decreased with film running in the aperture.

19.54 What does the term keystoning mean?—It is seldom that the projection machines in a theater or a review room are on the same level as the screen. As a rule, they are considerably higher than the center of the screen. This difference in height causes what is known as keystone distortion, because the shape of the picture on the screen resembles an inverted keystone. In Fig. 19-54 is shown how keystone distortion is developed. The projection machine is shown at A, the screen at B, the top of the picture at C, and the bottom of the picture at D. The line is drawn perpendicular to the screen through the center line of the picture. It will be noted that the distance from A to C is less than the distance from A to D. Therefore, the light rays traveling from A to D will diverge farther than those traveling from the projector to the top of the picture. The result is that the picture is projected in the shape of an inverted keystone, wider at the bottom than at the top. AEF is the angle of projection, and AE is the axis of projection. The projection angle is the angle formed by the projection axis and an imaginary line drawn through the center of the picture perpendicular to the screen. The greater the angle formed, the greater the keystone distortion. Keystone distortion may also be caused by the offset of the projectors in a horizontal plane. The effects are about the same. The effect of horizontal keystoning changes as the picture is changed from one projector to the other. With the advent of wide-screen projection, keystoning has become more noticeable because of the greater magnification and spread of the picture.

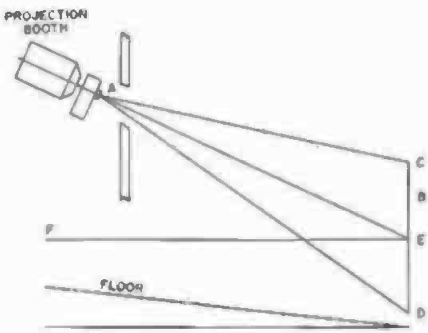


Fig. 19-54. Keystoning of a projected picture due to a high angle of projection.

Keystoning may be offset somewhat by tilting the top of the screen backward in an attempt to equalize the throw between the top and the bottom of the screen. The use of curved screens to compensate for these effects is of some help; however, keystoning cannot be entirely eliminated, because of the relationship between the projectors and the screen. To obtain a square frame line and eliminate the distortion of the aperture outline on the screen, the metal aperture in the picture head is filed to obtain a square frame line on the screen. Also, the screen is masked at the sides, top, and bottom into a small amount of the picture, to help clear up the effects of keystoning.

In the design of studio review rooms, keystoning must be kept to an absolute minimum. If possible, axes of the projectors lens should be above 8 feet from the floor level to clear persons standing at the rear of the room. The screen should be installed in such a manner that the vertical keystone is not more than a maximum of 5 degrees. The machines should be set as closely together as operation will permit to reduce the horizontal keystone effect to not more than 6 degrees. These factors are important, as such rooms are used to judge quality and composition of a picture. If the keystone is excessive, distortion and out-of-focus conditions will prevail at the edges of the picture.

19.55 What is the angle before keystoning becomes objectionable? — The maximum in the vertical direction is about 12 degrees. In rerecording stages and review rooms, if possible, the projection room should be set high enough for the beam to clear persons standing in the rear of the room. Also, the beam should be as near on the center axis of the screen in the vertical plane as possible. The keystone in the horizontal plane should be kept to a minimum. (See Question 19.54.)

19.56 How is a motion picture screen evaluated? — A motion picture screen in which the loudspeaker system is to be placed behind the screen must be selected with care, as considerable loss of high-frequency reproduction is possible. The manufacturers' data sheets should be consulted in advance to determine the loss at frequencies above 6000 Hz. The loss at 10,000 Hz should not be greater than 4 dB, with respect to 1000 Hz. Along with the transmission characteristics, the reflective qualities must also be taken into consideration. (See Questions 19.5 and 23.187.)

19.57 What type screen is used for 3-D projection?—An aluminized surface. The audience views the picture through polarized glasses. Flat or beaded surface screens will depolarize the light, thus destroying the three-dimensional effect. The analglyphic system of projecting 3-D pictures supplies the audience with red and green filter eye glasses. Both these systems are now obsolete and no longer used.

19.58 Describe the different type motion picture screens in use. — The three most commonly used screens are the matte white, beaded surface, and lenticulated. Matte white screens are employed with front type projection systems and may be a solid or perforated (sound) surface. Images projected on this latter type screen diffuse the projected light evenly in all directions. About 15 percent of the incident light is lost. However, viewing an image from far off-side results in considerable distortion of the image.

Lenticulated screens are constructed of minute horizontal and vertical reflective areas in the form of diamonds or rectangles. The size of the lenticulation affects the viewing angles and brightness. Such screens are capable of providing images several times brighter than those shown on white matte screen. Lenticulated screens are not perforated for sound.

Beaded screens are quite popular for 16-mm home projection, halls, and auditoriums. They may be considered to be a type of lenticulation accomplished by surfacing the screen with small glass beads approximately 0.5 to 0.1 mm in diameter. This results in a bright image but a narrow viewing angle. If used in a high ambient light, the brightness of the image is reduced. These screens are not perforated.

Screens used for rear projection consist of glass, Vinyl-latex plastic, and acrylic plastic and are used for audiovisual displays, with the projection system placed at the rear of the screen. The surface of the screen is not perforated.

Black lenticulated screens, developed by Sasuke Takahashi of Japan, are unique because of the black lenticulated vertical and horizontal surface. It is claimed by the inventor that in comparison with a matte white screen, it is 25 times greater in black and white contrast and ideal for color projection. Also, it absorbs scattered stray ambient light, and prevents halation on the surface with wide-angle viewing. It may also be used for stereographic projection. With no light on its surface, the screen appears black. This is advantageous for some types of display. Because of the lenticulation, the surface is not perforated for sound.

19.59 What is the process of background or rear projection?—A method of projecting background scenes on a transparent screen from the rear. It is used in both motion picture and television production. The actors and props are placed in front of the screen and the scene is photographed as a whole. If properly lighted, the finished picture appears as if the scene had been shot on location. To provide synchronization between the camera and projector shutters, they are driven from a selsyn-interlock system. Before each take the shutters are accurately aligned to prevent shutter-drag, blank spots, and flicker in the photography. Selsyn distributors are discussed in Section 3.

19.60 Describe a 16-mm television projector.—Projectors used for television require special features not found in theater projectors. Such a projector manufactured by RCA is shown in Fig. 19-60A and has several features making it peculiar to television, with complete remote and automatic control of its various functions. A rear view is given in Fig. 19-60B.

Earlier television projectors did not use an intermittent motion, but employed special shutters or an electronic shutter similar to a strobelight, synchronized with the power-line frequency supplying the station. In this RCA machine, an intermittent motion is used, having a precision claw-type movement, with a high degree of film registration. A three-toothed claw pulls the film downward. The center claw is sapphire lined, with the other claws made of hardened steel. The film speed is 24 frames per second, with a shutter speed of 720 rpm supplying 60 light pulses per second.

Starting time is 0.3 second. The projector can be run in reverse for rehearsals, thus eliminating the need for rewinding and rethreading. It may also be used to project a single picture for extended periods of time. A neutral filter is automatically inserted in the light path to prevent burning of the film with the shutter held open. Metallic cue marks may be attached to the film for automatically stopping and inserting a commercial and returning to the program again. The motor turns at 1800 rpm. Other features are an automatic

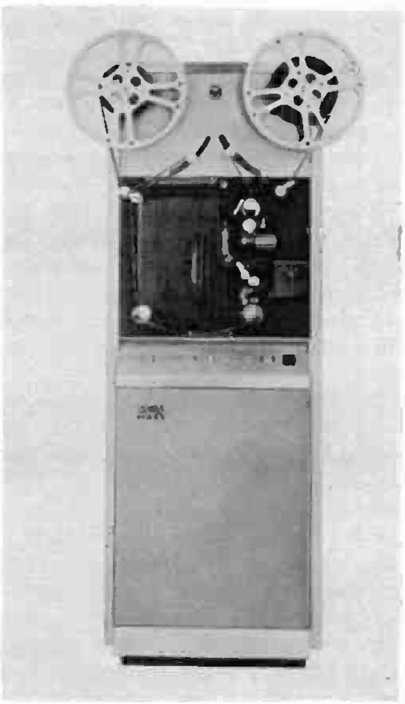


Fig. 19-60A. RCA Model TP66 16-mm television projector.

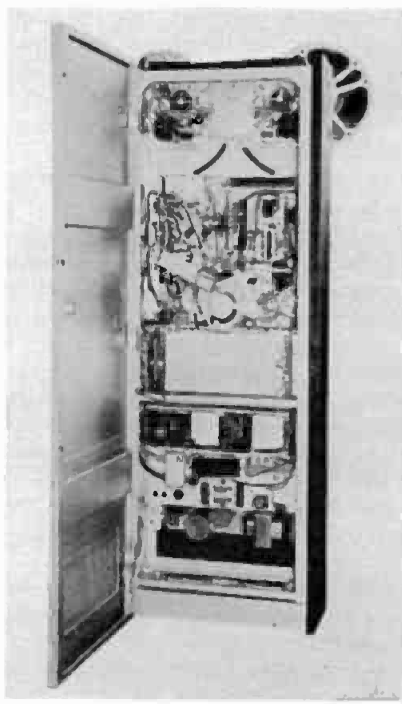


Fig. 19-60B. Rear view of RCA TP66 16-mm television projector.

loop restorer, projection lamp, and sound exciter lamp change. The electronics for the sound are completely transistorized.

In the earlier type projectors special shutters were employed which permitted the light to reach the film for $1/200$ second every $1/6$ second. Thus the first frame was projected on the camera tube twice, and the second frame three times, the third frame twice, and the fourth three times. This was an average scanning of each frame 2.5 times. Scanning the picture in this manner, 24 frames times 2.5 results in a television image of 60 fields or 30 frames per-second. This permitted the television to lock in with the power frequency, reducing flicker and other objectionable effects.

Television projectors also make use of electronic shutters, similar to a strobelight, which are synchronized with the power frequency. The light is out during the pull-down period, thus no shutter is required. The film in this type projector runs at a constant linear speed and requires no intermittent movement.

19.61 Describe test films available for optical and magnetic sound recording and reproducing equipment.—Test films for optical and magnetic sound reproducing equipment are many and varied. A few of the more important are: Theater test film containing dialogue music and sound effects; multifrequency films for frequency response measurement; sound focus; azimuth; and flutter. Such films are available for 8-, 16-, 35-, 55- and 70-mm equipment, and may be obtained from the SMPTE.

For those interested in video tape recording and reproduction, recommended practices and Standards are also available from the same source.

1962. Describe a Standard theater test reel.—A special test film containing picture and sound track from major studio productions. Each sound track has been selected for its special characteristics and quality of reproduction. This film is used by theater service organizations for adjusting the reproducing characteristics of a theater sound system. If the frequency characteristics of the system have been properly adjusted for a particular theater, good sound reproduction will result when this film is played on the system.

The film also contains a piano recording for checking flutter and heavy sound effects to determine the power-handling capabilities of the system. This film may be obtained in both 16- and 35-mm from the SMPTE.

19.63 Describe a flutter film.—Flutter films are used for measuring the amount of irregularity in the transport system of a magnetic or optical sound-reproducing machine. The flutter of this film must be extremely low, or it would be of little value. Flutter films made for measuring motion picture sound-reproducing equipment have slightly different Standards, depending on the type of equipment with which they are to be used. Therefore, the reader is referred to USASI (ASA) Standards PH22.43, 22.98, 22.113, and 22.128. Flutter films may be obtained from the SMPTE. (See Question 17.144.)

19.64 What is a buzz track?—A buzz track is a test film containing two square-wave frequencies used for the adjustment of the lateral placement of an optical film, while running through a sound head in a motion picture projector. The sound track consists of an 0.087-inch opaque center with a frequency of 1000 Hz on one edge and 300 Hz on the other. The positions of the sound tracks are accurately located on the film, so that when they are run in a projector sound head, no sound will be heard from either track, if the lateral adjustments of the film guide are correct. If the film is out of placement or the film weaves in its motion while running, one of the two tracks, or both, will be heard. The film guides are adjusted for a no-sound position. Such test films may be obtained for both 16- and 35-mm use. The standard for this film is given in Fig. 19-64.

19.65 What is a sound-focusing film?—A test film used for focusing the optical system of a photographic sound-reproducing head. USASI (ASA) Standard PH22.61-1963 specifies that the print is to be made from an original negative with a frequency of 7000 Hz. The sound track may be either variable-area or variable-density. The azimuth of the sound track is to be within plus or minus 3 minutes of arc. Such films may also be obtained for 16-mm projectors; however, for 16-mm, the sound track is made in a different manner, as discussed in Question 19.74.

American Standard Specification for

Buzz-Track Test Film for 35mm Motion-Picture Sound Reproducers, Photographic Type

ASA
Reg. U.S. Pat. Off.
PH22.68-1962
Revision of PH22.68-1949
*UDC [illegible]

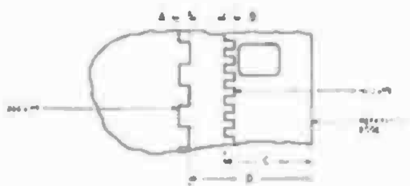
1. Scope

This standard describes a film that may be used for checking the lateral-scanning slit placement of photographic-type 35mm motion-picture sound reproducers.

2. Test Film

2.1 The test film shall be a direct positive recording or a print from an originally recorded negative and shall contain 300-cycle and 1000-cycle square-wave tracks on either side of the central exposed strip, as shown in the diagram.

2.2 The central exposed strip and the exposed portion of the two signal tracks shall have a minimum density of 1.4 and a maximum density of 2.0.



Dimensions	Inches	Millimeters
A	0.012 min	0.30 min
B	0.007 min	0.18 min
C	0.201 max 0.200 min	5.10 max 5.08 min
D	0.289 max 0.287 min	7.34 max 7.29 min

2.3 The film stock used shall be cut and perforated in accordance with American Standard Dimensions for 35mm Motion-Peiture Positive Raw Stock, PH22.36-1954.

2.4 The film stock used shall have a shrinkage of not more than 0.50 percent.

3. Revision of American Standard Referred to in This Document

When the following American Standard referred to in this document is superseded by a revision approved by the American Standards Association, Incorporated, the revision shall apply:

American Standard Dimensions for 35mm Motion-Picture Positive Raw Stock, PH22.36-1954

NOTE: A test film in accordance with this standard is available from the Society of Motion Picture and Television Engineers.

Approved April 25, 1962, by the American Standards Association, Incorporated
Sponsor: Society of Motion Picture and Television Engineers

* Universal Decimal Classification

Printed in U.S.A.
ASA3/[illegible]

Fig. 19-64. USASI (ASA) Standard for 35-mm buzz track.

19.66 What is a level-balancing loop?—It is a test film, recorded at 1000 Hz and then made into the form of a loop. Such loops are used for adjusting the output between projectors, or any type sound reproducers. In the studio, if the loop is magnetic, it is generally recorded at a level representing 100-percent modulation.

For optical sound track, maximum output is limited by the maximum modulation that may be applied to the light modulator. Therefore, to assure that the track will not be overloaded, it is recorded at 80-percent modulation. The maximum deviation from the established level for such sound tracks is plus or minus 0.25 dB.

19.67 What is a visual test film?—A special type print (picture only) containing four targets to check the focus and alignment, travel ghost, jump and weave, and lens abberation. This film is used when installing new equipment or making maintenance checks.

The focus-alignment target shows if the picture size and the screen masking are correct and the projected picture is properly centered on the screen. The travel-ghost target will show improper timing of the projector shutter and will give an indication of the correct adjustment as the timing is being corrected. The jump-and-weave target gives an accurate indication of the unsteadiness of the projected picture. Picture jump is measured in percent of picture height. Picture weave is measured in percent of picture width. The lens-aberration target shows picture distortion and gives an indication of the lack of sharpness that will be present in pictures run on a particular projector.

19.68 Describe multifrequency test films.—Multifrequency test films consist of a series of frequencies, which range from 30 Hz to 12,000 Hz. They are recorded on either magnetic or optical film, for both 16- and 35-mm use. Each print is individually calibrated. Generally, the frequencies are preceded by a voice announcement. The amplitude variation of such films is on the order of 0.25 dB. After a measurement has been made, the absolute response may be ascertained by the use of a calibration sheet supplied with the film. Such films are also called constant-amplitude or constant-frequency films.

19.69 What is a sound-transmission film?—A film similar to that described in Question 19.68, except it contains fewer frequencies than the multifrequency film. It is used by theater service organizations.

19.70 What is a warble film?—A film used for making acoustical measurements of an auditorium. This is accomplished by reproducing the film on the normal projection system and measuring the acoustical output for each frequency, using a sound-level meter as described in Question 22.94.

The sound track contains frequencies from 40 Hz to 8000 Hz. Each frequency is warbled plus or minus 12.5 percent of the nominal frequency. Above 1000 Hz, the frequencies are warbled 125 Hz plus and minus the nominal frequency. The warble rate is varied from 2.5 Hz to 5 Hz per second. The purpose of warbling the frequencies is to prevent the formation of standing waves in an auditorium or enclosure while making the measurements.

19.71 What is a scanning-beam illumination film?—A sound track used for checking the uniformity of illumination across the scanning slit in a sound head. The sound track consists of 17 incremental 1000-Hz tracks, all of an equal amplitude of approximately 0.007 inch. The tracks appear on the film in succession, each preceded by an announcement identifying the track number. The 17 tracks cover a width greater than the standard scanning beam. By running this test film and observing the indications on an output meter, it is possible to correct unevenness of illumination in the optical system and bring the variation in output to within plus or minus 1.5 dB, which is the recommended maximum variation. The adjustment is accomplished by replacing the exciter lamp or adjusting its position relative to the scanning slit. (See Question 19.72.)

19.72 What is a snake track?—A 1000-Hz sound track with an 0.007 amplitude and placed on the film in such a manner that the track moves across the scanning slit from one edge to the other at a uniform rate. The output from the film is measured using an output or VU meter, and adjustments of the exciter lamp are made to compensate for the variations. For 16-mm projectors, the

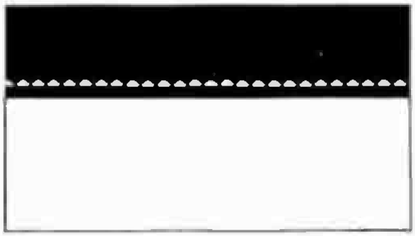


Fig. 19-72. A 35-mm snake-track for measuring the variation in slit illumination of a sound-reproducer optical system.

amplitude of the signal is 0.005 inch. A sample of a typical snake track is shown in Fig. 19-72. (See Question 19.71.)

19.73 What is a 16-mm projector-resolution target?—A glass slide 1 inch by 2 inches containing 19 resolution test patterns photographed within the standard 16-mm projector-aperture area. This slide is used in determining the resolving power of projector lenses in terms of the number of lines per millimeter which are resolved.

19.74 How does a 16-mm sound-focusing film differ from a 35-mm focusing film?—For 16-mm sound focusing, a special square-wave sound track is used because its output changes more rapidly with changes of focus than the output from the conventional sine-wave sound track. The square wave gives a more sensitive indication of errors of azimuth adjustment. The frequency used is 7000 Hz.

19.75 What is a 3-D and a 2-D projector-alignment film?—This film contains a special target for visually aligning projectors used for 3-D and 2-D projection. The film is made up in loop form. The projectors are aligned physically so that projected images fall on top of each other while both machines are running. It is absolutely essential that both projectors be perfectly aligned when projecting 3-D pictures, as both projectors are used simultaneously.

19.76 How are multiple magnetic sound head clusters tested?—By the use of multiple sound-track test films. These films are available in a variety of different types covering many tests and are somewhat similar to those used for the testing of photographic sound-track heads. Magnetic multifrequency test films have considerable greater frequency range than optical test films.

19.77 What is the height of the optical slit used in a motion picture projector sound head?—From 1.2 to 1.5 mils. A larger slit height is employed in a theater sound head because the frequency response is not as great as that required for rerecording sound heads. The attenuation characteristics for a 1.2-mil slit are shown in Fig. 19-77 as compared to the attenuation of a 0.5-mil slit.

19.78 Give the average frequency response for motion picture theaters.—For reproduction from magnetic sound track, the reproducing frequency characteristic is generally as shown in Fig. 19-78A. However, in some theaters a slight tilt up or down is required at the low- and high-frequency ends to compensate for theater acoustics. The amplifier system is equalized by using a resistive termination in place of the speaker system.

Reproduction from photographic sound track is quite different from that of magnetic film, as the high-frequency end is rolled-off in a manner similar to that shown in Fig. 19-78B. Theater amplifier systems are so designed that when switching from one type sound track to the other, the equalization is taken care of automatically.

Further extension of the frequency response for optical sound track beyond 6500 Hz is undesirable because of the decreasing sensitivity of the human ear, the small amount of energy in the higher frequencies, and the increased background noise from the film. Theaters that have attempted to widen the response for optical film, even up to 8000 Hz, have returned the response to the 6500-Hz cutoff because of the increased background noise from the film, the andience, and the air conditioning.

Most studios, during the time of transfer of the master sound track to the release print, tilt up both the low- and high-frequency ends to increase the signal-to-noise ratios at these frequencies. Therefore, the theater reproducing characteristic should be uniform, except for any correction required for theater acoustics.

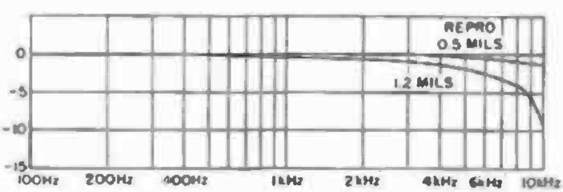


Fig. 19-77. Attenuation of 0.5-mil slit compared to that of a 1.2-mil slit.

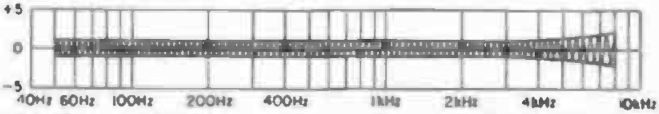


Fig. 19-78A. Suggested frequency response for magnetic sound track reproduction.

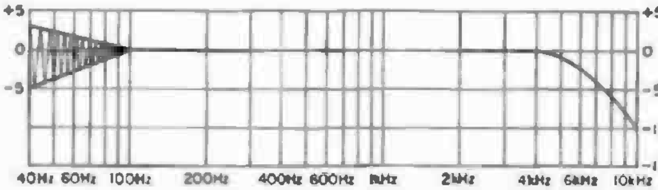


Fig. 19-78B. Frequency response for reproducing 35-mm optical sound track.

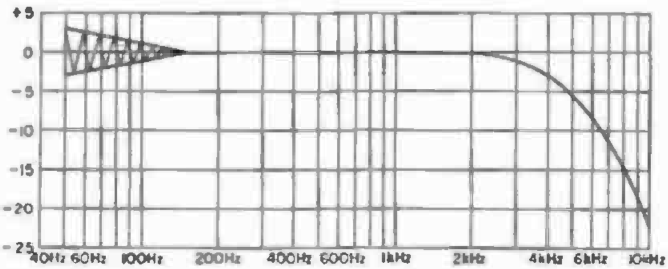


Fig. 19-79. Typical frequency response of a 16-mm projector with the controls set for a flat frequency response.

After the frequency response has been set, a Standard Academy test reel (see Question 19.62) is used to check the final response. Readjustment of the low and high frequency is made, if necessary.

19.79 What is the recommended frequency response for 16-mm projector systems?—Because of the wide variation in the frequency characteristics of 16-mm recordings and the conditions under which the films are projected it is difficult to make recommendations. However, the frequency response shown in Fig. 19-79 is that generally found when measuring 16-mm projectors, with the low- and high-frequency controls set to the flat position. For good reproduction, the high-frequency control should be adjusted to give presence and intelligibility. The low-frequency control should be adjusted for good low-frequency reproduction without its being boomy. The local acoustic environment will have a considerable effect on both the control settings.

19.80 Describe the principal components of a theater-type motion picture projector.—Pictured in Fig. 19-80A is a 35/70-mm theater-type motion picture projector, manufactured by Cinemeccanica of Milan, Italy. This machine is designed for both magnetic and optical sound-track reproduction. The major

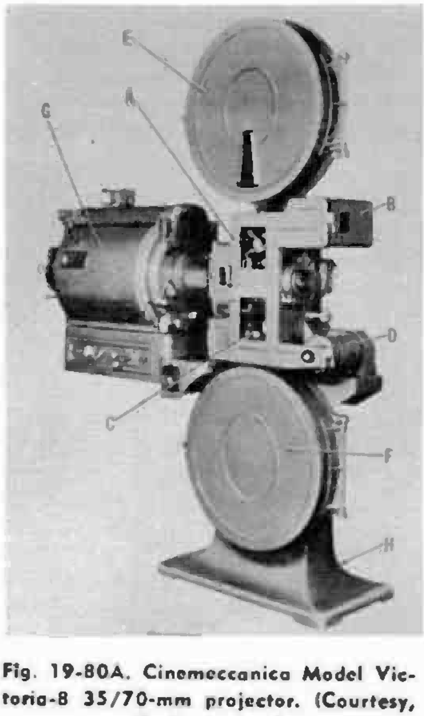


Fig. 19-80A. Cinemeccanica Model Victoria-8 35/70-mm projector. (Courtesy, Carbons, Inc.)

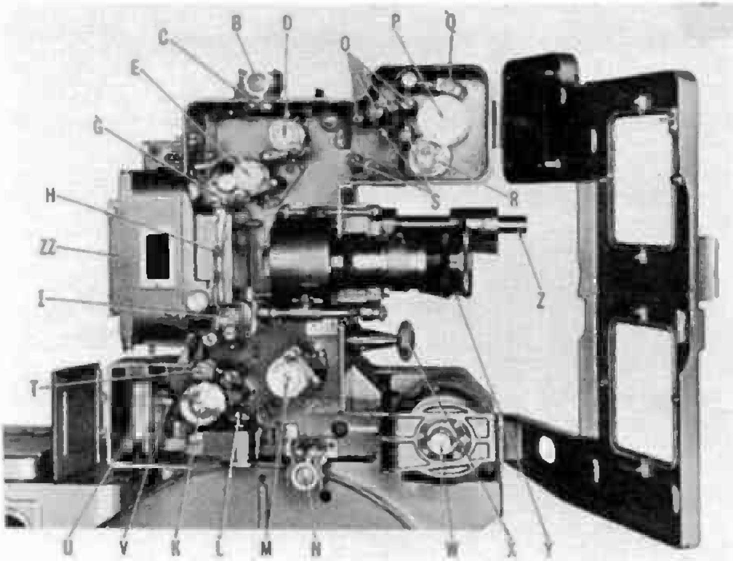


Fig. 19-80B. Interior view of Cinemeccanio Victoria-8 35/70-mm projector showing the penthouse, picture head, and optical film sound head. (Courtesy, Carbons, Inc.)

components are the picture head A, magnetic head compartment B, optical sound head C, drive motor D, upper and lower magazines E and F, lamp house G, and the base H. The whole assembly is so mounted on the base that it may be tilted upward 5 degrees, and downward 18 degrees. This pictured model is equipped with an XeTron lamphouse, rather than the conventional arc light. (See Question 19-105.)

Fig. 19-80B shows that the interior of the magnetic head compartment, the picture, and the optical sound heads are exposed. If the threading path of the film shown in Fig. 19-80C is referred to, a somewhat clearer understanding of the various components shown may be had. Combination 35/70-mm sprockets are used to facilitate changing from one size film to the other. The film sprocket idlers are turned over 180 degrees for the desired film size.

Starting at the top, the film A passes over magazine roller B, and through two fire-trap rollers C, to a constant-speed pull-down sprocket E, with its two idler rollers. Sprocket D is not used

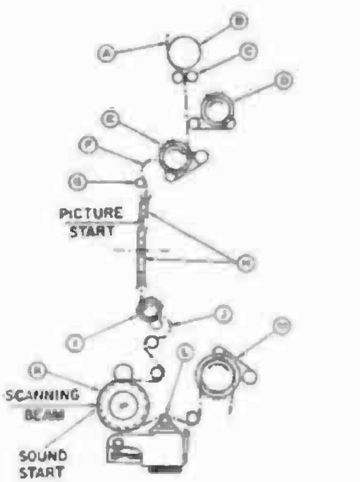


Fig. 19-80C. Film-threading path for 35-mm film using optical sound track reproduction.

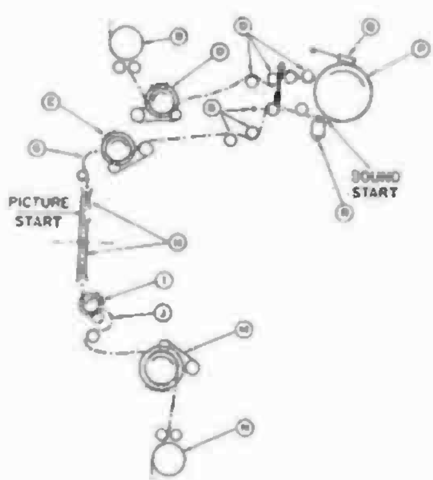


Fig. 19-80D. Film-threading path for 35- and 70-mm film using magnetic sound track reproduction.

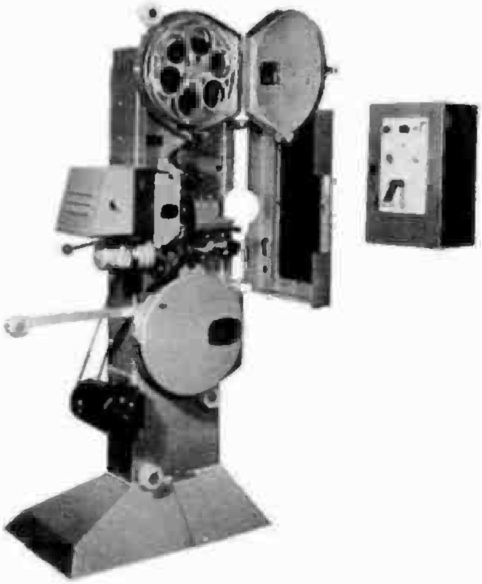


Fig. 19-80E. Magna-Tech Electronic Co. Inc., Model PJ-35, 35-mm projector.

with optical film. A loop F, about 17 perforations long, is made in the film and threaded over roller G above the picture-aperture plate H. The film is now pulled downward by intermittent sprocket I (see Question 19.86). A second loop J, of about 15 perforations long, is made before the film is passed over sound impedance drum K, then over an oil-filled damper F to constant-speed sprocket M, and to a magazine roller and fire-trap in the lower magazine. The sound start mark is placed over the impedance drum K in the exciter lamp light beam, and the picture start frame is placed in the picture-aperture plate H. The same callouts are used for Figs. 19-80C, D, and E.

For 35- and 70-mm screening using magnetic sound track, the threading path differs from that for optical film and is shown in Fig. 19-80D. After the film leaves the fire-trap rollers C, it passes over constant-speed sprocket D, then over a group of guidance and tension rollers O, to impedance drum P with its pad roller Q, over magnetic head cluster R. This cluster has in reality 10 magnet heads, 4 on one side and 6 on the other, and is rotated to the one desired. A second group of rollers S directs the film back to sprocket E, roller G, and to aperture plate H. For 70-mm film operation, loops F and J are decreased to 13 and 11 perforations, respectively. Leaving the aperture plate, the film is looped at J and passed directly to constant-speed sprocket M and down to the lower magazine through fire-trap rollers N.

Fig. 19-80B shows that there are additional items still to be considered. These are the pad roller T over the optical impedance drum K, the exciter lamp housing U, the optical lens system V for optical sound reproduction, drive motor W, framing control X, projection lens Y, and the shutter shaft Z. The housing ZZ at the rear of the picture head encloses a two-bladed conical picture shutter rotating at 1440 rpm and a safety shutter. Because the picture shutter cuts the light beam very close to the film, the efficiency is close to 50 percent for 70-mm film.

The motor drive system is designed for both 24- and 30-frame projection, by the use of a double pulley on the motor shaft. Because of the close tolerance and design of projection machines, a force-feed continuous lubricating system is necessary. This machine makes use of a gear pump, which forces oil to each gear, bearing, and intermittent movement. In addition to the above items, forced air and water cooling is provided around the picture gate area and the shutter housing. The upper and lower magazines will each accommodate 5900 feet of film. Because of the stabllizers used with both the optical and magnetic transport systems, the total flutter is less than 0.10 percent.

When two or more projectors are mounted in a projection room, some means of automatically cutting over from one to the other must be provided. The control system must change over the picture and sound simultaneously, arc lamp excluded. For Cinemeccanica installations, this is accomplished by the installation of a microswitch at the armature of the changeover solenoid douser (See Question 19.42). The microswitch operates a group of changeover relays for the optical sound or the 4 or 6 magnetic sound tracks. A pushbutton at the lamp house provides the control. This changeover system will provide facilities for three machines.

Pictured in Fig. 19-80E is a Magna-Tech Corp., 35-mm reversible projector designed particularly for dubbing for small studio operations. Basically, the projector is a Phillips-Norelco Model FP-20 35-mm projector, with several modifications.

This machine is equipped with a dual-purpose 220V 3-phase sync-interlock motor, upper and lower magazine torque single-phase motors, automatic gate-release mechanism, and an on/off control circuit for the lamp douser. The light source may be incandescent, arc, or a xenon type. When running interlock, all of the principal functions take place automatically. The lamp turns on and off whenever the projector is run in the reverse direction. Torque motors on the upper and lower magazine spindles provide rewinding, take-up, and feed for running in the reverse direction.

The remote-control features make this machine particularly desirable for dubbing operations. The projector can be threaded and the film can be run down into a reel. Difficult cues can be rehearsed by running the machine with the film in sync with the reproducers and recorders, both forward and backward. A satisfactory rehearsal can result from these operations. The film is then run back to the head end and rolled forward for a take. If the cue is missed, it may be stopped in sync and rolled back again. Regardless of where the projector is stopped, it will always remain in sync with the balance of the rerecording equipment, unless removed from the distributor line. Generally, for this type operation a xenon or incandescent-type projection lamp is used. A control box on the wall provides switching for different size reels of film, forward or reverse, or remote control of these functions. Rollers are provided above and below at the rear of the machine for loop operation (see Question 17.223). Reversible distributors are discussed in Question 3.56.

19.81 Describe a preview magazine and its use.—A preview magazine is generally a triple reel, lower magazine, which replaces the regular single magazine on a projection machine. It permits the screening of a picture with a separate sound track.

Preview magazines originated in Hollywood to be used in preview theaters for running sneak previews of pictures not yet released. The purpose of this was to get the audience reaction before making the final cut.

Such magazines, using separate sound tracks and picture for the editorial department, are also installed on machines in the studio for running dailies and rough cuts.

When these magazines are used, the picture print carries no sound track, since the sound track is on a separate magnetic film. Several different types of such magazines are available to fit projectors of different design and threading paths. The magazine shown in Fig. 19-81A has been selected for its simplicity and ease of explanation. This magazine is designed for use with a sound head where the optical and magnetic sound pickups are contained in the same compartment.

The threading path for a typical preview magazine of the design illustrated is given in Fig. 19-81B. Starting with the film A as it leaves the intermittent sprocket J, it is threaded over guide rollers B and C in the sound head, then to the take-up reel D in the magazine. As this is only the picture, the usual threading path over the sound drum K is omitted. The sound track is fed from supply reel E at the bottom, through chute F, over rollers G, H, and I, to intermittent sprocket J, over the sound drum K, to rollers B and C, to the take-up reel L. With this type arrangement, the sound and picture is kept in synchronization.

19.82 What is a film gate?—The mechanism that holds the film against

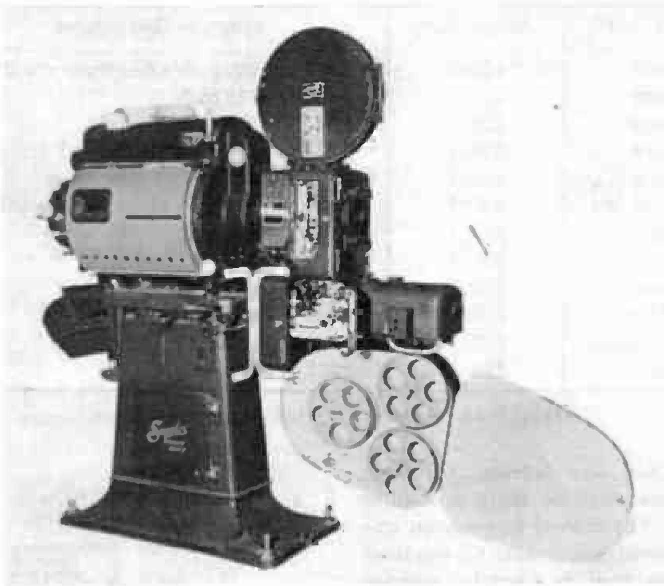


Fig. 19-81A. Preview magazine mounted on a Simplex 35-mm motion picture projector.

the aperture plate in the picture head of a projection machine.

19.83 Describe a projector aperture plate and its purpose. — The aperture plate consists of a rectangular hole in a removable plate that may be changed while the projector is in operation, if necessary. It is inserted in the picture head to establish the correct picture image ratio before the image arrives at the projection lens system.

Because of the many aspect ratios and different sizes of film, a number of plates are required for the average theater. Various size aperture plates are given in the tabulation in Fig. 19-83, along with the film size normally used.

Other ratios are: 1.66:1, 1:75:1, and 1.85:1. If the keystone is excessive, it may be corrected by filing the opening of the aperture plate to obtain a rectangular image on the screen. When the projection angle is greater than zero degrees downward, an undersized aperture plate is used and filed out to fit the particular projection angle. The bottom of the undersized plate is filed to the correct horizontal dimension to obtain the necessary width at the top of the picture. If the projection

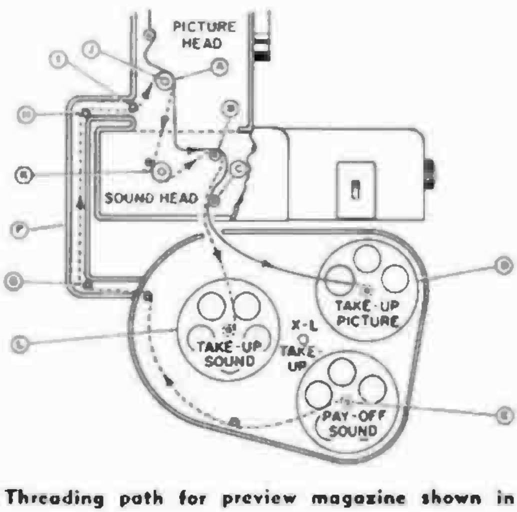


Fig. 19-81B. Threading path for preview magazine shown in Fig. 19-81A.

Film Width	Aspect Ratio	Aperture Dimensions
35 mm	1.34:1	0.825 × 0.600 inch (Academy standard)
35 mm	2:1	0.839 × 0.715 inch
70 mm	2.21:1	1.913 × 0.866 inch
35 mm	2.21:1	0.839 × 0.715 inch
70 mm	2.21:1	1.913 × 0.866 inch
3 × 35 mm	2.27:1	0.985 × 1.088 inch (Ultra-Panavision)
70 mm	2.27:1	1.913 × 0.866 inch
35 mm	2.34:1	0.839 × 0.715 inch
35 mm	2.35:1	0.839 × 0.715 inch
55.6 mm	2.55:1	1.340 × 1.050 inch
3 × 35 mm	2.77:1	0.985 × 1.088 inch (Cinerama)
70 mm	2.94:1	1.913 × 0.811 inch

Fig. 19-83. Aperture plates of various sizes.

angle is minus zero degrees (projector shooting upward), the filing procedure is reversed. For curved screens, an undersized aperture will also be required and must be filed in a similar manner in order that the top and bottom will appear horizontal and parallel to each other on the screen. After the inside edges are filed, they are beveled about 30 degrees, with the sharp edge toward the film, to eliminate fringe effects at the edges of the picture. The bevel is then painted dead black or black anodized to prevent reflections. (See Question 19.54.)

19.84 What is an electronic shutter?—A special electronic control device connected to the photocathode of an image-orthicon tube used for the transmission of motion pictures by means of television. The control device keys the image-orthicon tube in such a manner that it operates only during the vertical blanking period. The film travels through the projector at a continuous linear speed, even during the pull-down period, and this electronic device replaces the conventional mechanical shutter. (See Question 19.60.)

19.85 What is a rotary stabilizer?—A hollow drum similar to a flywheel and filled with oil. It is used on motion picture projectors and sound heads. The stabilizer is attached to the film-drum (impedance-drum) shaft. When the film is pulled over the drum by a constant-speed sprocket, friction of the film on the surface of drum causes it to rotate.

The outer shell of the stabilizer is made of thin metal so as not to impose a heavy load on the impedance drum when it first starts to rotate. Inside the light shell is a heavy flywheel mounted

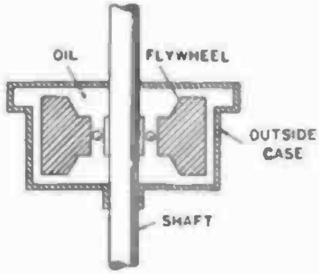


Fig. 19-85. Cross-sectional view of RCA rotary stabilizer.

on ball bearings. The inside of the stabilizer is completely filled with oil. When the outer shell begins to rotate, it transmits its motion to the flywheel through the oil. After attaining its normal speed, the stabilizer tends to stabilize and filter out variations in the linear speed of the film. A cross-sectional view of a rotary stabilizer developed by RCA and used on their projectors and sound heads is shown in Fig. 19-85.

19.86 What is an intermittent movement? — A mechanical movement used in a motion picture projector to pull the film downward intermittently at a given rate of speed. 35-mm sound projectors pull the film downward at a linear speed of 90 feet per minute, or 24 frames per second.

While the film is in motion, a mechanical shutter blanks out the light from the lamp house and hides the downward motion from the viewer. Due to the persistence of vision of the human eye, the projected series of pictures appears to have continuous motion.

19.87 How does an intermittent motion operate?—A Geneva intermittent movement is shown in Fig. 19-87A. The

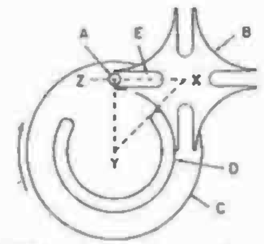


Fig. 19-87A. A 4:1 Geneva intermittent movement used in 35-mm motion picture projectors.

movement consists of four parts: a pin A, a start or cross B, cam C, and ring D.

If point X, the center of the star (connected to the intermittent sprocket), and point Y, the center of cam C, are continued to the center Z of the pin A, a right angle is formed. The arrow indicates the direction of travel of the cam. Pin A is shown at the exact instant it starts to engage the slot in the star B. Until cam C has reached this position there is no movement of the star.

At the instant pin A engages the slot E, the pin travels along the line ZX. Therefore, the pin will enter the slot cleanly with no chatter along the side walls of the slot. Observation will show that at the exact instant the pin enters the slot there is still no movement of the star, inasmuch as the motion of the pin coincides with the slot. After the pin passes completely into the slot, its motion no longer coincides with the slot

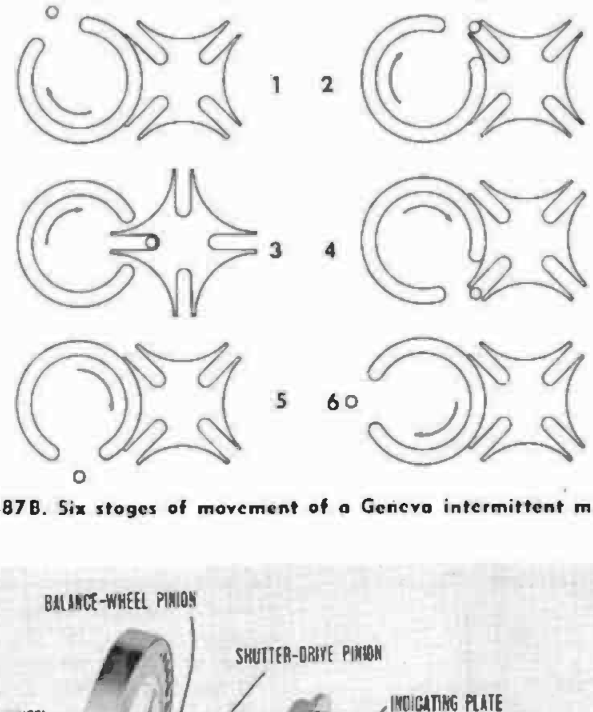


Fig. 19-87B. Six stages of movement of a Geneva intermittent movement.

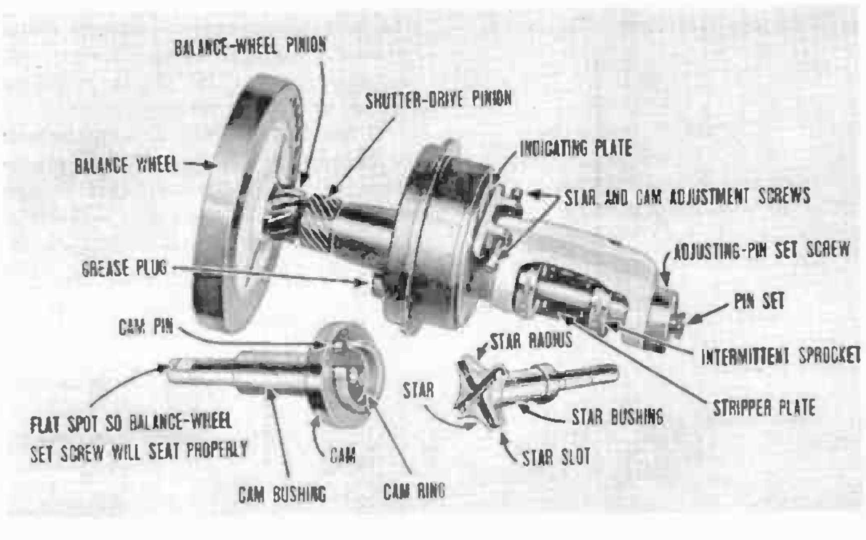


Fig. 19-87C. 35-mm motion picture projector intermittent movement.

and the star begins turning. This movement is slow at first and gradually increases until the star has reached its maximum speed when the pin is completely in the slot and is in line with the centers X and Y. When the pin leaves this position, the star (and consequently the film) begins to slow down. At the end of the movement, the pin leaves the slot and returns to its original position. When the pin leaves the slot, the star has stopped moving.

The purpose of the ring D is to hold the star stationary after the pin leaves the slot. A a result, the film is motionless during the time the cam completes its movement.

A study of the six positions shown in Fig. 19-87B will show that, from the time the pin enters and leaves the slot, the star has made a quarter revolution. For one complete revolution of the cam, one-fourth or 90 degrees of its time is devoted to moving the film. This is called a 4:1 or 90-degree movement. 35-mm film travels at a speed of 90 feet per minute, or 24 frames per second. Therefore, each frame has 1/24 of a second devoted to it. The film is in motion one-fourth of the time or 1/96th of a second for each frame. To assure quiet and smooth operation, the complete movement is placed in a sealed casting where it runs in a continuous bath of oil. An assembled intermittent and the principal parts of the movement are shown in Fig. 19-87C.

19.88 What is an Askania claw movement?—An intermittent movement used in 16-mm projectors (Fig. 19-88).

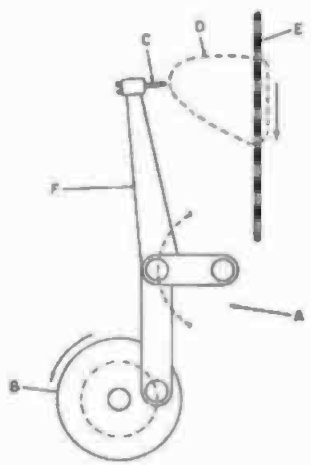


Fig. 19-88. Askania claw movement used in 16- and 8-mm projectors.

This movement has a ratio of 1:1. The hinged lever F modifies the motion of the cam B and imparts to the pulldown claw C the motion indicated by the dotted line D which moves the film E in the direction shown by the arrow. Because of the film wear induced by this movement, it is used only in 16-mm and 8-mm projectors. Intermittent, Geneva-type movements used in 35-mm projectors are too expensive for the average 16-mm projector.

19.89 What is a beam splitter?—A prism used in the optical system of a sound head designed for push-pull operation. The beam splitter divides the light beam in such a manner that the modulations of the push-pull sound track fall on the two plates of the photocell. This is illustrated in Fig. 19-94.

19.90 What is a fader?—A dual gain control connected between the outputs of two projectors, as shown in Fig. 19-90.

In the early installations of projection equipment, the fader was mounted on the front wall of the projection booth. When changing over from one machine to the other, the operator faded the sound from one machine to the other, thus obtaining a smooth transition between machines. Present-day equipment uses automatic changeover devices which permit smooth cutovers without the operator being required to fade the sound. (See Question 19.42.)

19.91 What causes noise in an optical sound head?—Several different things may cause such noise. Among the most common are mechanical vibration of the exciter-lamp filament, microphonic tubes in the preamplifier stages, and the phototube.

If the exciter-lamp filament is loose or improperly adjusted, the slit modulates the scanning-beam image because of vibration from the motor and drive-gear train. The sound generated appears as a low-frequency gear noise.

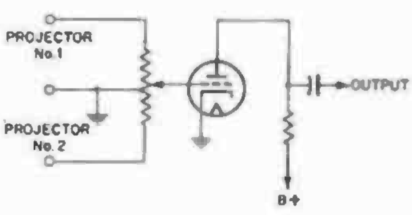


Fig. 19-90. Fader control for cutting over from one projector to another.

Replacement or adjustment of the lamp will generally remove the noise. To check the exciter-lamp noise, run the projector without film and turn up the gain about 20 dB above normal. If the exciter lamp is vibrating, the noise can be made to disappear by interrupting the light to the phototube with a card.

Background noise is also generated by the vibration of the phototube coupling transformer, or the phototube itself, because of loose elements. If the transformer is at fault, the only cure is to float the transformer in sponge rubber, then place it in a nested shield to prevent pickup from surrounding magnetic fields. (See Question 8.51.)

19.92 What type optical sound tracks are most sensitive to sound-head misalignment?—Variable-area tracks. It is quite important that the slit azimuth and slit illumination be properly adjusted; if they are not, serious harmonic distortion may result. If the reproducer is the push-pull type, the photocell circuit must be carefully balanced for maximum cancellation. This subject is discussed more fully in Question 19.64.

Variable-area sound tracks must be aligned more carefully than variable-density ones. However, in a properly maintained system, little trouble is experienced with either type of track.

19.93 Describe the construction of a phototube.—A phototube consists principally of two electrodes in a glass evacuated envelope (Fig. 19-93A and Fig. 19-93C). The cathode emits electrons when its sensitized surface is exposed to a source of light or other radiant energy. Electrons are drawn to the anode because this electrode is at a positive potential. The number of electrons emitted by the cathode depends on the wavelength and the amount of radiant energy falling on the cathode. The phototube thus provides an electric current whose magnitude can be controlled by the light or other radiant energy.

The sensitivity of a phototube is basically defined as the quotient of the current through the tube by the radiant flux received by the cathode. The sensitivity of the tube depends on the color of the light or spectral distribution of the radiant flux to excite the tube. When two such tubes are being considered for use with a given light source, and if the tubes have a different color response, a comparison of the sensitivity may be misleading, unless both ratings are for the same light color. Sensitivity can be measured with the radiation supplied by a tungsten lamp, operated at a filament color temperature of 2870 degrees Kelvin, since the exciter lamp in a projector is operated at or near this temperature.

The presence of inert gas in a phototube will increase its sensitivity, therefore the manufacturer of the tube introduces a controlled amount of gas after evacuation. Gas-type phototubes have higher sensitivity than corresponding vacuum types. Insertion of the gas in the tube causes the following action. Electrons moving from the cathode to anode collide with gas atoms. In such a collision, the electrons may disrupt the atom, knocking an electron out of the atom, leaving a position ion. This disruption of the atom increases the current through the tube because the new electron is drawn to the anode and the positive ion is drawn to the cathode. The positive ion can further increase the current when it arrives at the cathode, by causing secondary emission from the cathode. Therefore, the presence of gas increases current in two

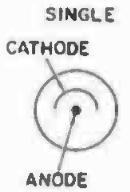


Fig. 19-93A. Phototube construction. Single tube plan view showing the cathode and anode.

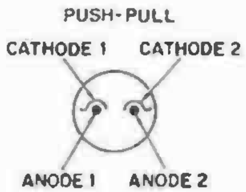


Fig. 19-93B. Dual phototube used for the reproduction of push-pull optical sound track.

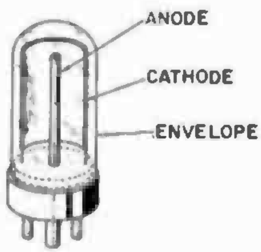


Fig. 19-93C. Interior construction of a single phototube showing the major parts.

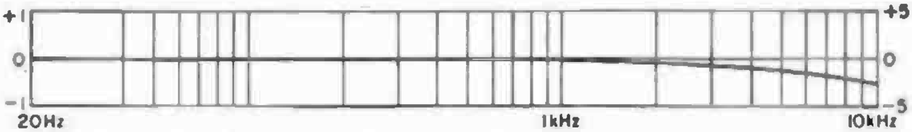


Fig. 19-93D. Typical frequency characteristic for a gas phototube used for sound reproduction. Equalization is used to compensate for the loss at the higher frequencies.

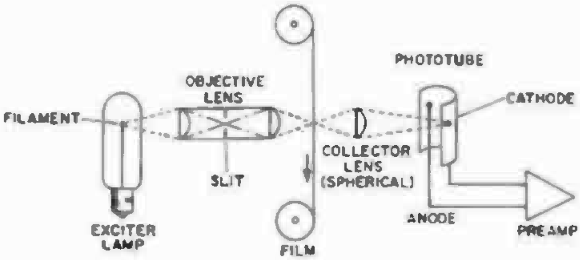


Fig. 19-93E. Optical train for conventional phototube sound reproducer.

ways—by the production of ions and by increasing the amount of cathode emission present.

When phototubes are used for the reproduction of sound, the light to the phototube is varied at an audio-frequeny rate. Under these operating conditions, the luminous sensitivity of the tube is defined as the quotient of the amplitude of variation in phototube current by the amplitude of the variation of the light source. Increasing the load resistance increases the output voltage, but with higher distortion. A typical frequency response for a gas phototube used for sound reproduction is given in Fig. 19-93D.

Push-pull phototubes have characteristics similar to a single tube, except that the two sets of elements are treated as a single phototube. The most important factor in their use is that the two sides of the tube must be balanced electrically, as discussed in Question 19.96. Fig. 19-93B shows how the elements are placed in a push-pull phototube. The term *phototube* is used only with vacuum-type photosensitive devices. Other types are referred to as photocells.

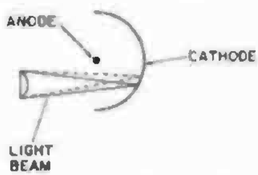


Fig. 19-93F. Method of installing phototube so that the light beam clears the anode element (rod).

When installing phototubes in a sound head, care must be taken that the exciter lamp light beam is so placed with respect to the anode and cathode of the phototube that the light beam clears the anode (rod) and falls on the cathode element, as shown in Figs. 19-93E and F.

19.94 How is the optical train for a push-pull sound head constructed?—In Fig. 19-94 is shown a typical optical train designed to reproduce both single and push-pull sound tracks.

The image of the exciter-lamp filament A is projected on a mechanical slit B in the objective-lens barrel C. The image of the slit is focused as a long narrow beam F on the sound track D by the objective lens E. Leaving the sound track, the light falls on a cylindrical lens G, then passes through a coudenser lens H to a double prismatic lens I. The prism turns the beam upward 90 degrees to a second prism J where it is directed to the anode K and L of a push-pull phototube M. Variations in the amount of light falling on the phototube anodes generate minute changes of current in the cell. These changes are amplified and passed on to the main amplifier equipment.

Single and push-pull sound tracks are reproduced by connecting the cathodes together, thus making the phototube operate as a single cell. For proper operation in the push-pull position, it is necessary that the voltages applied to the anodes have exactly the same value. (See Questions 19.96 to 19.99.)

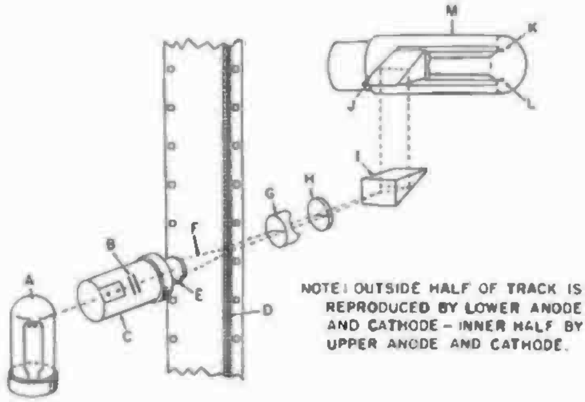


Fig. 19-94. Push-pull beam-splitting optical system, used for the reproduction of push-pull optical sound track.

19.95 What precautions must be taken in the operation of gas-type phototubes?—The actual voltage at the anode of the phototube is generally 90 Vdc. This voltage must not be exceeded. If it is, a gas discharge is likely to occur. This discharge is indicated by a blue glow within the tube. Once started, the glow will continue until the anode voltage is reduced or disconnected. If the glow is permitted to continue, the tube will be severely damaged. The anode voltage can only be measured using a vtvm.

19.96 Describe an optical sound head circuit for both standard and push-pull sound track.—A basic circuit for a single phototube is given in Fig. 19-96A, with the circuity used for push-pull phototubes given in Fig. 19-96B. To achieve a balance between the two halves of the phototube, it is necessary to balance the voltage applied to the anode of each half of the tube. When a balance has been achieved, the signal voltage at the cathodes will be of the same amplitude.

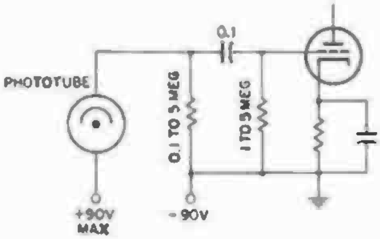


Fig. 19-96A. Basic circuit for a single-section phototube.

The applied voltage is balanced by a voltage-divider circuit and a potentiometer P1 connected in the center of a voltage-dividing network. When the switch is in the push-pull position, the cathodes are connected to a push-pull coupling transformer T1 which has a

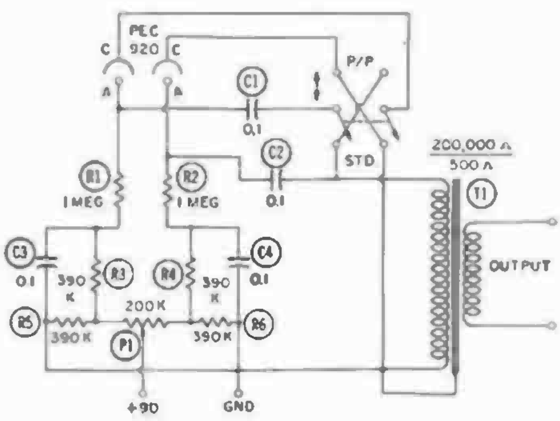


Fig. 19-96B. Schematic diagram for optical sound head for reproducing both standard and push-pull sound track.

high-impedance primary and a low-impedance secondary.

When the switch is in the standard position, the cathodes are connected in parallel and returned to one side of the output-transformer primary. For this connection, the phototube operates as a single-ended phototube.

Capacitors C1 and C2 and resistors R1 and R2 are for frequency correction. The capacitors C3 and C4 in the balancing network provide a low-impedance path to ground for the audio currents and keep them out of the power supply. Resistors R3, R4, R5, and R6 constitute the voltage-balancing circuit.

19.97 How Is the cancellation of a push-pull phototube sound head adjusted?—If the phototube circuits are similar to those of Fig. 19-96, the cancellation is adjusted by the use of the test circuit of Fig. 19-97 and by a loop of negative-bilateral sound track modulated about 80 percent. A voltage amplifier and VU meter are connected across the output of the sound head. (No termiuation is required for the voltage amplifier.) With the sound head phototube switch in the standard position (non push-pull) the gain of the voltage amplifier is increased to a reading of plus 30 dBm on the VU meter. The phototube switch is now thrown to the push-pull position until a reading is obtained on the meter.

The voltage-balancing potentiometer in the sound head is adjusted for a minimum reading on the VU meter. The plus reading is added to that of the minus reading as if they were both positive values. As an example: assume the meter reads a plus 30 dBm in the standard position and the reading in the push-pull position, for the best balance, is minus 10 dBm. Adding of the two readings results in a figure of 40 dB. Thus, the cancellation is 40 dB below the output level of the track when reproduced in the standard position. The greater the cancellation, the lower will be the noise from splices and bloops.

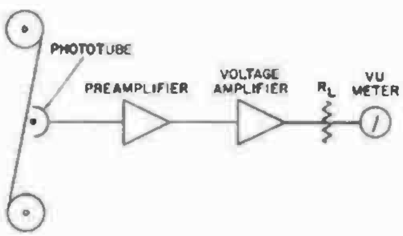


Fig. 19-97. Test circuit for adjusting the cancellation of a push-pull phototube optical film reproducer.

19.98 How is cancellation in a push-pull sound head achieved?—Assume that a 1000-Hz negative-bilateral sound track is playing back on an optical sound head in the push-pull position. The bilateral sound track generates a signal in both anodes of the dual phototube simultaneously. If the two halves of the phototube are in alignment and balance, the signal is the same at each cathode. When the cell is in balance, the signal voltage across the primary of the coupling transformer will be of the same amplitude for both tracks but will be 180 degrees out of phase. Therefore, the signals cancel, and the output signal across the secondary is zero.

When a push-pull sound track is played in the push-pull position, the sound tracks on the film are 180 degrees out of phase with respect to each other; therefore, a signal which is 180 degrees out of pbase and similar to that generated in a push-pull amplifier described in Question 12.226 is generated in each cathode.

With the switch in the standard position (non push-pull), the cathodes of the phototube are connected in parallel and the tube functions as a single phototube. Push-pull sound heads are seldom found in a theater installation and will only be found in older studio equipment.

19.99 What factors affect the cancellation in an optical push-pull sound head?—The mechanical placement of the sound track, azimuth adjustment, optical-system adjustments, electrical balance between the two halves of the phototube, and the general mechanical condition of the sound-head mechanism. (See Question 19.64.)

The cancellation of a sound head should not be adjusted until the machine has been thoroughly warmed up. Mechanical adjustments should be made according to the sound-head manufacturer's instructions.

19.100 How is the sound level from two or more sound heads balanced?—By means of balancing loops which are built from constant-frequency, constant-amplitude stock—either optical or magnetic film. The loops are made just long enough to run in the sound head.

With the machines in motion and a VU meter connected at the output of the main amplifier system, the output levels for either optical or magnetic reproduction are adjusted to match between machines. As a rule, theater equipment includes a test position for this purpose. Constant-frequency film for making such loops can be obtained from the SMPTE.

19.101 How are magnetic head clusters mounted in a penthouse sound head?—They are mounted in two different ways. One method employs two separate head clusters, one for 35-mm heads and a second for 70-mm heads. The second method employs a single head cluster with 35-mm heads on one side and 70-mm heads on the other. The cluster is rotated for the desired-type head. Such a scheme is discussed in Question 19.80. Typical threading paths for a penthouse using separate magnetic heads for 35- and 70-mm reproduction are given in Fig. 19-101.

19.102 Are magnetic sound tracks affected by the steel parts of a motion picture projector?—Not to any great extent; however, as a precaution, it is good practice to demagnetize, at least once each day, all parts that come in contact with the magnetic film. The use of nylon rollers has greatly facilitated the running of magnetic sound tracks. (See Question 17.90.)

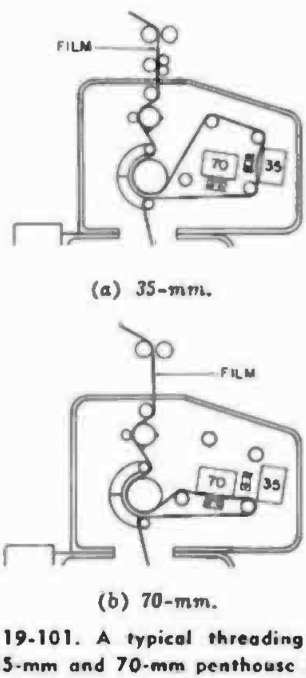


Fig. 19-101. A typical threading path for 35-mm and 70-mm penthouse sound heads, using separate 35- and 70-mm magnetic head clusters.

19.103 What is the width and height of the scanning beam in an optical sound head?—At the present time, the standard for 35-mm is 84 mils in width and 1.2 mils in height. For 16 mm, the width is 72 mils; no height is stated.

19.104 Describe a phototransistor.—Phototransistors are of the junction-type construction, with leads attached to the collector and emitter regions of the semiconductor wafer, but with none attached to the base. Packaged, the wafer is mounted so that the base region may be illuminated by an exciter lamp, through a tiny lens in the nose of the enclosing shell (Fig. 19-104).

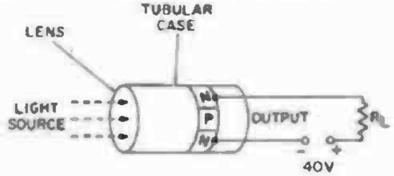


Fig. 19-104. Basic design for a phototransistor.

A dc voltage is applied between the emitter and cathode in the conventional manner. The circuit then becomes that of a common-emitter amplifier. When the wafer is dark, the cutoff current is extremely small. When it is illuminated, current carriers are injected into the base region. The base current is amplified beta times by the transistor structure, and a large collector current proportional to the light intensity is caused to flow.

Silicon planar npn phototransistors are operated at a dc voltage of about 40 volts and provide approximately 9 mA of collector current for 1000 foot-candles illumination. The dark current is about 0.25 microampere. The upper frequency limit is around 25,000 Hz. The advantage of such a device is that it provides amplification for relatively low light intensities with a comparatively large output current. The device may be operated with a steady or modulated source of illumination. Such devices are being used to replace the conventional phototube when the amplifier systems are of transistor design. The circuit for a phototransistor is given in Fig. 19-104. (See Question 11.161.)

19.105 Describe a high-pressure xenon lamp used for projection.—Xenon lamps are high-pressure, gas-filled lamps designed for dc operation. The use of such lamps in motion picture projection is only one of its many uses. The outstanding features of these lamps include a color temperature that resembles natural daylight (which is a mixture of direct sunlight and reflected skylight), immediate readiness for operation, extreme brightness, and extremely good stability. (See Question 19.33.) The light color of xenon lamps is independent of variations in the supply voltage and will remain unchanged even when the light output is being regulated. Xenon lamps emit a strong medium and long wave ultraviolet radiation with a continuous spectrum and have several radiation maxima in the short wavelength infrared range between 8000 and 10,000 angstrons. Such lamps can be used in 16-, 35-, and 70-mm projectors, in small devices such as slide projectors, in light-scanning devices, and for many industrial purposes. They are particularly well suited to wide-screen projection. A typical installation of such a lamp is shown in Fig. 19-105. It will be observed that the mirror from the arc lamp has been retained and only the arc mechanism has been removed. The illustration is the interior of the lamphouse shown in Fig. 19-80A.

Xenon lamps are manufactured by Osram of Berlin, West Germany, Toshiba of Japan, and others. This lamp is rapidly replacing the conventional arc lamp and incandescent light, because of its long life and color spectrum. Although many installations employ the lamp in a vertical mounting position, it may be mounted horizontally, with the length parallel to the sides of the lamp house. One of the advantages of such lamps is that as the lamp output is reduced in intensity, the color temperature remains the same.

19.106 Describe the construction of a xenon projection lamp.—The envelope of a xenon lamp consists of quartz of ellipsoidal shape, with two electrodes placed diametrically opposite. The cathode (negative) must be made small. The anode (positive) must be relatively large in comparison, in order to dissipate the heat. The overall length of the lamp is determined by the temperature gradient between the electrodes and the bases. Because of the prefocus base, the

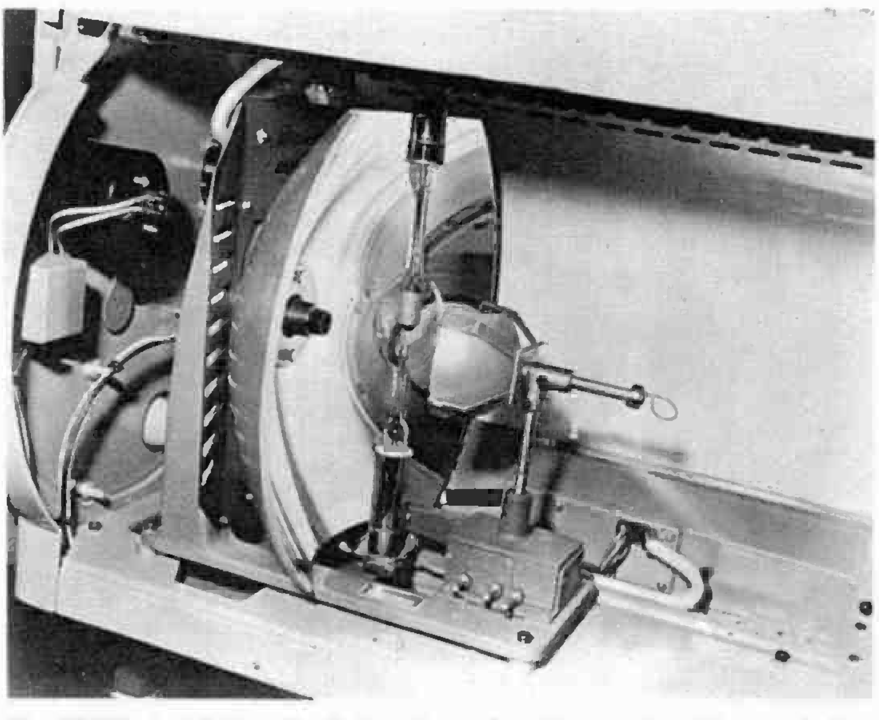


Fig. 19-105. Installation of a XeTron (xenon lamp) in a motion picture projection lamphouse. (Courtesy, Carbons Inc.)

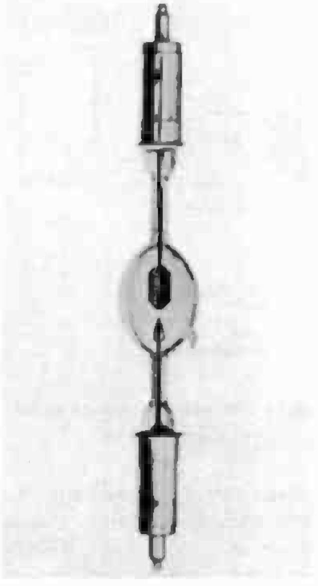


Fig. 19-106. A 1600-watt xenon high-pressure projection lamp. Manufactured by Osram of West Germany. (Courtesy, Macbeth Sales Corp.)

lamp in use may be replaced by a similar lamp without making any adjustment. A special property of the xenon arc lamp is that it absorbs only a minimum of its own radiation. When reflecting an inverse image of the arc in the mirror used in the original arc-lamp installations, the inverse image must be aligned so that the bright parts of the arc (cathode hot spot) are not imaged on one of the two electrodes. Although the concentration of power is high, xenon lamps need no forced-air cooling because of their shape. Because of the glare, the ultraviolet radiation, and the high operating pressure, the lamp must be completely enclosed in the lamphouse with air vents.

The gas filling has an excess pressure of several atmospheres (14 lbs per sq. in.) even when the lamp is not in use. Therefore, protective *gloves and mask* must be worn when handling the lamp. The burning position is always vertical, with the smaller electrode downward, as shown. To ignite the lamp, a special ignition device is used. The lamp may be started quickly by pushing a button for less than one-half second. A 1600-watt xenon lamp is shown in Fig. 19-106.

19.107 What is the life of a xenon tube?—The life of these lamps is generally terminated by the blackening of the discharge vessel. The degree of blackening is influenced by the magnitude of the current pulsations and the number of starts. The average life for a 150-watt lamp is 1200 hours; for 450 to 1600-watt lamps, the life is 200 hours; for 2500-watt lamps, the life is about 100 hours.

19.108 Give the characteristics for xenon projection lamps.—As an illustration, three sizes of lamps will give 450, 900, and 2500-watts. Xenon lamps are generally powered from single or three-phase silicon rectifiers, having fairly good regulation. A switch is provided in the primary wiuding to set the operating voltage to its correct value. In addition, au igniter circuit is also required. Voltage, current, and luminous flux will vary with tubes of different manufacture.

19.109 Give the circuit for a high-pressure xenon lamp.—The circuit for igniting the lamp and the keep-alive low-voltage power supply is shown in Fig. 19-109. The circuit consists of a transformer T1 which steps up the line voltage to 5000 volts and applies it to a 5000 to 30,000-volt step-up transformer T2, which is connected across the electrodes of the xenon lamp. When the lamp is energized, the low voltage is applied first, aud then the push button for the igniter circuit is held down for less than one-half second. This causes a spark across the quenched gap, which, in turn, generates a 30,000 V spark across the electrodes of the lamp. This spark gap breaks down the internal re-

Rated watts	*Lamp supply vlt.*	*Operating vlts.*	*Current Amperes*	*Luminous flux lumens*	*Light output to screen (lumens)*
450	70 Vdc	18 Vdc	28	13,500	2400
900	70 Vdc	22 Vdc	50	32,000	7000
2500	85 Vdc	30 Vdc	95	100,000	18,000

Fig. 19-108. Characteristics of some xenon projection lamps.

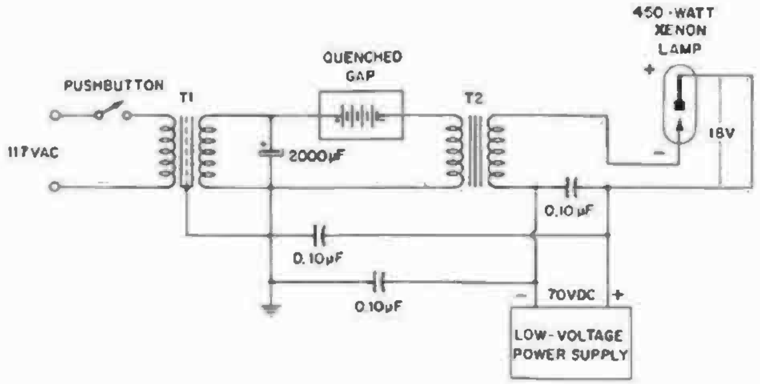


Fig. 19-109. Basic circuit for the igniter and keep-alive low-voltage power supply for a 450-watt xenon projection lamp. (Courtesy, Carbons Inc.)

sistance between the electrodes. As the resistance is now low between the electrodes, the low voltage power supply is turned off. The quenched gap consists of several small gaps in series with cooling fins similar to heat sinks. These fins radiate the heat and cool the gaps almost immediately and cause deionization, and thereby conduction.

19.110 What is the spectral sensitivity characteristic of a gas phototube used for sound reproduction?—The maximum sensitivity centers around 8000 angstroms, or in the red region. The spectral sensitivity of a typical gas-filled phototube is shown in Fig. 19-110. (See Question 19.111.)

19.111 What is the sensitivity of the human eye compared to a phototube for a given color?—A phototube will respond to variations of light intensity in the order of 0.10 percent. The human eye will not recognize a change less than 2 percent The spectral sensitivity characteristic for the human eye is given in Fig. 19-111.

19.112 What is the strongest light the average human eye will accept?—About 16 billion times brighter than the least perceptible light. This is a range of 0.000001 to 16,000 millilamberts.

19.113 What methods are used to couple phototubes to a preamplifier?—Phototubes may be connected by a special coupling transformer or by a low-capacity cable. In the latter type of connection, the output of the phototube is taken directly to the control-grid circuit of the first tube in the preamplifier. Two methods of connection are shown in Fig. 19-113.

19.114 What causes a picture flicker?—If a vacuum-tube rectifier is being used to supply direct current for the arc lamp, the rectifier is generally designed to use six rectifier tubes in a

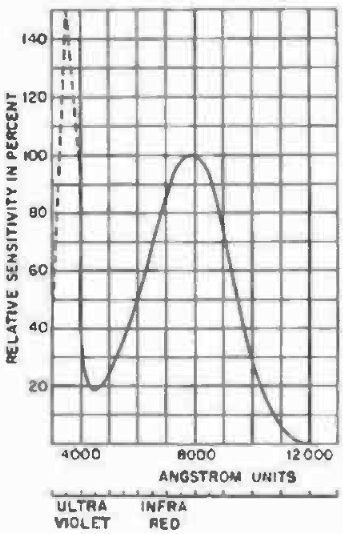


Fig. 19-110. Spectral sensitivity characteristics for a typical gas phototube used for sound reproduction.

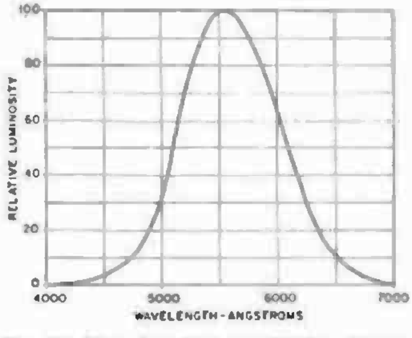
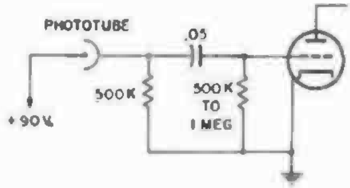
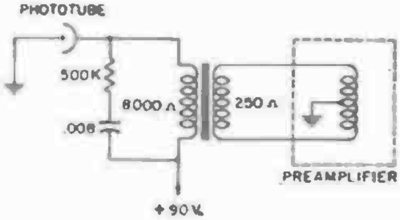


Fig. 19-111. Spectral sensitivity characteristic of the human eye.



(a) *Direct coupling using low-capacity cable.*



(b) *Transformer coupling.*

Fig. 19-113. Methods of coupling a phototube to the input of a preamplifier.

full-wave, 3-phase rectifier circuit. (See Fig. 19-114.) If the rectified current is not the same in each tube, a flicker will be noted on the screen.

Since the projector shutter frequency is 48 Hz, and if the rectifier is operating from 60-Hz power, the flicker rate will be 12 Hz, which is below the rate of persistence of human vision. Thus, a flicker is seen. The current passed by each rectifier tube should balance with respect to that passed by the others within 1 ampere.

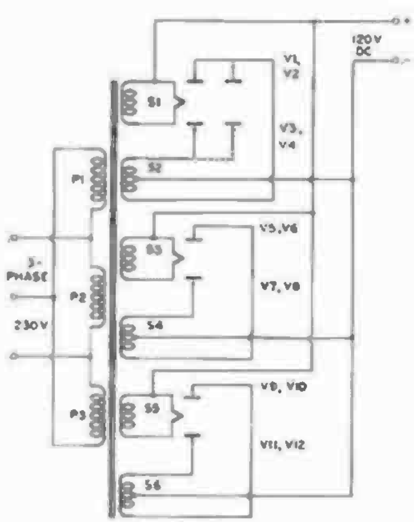


Fig. 19-114. Three-phase, vacuum-tube rectifier for motion picture arc lamp use. Each rectifier actually consists of two tubes connected in parallel.

With rectifiers such as silicon diodes, flicker can often be traced to a voltage difference between the three phases of the power source feeding the rectifier unit.

Regardless of the number of phases, flicker can be caused by voltage unbalance between the phases. The greater the unbalance the greater the flicker. If possible, the voltage between phases should balance to within 2 percent.

Balancing of the voltage between phases is a must with either type rectifier. In cutting over from one machine to another, the ac voltage to the primary and not the dc voltage to the lamp should be broken, otherwise damaging transients can be produced and cause the rectifier tubes to burn out or cause damage to the rectifier units because of heavy surges. As a rule, when the ac to the rectifier is opened, a relay opens the dc line to the arc lamp.

19.115 What is the cause of flicker in a 16-mm projector?—Generally, it is due to the type shutter and the pull-down mechanism. However, in the newer projectors, because of the improvement of the illumination systems, the flicker is reduced by the use of a three-bladed shutter, as it increases the flicker rate to 50 Hz.

19.116 What is the procedure for adjusting an anamorphic lens?—Using a picture or test film as described in Question 19.129. With the regular projection lens in place, the image is focused on the screen, and the image size is adjusted to fill the screen in a vertical direction. The anamorphic lens is then attached and adjusted to expand the picture to the right dimension in the horizontal plane; this is the usual plan of adjustment.

When a long throw is encountered, the light requirements may become impractical. This condition may be partially overcome by what is termed reverse anamorphosis. The normal projection lens is changed to one that will fill the screen in a horizontal plane. The anamorphic lens is then installed and adjusted so that the image is reduced in vertical plane to fit the screen height. In this manner, considerably more light is obtained at the screen. (See Question 19.118.)

19.117 Is the regular lens system of a projector used with an anamorphic lens?—Yes. The anamorphic lens system

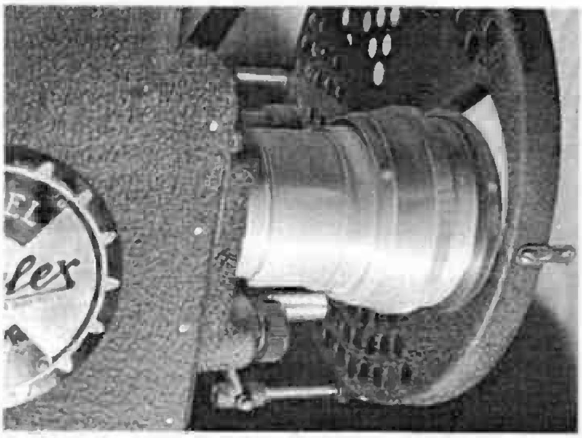


Fig. 19-117. Anamorphic lens mounted on a Simplex projector.

is placed in front of the regular lens, as pictured in Fig. 19-117.

19.118 Describe the construction of an anamorphic lens.—When the picture is originally photographed, an anamorphic lens is mounted ahead of the regular lens used on the camera. The image is squeezed horizontally and looks very much like a television picture in which the horizontal linearity is out of adjustment. This causes the characters to have narrow bodies and elongated heads. Squeezing the image permits a greater angle in the horizontal plane. When the film is projected, the anamorphic lens on the projector spreads the squeezed image on the film back to its original proportions.

The projected picture can use any one of the ratios given in Question 19.83. The most common of these ratios is 2.55:1.

19.119 Where is an optical sound track placed on a CinemaScope magoptical print?—The four sound tracks are placed in the same manner as for normal CinemaScope release print. The monophonic optical sound track is placed in a position in which it will reproduce on any standard projector; however, the sound track is somewhat narrower than normal. If the print is to carry only an optical sound track (no magnetic), the same print is used minus the magnetic striping.

19.120 Describe the difference between a CinemaScope sprocket and standard sprockets.—In the original CinemaScope production of *The Robe* (released in 1953), special perforations were used in the film base, with special sprockets on the projector. The sprockets were standard except for the width of the teeth, which were narrower in order to provide more room for the magnetic sound tracks. These special perforations and sprockets are still used for CinemaScope productions. Standard 35-mm film can be run on these sprockets without difficulty. However, CinemaScope perforations cannot be run with standard sprockets, as to do so will damage the perforations. CinemaScope release prints, both 55- and 70-mm, also use the special perforations and sprockets.

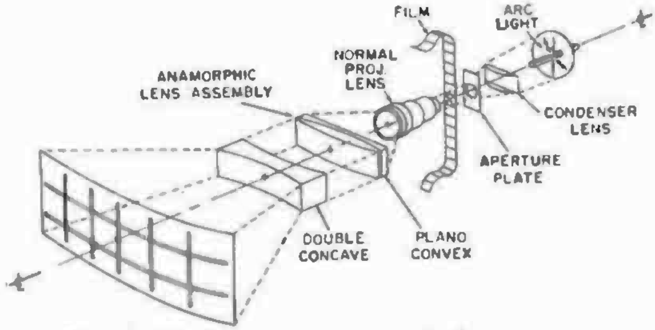


Fig. 19-118. Schematic of anamorphic lens system.

19.121 Why are curved screens used with wide-screen projection?—To bring the picture focus over the entire surface of the screen. (See Question 19.124.)

19.122 How is the curvature of a curved screen plotted?—From the angle of projection as shown in Fig. 19-122. It will be noted that the screen is curved inward and is a segment of the projection arc.

If the vertical projection angle exceeds 12 degrees, the screen may be tilted back at the top to reduce the effects of keystoning. Curving the screen reduces the out-of-focus condition at the edges of the screen by bringing the edges forward, thus shortening the distance of the throw at the sides of the screen. (See Question 19.121.)

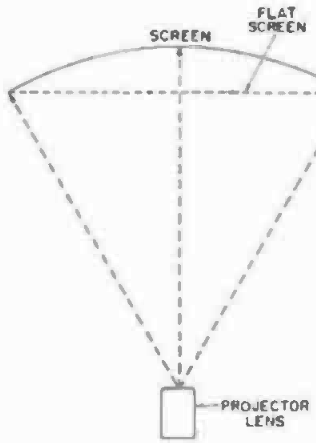


Fig. 19-122. Layout for a curved projection screen.

19.123 What determines the width of a wide-screen picture?—The maximum height of the screen image, the ratio of the projection aperture plate, and the expansion ratio or magnification of the anamorphic lens system in the horizontal plane. Thus, for 55.6-mm film, using aperture dimensions of 1.340/1.050 with an anamorphic lens expansion of 2:1, the aspect ratio is:

$$\frac{1.340}{1.050} = 1.275 \times 2 = 2.55:1$$

Aperture plate dimensions and ratios are given in Question 19.83. (See Question 19.126.)

19.124 What is the cause of out-of-focus conditions at the edges of a wide screen?—Using a flat screen, the projection beam has to travel a greater distance to reach the edges, than to reach the center, as shown by the dotted lines in Fig. 19-124. Thus, it may be seen that the center is out of focus with the edges. Current wide-angle lenses are designed to bring the edges into sharp focus when a curved screen is employed. Not all out-of-focus conditions are caused by the projection arc. Focusing can also be affected by the lens system in use, keystoning, and the design of the screen. (See Question 19.122.)

19.125 Define the term "anamorphic ratio."—Anamorphic lenses are specified by their magnification or spreading ratio, which indicates how much the picture can be spread on the horizontal plane. The height of the picture remains the same. Thus, for a 2:1 ratio, the picture is twice the size. The height remains the same for a given value.

19.126 Give the aspect ratios, past and present.—Aspect ratio is the ratio of the picture height to the picture width. The various ratios used by different studios, and their origination is given below. Many of these ratios are still in use and vary with different productions and film size.

Film Width	*Aspect Ratio*
35 mm	1.34:1 (Academy Standard)
35 mm	2:1 (Superscope 1954)
70 mm	2.21:1 (Todd-AO 1955)
35 mm	2.21:1 (Panavision 1954)
70 mm	2.21:1 (Super-Panavision 1958)
3 × 35 mm	2.27:1 (Ultar-Panavision 1961)
70 mm	2.27:1 (Ultra-Panavision)
35 mm	2.34:1 (Techniscope 1963)
35 mm	2.35:1 (CinemaScope 1953)
55.6 mm	2.55:1 (CinemaScope 1956)
3 × 35 mm	2.77:1 (Cinerama 1952)
70 mm	2.94:1 (MGM-65 1957)

Other ratios used are, 1.66:1, 1.75:1, and 1.85:1. However, these latter ratios are used primarily for cropping. (See Question 19.48.)

19.127 Name the various wide-screen systems.—Several different systems have been developed, using 35-, 55-, and 70-mm film. Many variables are included in these systems: linear speeds greater than 90 fpm, width of film, frame rate, perforations, horizontal magnification, aspect ratio, and the number of sound tracks with two systems using three projectors simultaneously. Also, in certain systems, separate sound heads are used which run interlocked with the projectors. Although the camera is not a problem of the projectionist, there are many variatious in the method of photographing the picture and in the printing processes.

The wide screen systems both past and present are:

CinemaScope, 35- and 55-mm; Panavision; Super-Panavision; Ultra-Panavision; MGM-65; Cinerama; Cinemiracle; Perspecta Sound; Superscope; Techniscope; Vistavision; Todd-AO; Technirama. (See Question 19.128.)

19.128 Describe the special wide-screen systems.—Wonderama: The Wonderama system uses a split image. In this system, the camera photographs the image through an image converter which rotates the left and right halves of the picture 90 degrees in opposite directions, and produces a nonanamorphic image, with an aspect ratio of 3:1. The picture is projected through binocular-type lenses that return it to its correct ratio, but in enlarged dimensions.

Poleyeran: Poleyeran is a system developed in Czechoslovakia. It uses eight screens and is used in combination with live action.

Circarama: At Disneyland and at several world fairs a system called Circarama was used, in which the spectators stood in the center of a circular screen. The screen consisted of 11 separate panels, each 8 feet in height set in a 40-ft circle. The projection system encompassed a group of interlocked 16-mm projectors.

Cinetarium: Cinetarium employs a single projector and presents a 360-degree circular motion picture in a dome-type theater. This system was developed in Germany.

VertiVision: This system of projection was developed by H. M. Tremaine and John H. W. Guselle for the British Columbia (Canada) Expo-70 Pavilion, Osaka, Japan. The system is front projection employing a vertical screen 13′ 6″ wide and 40′ 6″ high. Special optically printed films are used for spectacular effects. A vertical screen 17′ wide and 37′ high was also used at Expo-67 in the Labyrinth Pavilion, Montreal, Canada. This latter system of projection was designed by the National Film Board of Canada.

19.129 Show an amorphic focusing chart.—A special chart is made, similar to that shown in Fig. 19-129. The chart is then photographed, using an anamorphic lens system to compress the image in the horizontal plane. The chart is projected on the screen, and the lens system is focused and rotated until the circles become round and the lines become straight. Although the chart shown is for 20th Century-Fox CinemaScope, such charts may be made for any ratio. (See Questions 19.116 and 19.118.)

19.130 What projection systems are more than four perforations per frame?—Technarama, Super-Panavision, and MGM-65 use five perforations per frame. At the present time, Ultra-Panavision and Cinerama use six perforations per frame.

19.130 What projection systems use more than 24 frames per second?—Cinerama and Ultra-Panavision use 26 frames per second, while Todd-AO uses 30 frames per second. Other systems use 24 frames per second at the present time.

19.132 What projection systems use a linear speed greater than 90 feet per minute?—Technarama and MGM-65 use 112.5 feet per minute; Ultra-Panavision, 122 feet per minute. CinemaScope-65 uses 135 feet per minute. Super-Panavision and Todd-AO use 140.6 feet per minute. Cinerama and Ultra-Panavision use 146.25 feet per minute.

19.133 When a single magnetic sound track is used, where is it placed?

Fig. 19-129. Test chart for adjusting wide-screen projection images, using an anamorphic lens system.

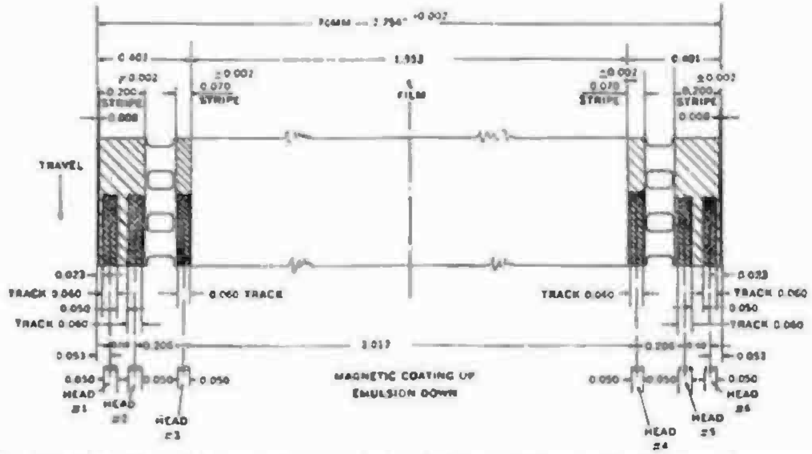


Fig. 19-136. Sound-track placement for 70-mm release prints using 6 magnetic tracks on the film.

—In head position number 1, per USA (ASA) PH22.137-1963, as shown in Fig. 17-189.

19.134 When three magnetic sound tracks are used, how are they designated?—1, 2, and 3. Number 1 feeds the left speaker, Number 2 the center, and Number 3 the right speaker. (See Question 17.169.)

19.135 When four magnetic sound tracks are used, how are they designated?—The same as for three, except the fourth sound track is used for controlling surrounding speakers in the auditorium.

19.136 Give the placement for six magnetic sound tracks on a 70-mm release print.—A recommended track placement is given in Fig. 19-136, but no standard has been set at this particular time. The track placement given is used by Todd-AO and other studios using six sound tracks. The stripes are placed on the film after processing. Two sound tracks, each 60-mils wide and separated by 50 mils, are placed on the outer edges of the base, with two additional 60-mil tracks placed inside the sprocket holes.

19.137 How is the distance between loudspeakers for wide-screen projection determined?—The distance may be calculated thus,

$$\text{distance} = \frac{\text{screen width} \pm 10\%}{3}.$$

The speakers in the system must be of identical types and of the same sensitivity. This prevents the level and quality of dialogue from changing as the characters move across the screen. The dialogue and other sounds recorded in the studio are based on the preceding relationship of the speakers in the theater; therefore, it is important that the speakers be placed in this relationship to the picture frame.

The speakers may also be placed for an aspect ratio of 2.55:1, 0.85 times the picture height. The center line of the high-frequency unit should be located between one-half and two-thirds of the picture height. The horn mouth of the high-frequency unit must be as close to the screen as possible and, since there is a definite relationship between the high-frequency and low-frequency units for correct phasing at the crossover frequency, this usually demands that the low-frequency unit also be placed close to the screen.

The low-frequency unit will generally be pointed slightly toward the center of the audience. The high-frequency units should be twisted around so that they point directly out rather than toed in; and they should, insofar as possible, cover the front rows of seats, favoring the near side if a compromise must be made. Loudspeaker phasing is discussed in Question 20.126.

19.138 What is the minimum acceptable cross-talk level for a stereophonic system?—Not less than 40 dB in a frequency range of 50 to 8000 Hz.

19.139 How important is phasing in a stereophonic system?—Extremely important, as cancellation between various speakers may take place. The reproducing channels must be in phase from the magnetic reproducing heads to the loudspeakers, both electrically and acoustically.

19.140 What is the maximum flutter that may be tolerated in a sound head?—Not more than 0.15 percent rms total flutter.

19.141 What should the signal-to-noise ratio be for a properly adjusted projection system?—Approximately 50 dB. Run a piece of 400 Hz, 100-percent modulated sound track through the projector with the output of the power amplifier terminated in a resistive load. Set the gain control for a normal house level. Read this level with a vacuum-tube voltmeter. Next, run an unmodulated sound track or bias lines through the projector. Increase the sensitivity of the vacuum-tube voltmeter until a reading is obtained. The difference between the two readings is the signal-to-noise ratio of the system.

For stereophonic reproduction, the signal-to-noise ratio should be 50 dB over a frequency range of 50 to 8000 Hz. Under the above conditions, hum in the auditorium will be inaudible.

19.142 What type systems are used for interlocking sound heads and projectors?—Two methods can be used, although the second method is preferred. 1. Selsyn interlock distributor system, using single-phase motors and distributors turning at 1800 rpm. 2. Selsyn interlock distributor, using 230-volt three-phase motors turning at 1200 rpm. In early installations, the machines were interlocked by a shaft on the front wall of the projection room, and chain drives were connected to the machines. However, this is not a satisfactory manner and is not recommended.

19.143 Describe the basic plan for cinerama projection.—The Cinerama system, although not stereoscopic photographically, achieves this stereoscopic illusion because of the very wide curved screen. Referring to the floor plan of Fig. 19-143, three separate projectors are employed, with each projector covering an angle of 48 degrees. The screen is semicircular and covers an arc of 144 degrees horizontally, and 55 degrees vertically. The geometry of the camera lens and that of the projection system is such that the resulting picture is very nearly that experienced by the human eye. The projectors run at a linear speed of 146.25 feet per second using 35-mm film, with 26 frames per second and six perforations per frame. The sound is a double system; that is, two sound heads, each carrying seven magnetic sound tracks, are electrically interlocked with the projectors. Five of the sound tracks feed loudspeakers behind the screen, and two others feed speakers on the left and right of the auditorium. The rear speakers are cued in and out as required.

19.144 What is fantasound?—A special sound recording and reproducing system developed in 1939-40 by J. N.

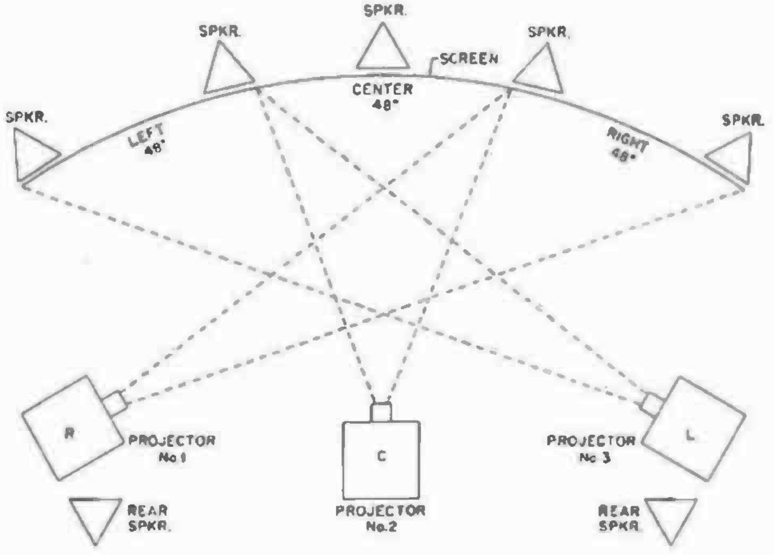


Fig. 19-143. Projection plan for Cinerama projection.

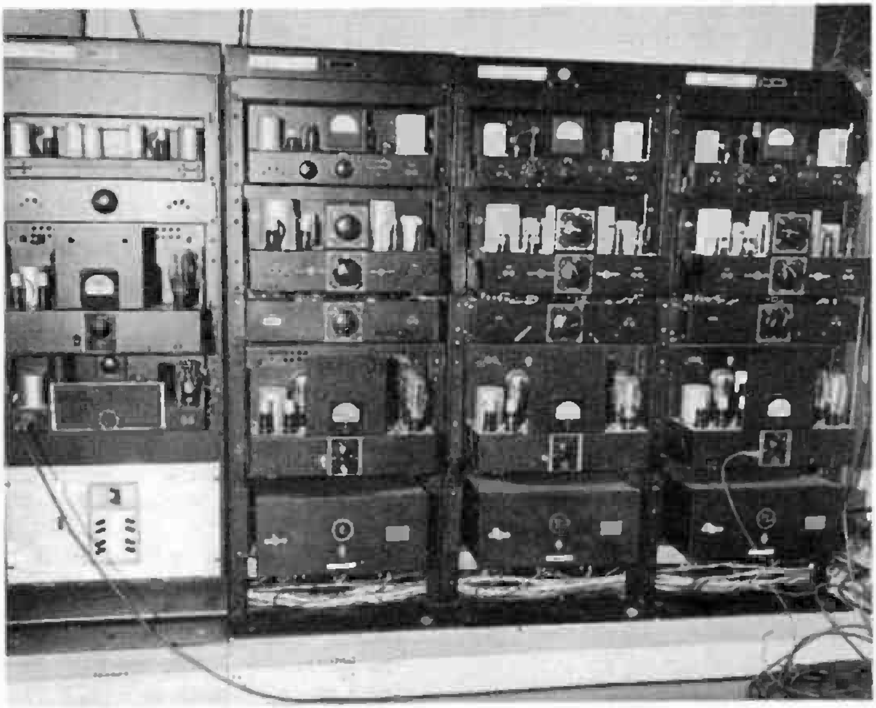


Fig. 19-144A. Section of the reproducing amplifiers used for the Walt Disney Production "Fantasia."

A. Hawkins, William E. Garity, and H. M. Tremaine for Walt Disney Productions and used with the production *Fantasia*. The system consisted of three program tracks and one control sound track on a single optical film. The control track consisted of several different frequencies which controlled the opening and closing of the amplifier systems in relation to the action on the screen. The three loudspeaker systems used were: (1) behind the screen, (2) on the side walls, and (3) at the rear of the auditorium, to create the illusion of the sound moving with the action on the screen. Although the original music was recorded stereophonically, the final recording was made pseudostereophonic. A section of the equipment for reproducing Fantasia is shown in Fig. 19-114A. Present releases of this production use magnetic sound tracks, with less speaker systems than the original plan of the basic system shown in Fig. 19-114B.

19.145 What are the minimum audio power requirements for a theater of a given size?—The audio requirements vary, depending on the cubic footage of the theater. However, the minimum requirements can be found by referring to Fig. 2-47. Although the graphs indicate fairly modest values for smaller theaters, the trend is to use greater power than indicated to obtain more realistic reproduction. If the auditorium of the theater is fairly well damped (dead), greater power than indicated will be required. A minimum of 30 watts is recommended. It is not unusual

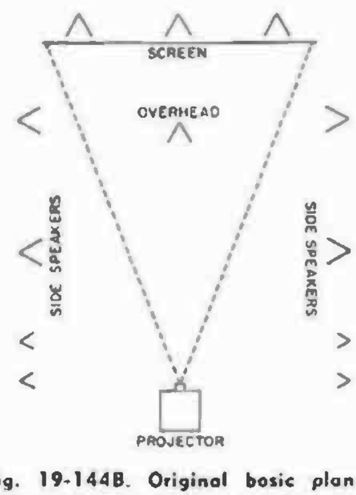


Fig. 19-144B. Original basic plan for Walt Disney production "Fantasia."

in larger theaters to find systems using 250 watts of audio power.

19.146 Describe the various shutter designs, past and present, used for 35- and 70-mm projection.—The sole purpose of a shutter in a projection machine is to cut the illumination off when the intermittent motion pulls the film downward. This occurs at the rate of 24 frames per second. Many different designs of shutters have appeared over the years. One of the original shutters was that of the single-blade type, used on the first projection machines. This was followed by a two-bladed shutter, used at the front or rear of the picture head. Then similar shutters were placed in front and at the rear of the picture head. During this time, the three-bladed shutters made their appearance. All of these shutters were fairly efficient; however, the two-bladed shutter at the rear of the picture head is the most inefficient of all types.

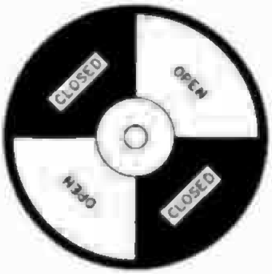


Fig. 19-146A. Two-bladed motion picture projector shutter.

Conventional shutters are required to cut off the light during the intermittent pull-down period of 1/96 second (90 fpm) and to provide a balanced cutoff of equal duration in the middle of the dwell period, resulting in a 48-Hz exposure frequency. Under these conditions, the two blades of the shutter must be of equal angular width. Such shutters have a light transmission of 50 percent.

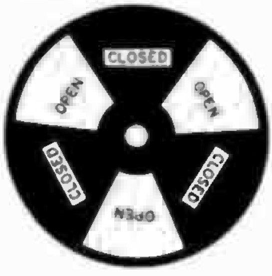


Fig. 19-146B. Three-bladed motion picture projector shutter.

The most efficient shutter at the present time is the barrel-and-cone type, placed at the rear of the picture head. It is the type used with most 70-mm projectors. The closer a shutter can be placed to the picture aperture, the more efficient it becomes.

In the design of projector shutters, several factors must be considered. Among these are the light level at the screen, the speed of the lens system in the lamphouse (concave mirror), and the type of intermittent movement. Plans for two- and three-blade shutters are given in Figs. 19-146A and B. (See Question 19.80.)

19.147 How is the footage of 16-mm film converted to 35-mm film footage?—16-mm film runs at 36 feet per minute, while 35-mm film runs at 90 feet per minute, a ratio of 2.5 to 1. Thus, for a given running time, it requires 2.5 times the amount of 35-mm film compared to 16-mm film. A conversion chart for quick reference is given in Section 25.

19.148 What type glass is recommended for projection room portholes?—For the projectors, a good grade of optical glass about 3/8-inch thick. Two pieces are mounted in removable resilient frames set apart by the thickness of the front wall. The glass in the projection room side is set vertical, while the one on the opposite side is tilted at the top about 5 to 7 degrees, to prevent reflections on the screen. Observation windows should be a double glass to prevent the transfer noise into the auditorium. These too should be tilted to eliminate reflections.

19.149 What type test tracks and film are required for servicing stereophonic sound systems?—The following test tracks and picture are recommended. Certain tests will be required daily, while others will only be necessary when adjusting the system components. Special test tracks and picture may be required for certain systems, in which instance the studio generally supplies them. Test films may be obtained from the SMPTE.

a. Level balance for setting the level of four magnetic sound tracks simultaneously.

b. Multifrequency sound track, for adjusting the frequency characteristics of four sound tracks simultaneously.
c. Loudspeaker-balance test, for testing the response using dialogue and music. The sound progresses from track 1 to track 4.
d. Picture and stereophonic sound tracks with a 12-kHz control track for checking auditorium surround speakers.
e. Flutter test, 3000 Hz, four tracks simultaneously.
f. Loudspeaker-phasing track which is used for adjusting the phasing of individual and complete speaker systems.
g. Picture only, with alignment chart which is used for adjusting the anamorphic lens system.

19.150 When is it necessary to resync the sound track and picture?—When the throw is excessively long. This is accomplished as explained in Question 19.30.

19.151 What is a prefocused projection or exciter lamp?—An exciter or projection lamp with a ring soldered to the base by the manufacturer and in such a position that when the lamp is put into its socket, it will be in focus without further adjustment. The purpose of the prefocus base is to permit a rapid change without adjustment.

19.152 What is a burnishing film?—A special film with a very fine polishing abrasive imbedded in it and used for polishing the aperture plate in a motion picture projection machine. This film may be obtained for either 16-mm or 35-mm projectors and with single or double perforations. (See Question 17.126.)

19.153 What is green film, and how is it projected?—Green film is film just received from the process laboratory that has not been waxed or treated for projection. Such film is often run in the studio projection room if time is a factor, or if the picture is a work print. The running of green film should be avoided, if possible. Green film, when projected for the first time, will generally stick to the aperture plate because of the heat. This will cause a piling up of the emulsion and a consequent scratching of the picture area, plus in and out of focus, and warping of the picture.

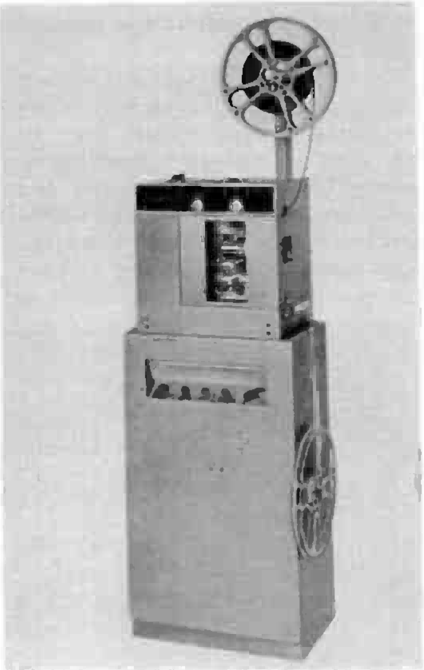


Fig. 19-156A. Eastman Model 30 16-mm projector.

In the absence of a waxing solution, a mixture of beeswax and carbon tetrachloride may be used. This solution is applied with a velvet or lintless cloth while running the film on rewinds. The fumes from the solution must not be inhaled. Normally, motion picture film is lubricated over its entire surface to prevent it from sticking to the shoes in the projector.

19.154 What is the difference between a squeezed and an unsqueezed print?—A squeezed print is one made using an anamorphic lens in the motion picture camera and is projected utilizing an anamorphic lens with reverse characteristics. An unsqueezed print is one shot using a regular lens.

19.155 Describe the construction of polybelts.—Polybelts are wide-base belts containing several V belts. Such belts have greater driving power and steadiness than single V belts. Belts of this type are used on projectors, between the motor and main drive shaft.

19.156 Describe a heavy-duty 16-mm projector.—Fig. 19-156A shows an Eastman Model 30 16-mm projector. This machine is similar in many respects to a regular 35-mm projector and has been designed for use in motion picture studios and theaters where the picture must be comparable to standard 35-mm projection.

The mechanism consists of an intermittent sprocket pull-down with an accelerated Geneva movement with its own directly connected synchronous motor. The remaining sprockets and shutter are driven by a second synchronous motor. The two systems are engaged temporarily for starting and are then run mechanically independent to eliminate shock forces and to obtain an inherently flutter-free sound-transport system.

The schematic diagram for the phototube and magnetic head preamplifier is given in Fig. 19-156B, with the main amplifier shown in Fig. 19-156C, and the various equalizer characteristics in Fig. 19-156D.

Also included is a dc power supply for the sound exciter lamp and a high-

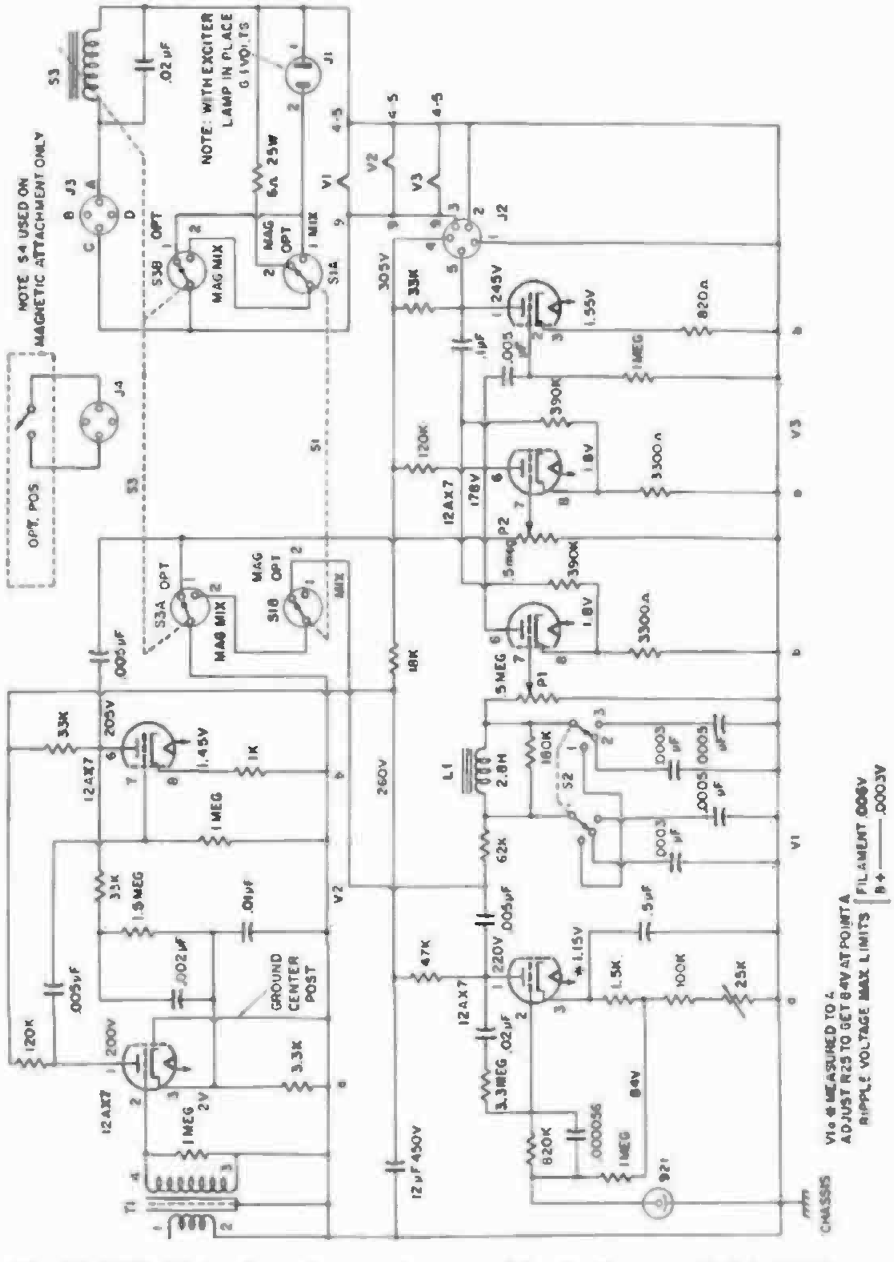


Fig. 19-156B. Phototube and magnetic preamplifier for Eastman Model 30 16-mm motion picture projector.

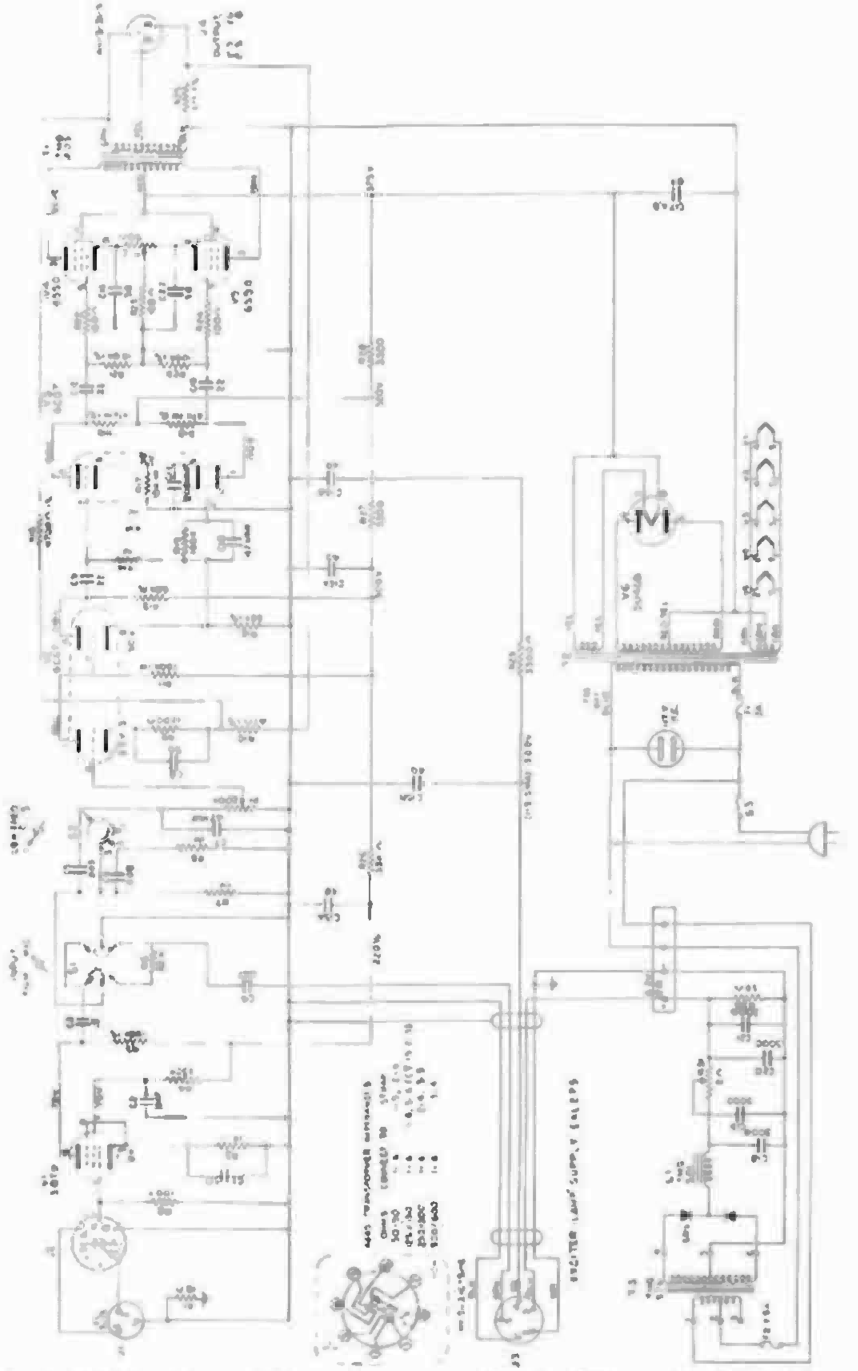


Fig. 19-156C. Main amplifier and power supplies for Eastman Model 30 16-mm motion picture projector.

voltage power supply for the amplifier section. This machine may also be obtained for arc-lamp operation.

19.157 How can the projector in Question 19.156 be converted to selsyn interlock? — This particular projector (and previous models) can be modified to operate both single-phase and three-phase selsyn interlock, by the addition of a 1200-rpm three-phase selsyn interlock motor, as shown in Fig. 19-157A. The selsyn motor is connected to the

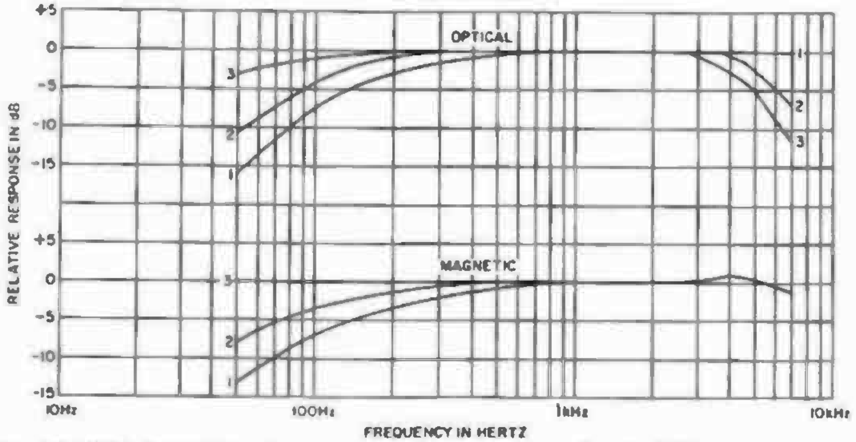


Fig. 19-156D. Frequency response curves for magnetic and optical film reproduction, Eastman Model 30 16-mm motion picture projector.

vertical drive shaft of the projector, through a Gilmer belt and two gears.

In the single-phase mode of operation, the selsyn motor is disconnected electrically and idles with the vertical shaft. In the selsyn mode, the selsyn motor is the prime mover, and the acceleration and deceleration of the single-phase motor is controlled by means of a three-pole relay, connected in conjunction with the single-phase and the interlock system. Such a modification will permit the projector to be interlocked with rerecording equipment or a separate sound head.

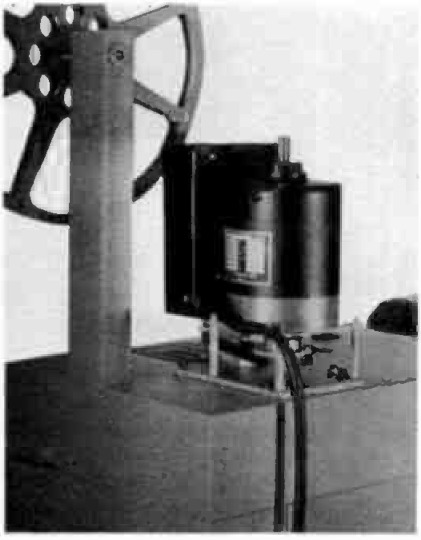


Fig. 19-157A. Three-phase selsyn interlock motor mounted on Eastman Model 30 16-mm projector. (Courtesy, Magna-Tech Electronic Co. Inc.)

The power-circuit modification is shown in Fig. 19-157B. The connections to the selsyn interlock motor and projection machine proper are made through connector plugs P1, J1, and J2. A three-pole relay K1, installed in the projector base, connects to the four-pole switch on the control panel of the projector. After the direction of the motor rotation has been established, the floating gears in the projector head may have to be readjusted because of a slight difference in the phasing. The manufacturer's instruction book should be referred to for this adjustment. The four-pole switch on the projector for controlling the blower, the lamp, and the single-phase drive remains intact. Kits are available for this modification.

19.158 Describe a 16-mm projector-recorder.—A unique design in dual purpose 16-mm projector-recorders pictured in Fig. 19-158 is manufactured by Siemens and Halske of Berlin, West Germany. This machine may be used as a sound projector for both optical and magnetic reproduction, using either 8- or 16-mm film. It also permits recording on 8-mm striped film on regular 16-mm magnetic film. Both the picture and recording are mechanically interlocked for absolute synchronization.

The machine also provides facilities for a narrator to record sound track while watching a silent picture, or the machine may be used to preview style for editorial work. (See Question 19.81.) Nine different modes of operation are possible.

Fig. 19-158 shows that, for magnetic reproduction or recording, the film

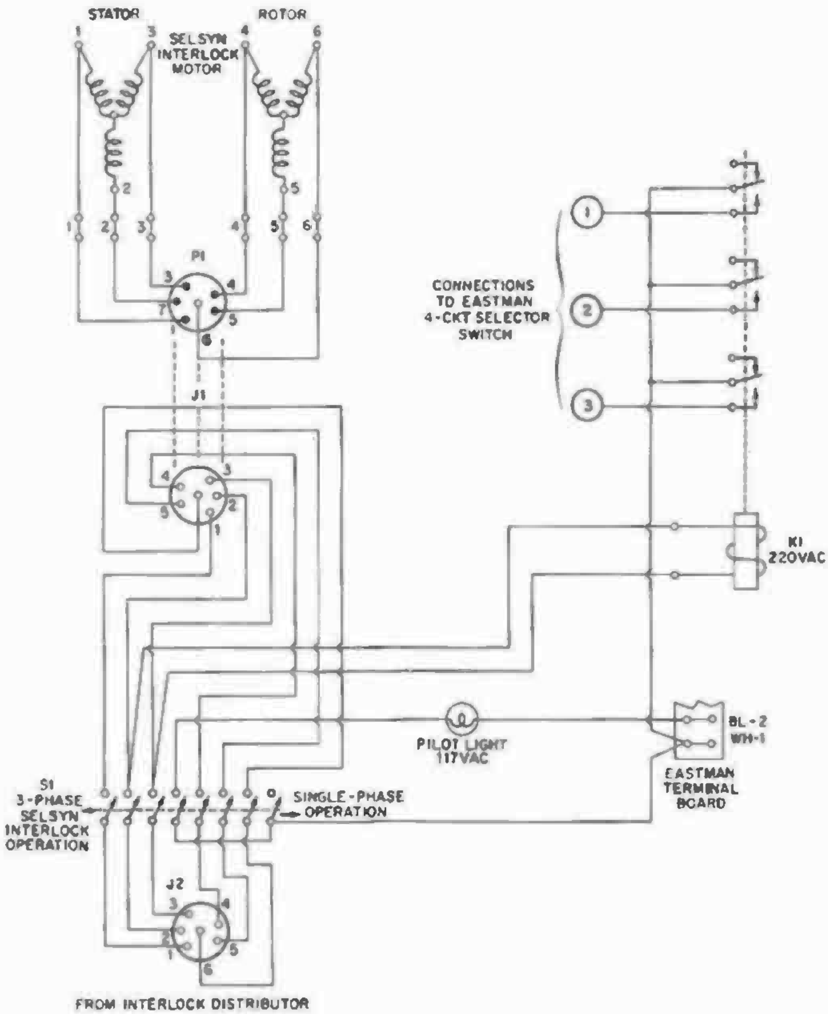


Fig. 19-157B. Schematic diagram for converting Eastman Model 30 16-mm projector for selsyn interlock operation.

leaves the supply reel A and is threaded under sprocket B, then downward to a sound drum C and to its pad roller D. From this point, the film travels upward again over sprocket B, over rollers E and F, to take-up reel G. The magnetic record-playback and bias heads are mounted in a Mumetal shield at H. At I is an amplifier stage used for recording, and at J is the main amplifier. A microphone is shown at K. Item L is an optical-magnetic head assembly for the opposite side of the machine. Provisions are made for patching in other sources of program material by means of jacks at the lower left of the base.

19.159 What is the relationship of lens size, picture size, and throw for 8-mm and 16-mm projection?—This relationship is given in Fig. 19-159. (See Question 19.163.)

19.160 Give the relationship of lens size to throw, and picture size for 35-mm projection, using a standard picture aperture (Academy).—A tabulation of the lens sizes and picture dimensions is given in Fig. 19-160. These figures are for a standard aperture (Academy) 0.825/0.600 inch. The sizes are given to the nearest tenth of an inch. For a projection distance greater than 200 feet, the distance of one-half the required length is selected, and the image size is doubled.

19.161 Describe the Todd-AO system of projection.—The Todd-AO sys-

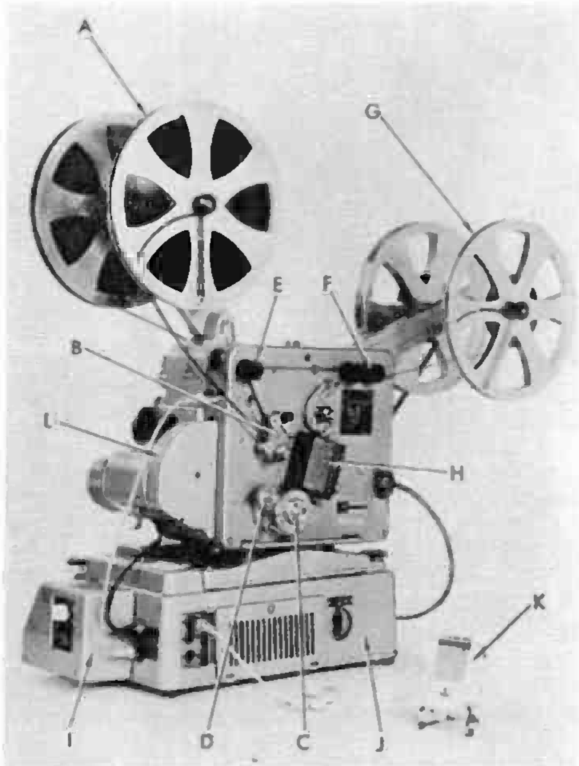


Fig. 19-158. Dual 16-mm projector-recorder manufactured by Siemens and Halske of West Germany. (Courtesy, R and H products Ltd., Canada).

8MM MOTION PICTURES

APERTURE WIDTH — .172"

LENS FOCAL LENGTH	SCREEN WIDTH									
		40"	50"	60"	70"	84"	8'	9'	10'	12'
9/16"	Projection Distance in Feet	10.9	13.6	16.4	19.1	22.9	26.2	29.6	32.1	39.2
5/8"		14.5	18.2	21.8	25.5	30.5	34.9	39.2	43.6	52.3
22 mm.		16.8	21.0	25.2	29.4	35.3	40.3	45.4	50.4	60.5
1"		19.3	24.2	29.1	33.9	40.7	46.5	52.3	58.3	69.8
1¼"		24.2	30.3	36.3	42.4	50.9	58.2	65.4	72.7	87.3

16MM MOTION PICTURES

APERTURE WIDTH — .380"

LENS FOCAL LENGTH	SCREEN WIDTH																		
		40"	50"	60"	70"	84"	8'	9'	10'	12'	14'	16'	18'	20'	22'	24'	26'	28'	30'
5/8"	Projection Distance in Feet	5.5	6.9	8.2	9.6	11.5	13.2	14.8	16.4	19.2	23.0	26.3	29.6	32.9	36.2	39.5	42.8	46.1	49.3
1"		8.8	11.0	13.2	15.4	18.4	21.1	23.7	26.3	31.6	36.8	42.1	47.4	52.6	57.9	63.2	68.4	73.7	78.9
1½"		13.2	16.4	19.7	23.0	27.6	31.6	35.6	39.5	47.4	55.3	63.2	[illegible]	79.9	86.9	94.7	102.6	110.5	118.4
2"		17.5	21.9	26.3	30.7	[illegible]	42.1	47.4	52.6	63.2	73.7	84.2	94.7	105.3	115.8	126.3	136.8	147.4	157.9
2½"		21.9	27.4	32.9	38.4	46.1	52.6	59.2	65.8	78.9	92.1	105.3	118.4	131.6	144.7	157.9	171.1	184.2	197.4
2¾"		24.6	30.2	36.2	42.2	[illegible]	[illegible]	65.1	72.4	86.9	101.3	115.8	125.0	144.7	159.2	173.7	188.2	202.6	217.1
3"		26.3	32.9	39.5	46.1	[illegible]	[illegible]	[illegible]	78.9	94.7	110.5	126.3	[illegible]	157.9	173.7	189.5	205.3	221.1	236.8
3½"		30.7	38.4	46.1	[illegible]	[illegible]	[illegible]	[illegible]	92.1	110.5	128.9	147.4	165.8	184.2	207.6	221.1	239.5	257.9	276.3
4"		35.1	43.9	52.6	61.4	[illegible]	84.2	94.7	105.3	126.3	147.4	168.4	189.5	210.5	231.6	252.6	273.7	294.7	315.8

Fig. 19-159. Picture size, lens focal length, and throw for 8-mm and 16-mm film projection.

	PROJECTION DISTANCE—FEET																
Focal Length Inches	40	50	60	70	80	90	100	110	120	130	140	150	160	170	180	190	200
2.00	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]									
2.25	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]							
2.50	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]						
2.75	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]				
3.00	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]			
3.25	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]		
3.50	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
3.75		[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
4.00		[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
4.25		[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
4.50			[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
4.75			[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
5.00				[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
5.25				[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
5.50				[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
5.75					[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
6.00					[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
6.25					[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
6.50						[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
6.75						[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
7.00						[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
7.50							[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
8.00							[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
8.50								[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]
9.00									[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]

Fig. 19-160. Projection throw, lens, and picture size for 35-mm film using a standard Academy aperture. (Courtesy, International Projectionist Magazine)

tem is a wide-screen process developed by Dr. Brian O'Brien, of the American Optical Co., in conjunction with Magna Pictures Corp., in 1953. This system, like Cinerama, creates the illusion of audience participation with only the use of a single camera for the photography, and one projector in the theater. The release print is 70 mm wide, carrying six magnetic sound tracks (Fig. 19-

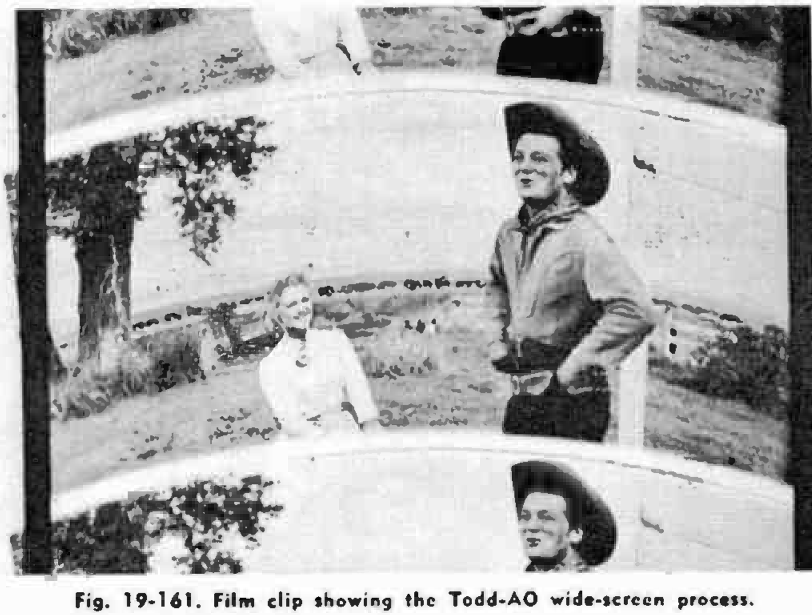


Fig. 19-161. Film clip showing the Todd-AO wide-screen process.

136), with 65-mm perforations, or five perforations to the picture frame.

The curved image shown in the film clip of Fig. 19-161 is necessary to obtain a linear image on the screen. This is accomplished in the projector optical system. The linear speed of the projector is 140.6 feet per minute, 30 frames per second.

19.162 What is a perspecta sound projection system?—A system developed by Paramount Studios for the projection of pseudostereophonic sound, using only one reproducing channel. The system employs three low-frequency tones of 30, 35, and 40 Hz which are superimposed on the sound track and control the speaker pattern. To prevent the control frequencies from being heard, filters are used in the amplifier circuit to remove them. The speakers are selected by relays. In some respects, the system operates in a manner similar to the Dorsett system described in Question 19.170.

19.163 Describe a folded-throw projection system.—A folded-throw projection system is one in which the normal picture is folded by the use of one or several mirrors, to reduce the physical length of the projection beam. Such a system using a 16-mm projector and a front-surfaced mirror is shown in Fig. 19-163A. For this particular system, the picture is projected on the mirror, then reflected to a rear projection screen. The total length of the throw is normally 18 feet, but when folded, the required distance is approximately 9 feet from the screen.

The basic plan for folding the beam is given in Fig. 19-163B. Critical adjustments in such systems are the placement of the projector, the angle of the mirror, and the position and distance of the screen from the mirror. Four such projectors are pictured in Fig. 19-164A, using a common projection area. These projectors were designed by Forest E. Jaquart, of Lytle Engineering Co., and were used in conjunction with a multilingual sound reproducing system, described in Question 19.164. Although the image is reversed at the screen, when viewed from the front by an audience, it is in correct relationship.

It should be remembered that, when using mirrors, they must always be used in odd numbers. If a straight-line beam is used (no mirrors) a prism is employed at the projector to reverse the projected image; otherwise, the image on the film must be printed in reverse. If the print carries a sound track, this

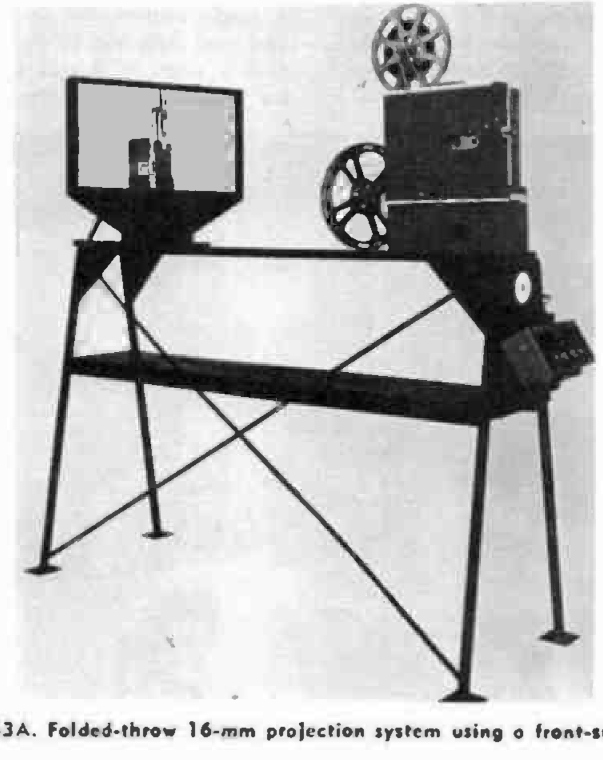


Fig. 19-163A. Folded-throw 16-mm projection system using a front-surfaced mirror.

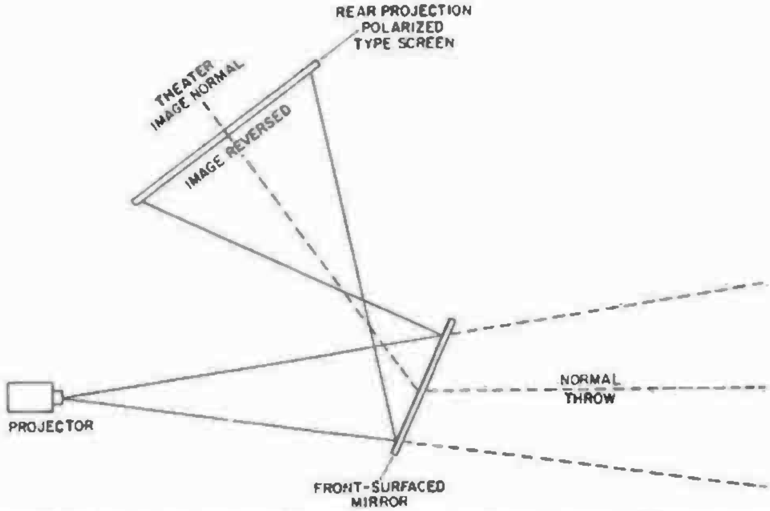


Fig. 19-163B. Basic plan for folded-throw rear projection system. For front-surface projection a second mirror must be placed in the projection beam.

could present difficulties. The prism at the projector does not shorten the throw distance, but only reverses the image. Therefore, when using a reversing prism, only two mirrors are required.

Short focal-length lenses (under 1 inch) should be avoided, if possible. They tend to distort the image, unless they are specially ground. Image width varies in direct ratio to the throw distance. Image height is related to width by the ratio of the image on the film and the projector aperture ratio. (See Questions 19.105 to 19.109.)

19.164 Describe a multilingual sound-projection system. — Pictured in Fig. 19-164 is a multilingual sound and projection system, employing four 16-mm rear projection systems, with each projector using a folded-throw projection beam. Each projector is separately interlocked with a four-track 16-mm magnetic film reproducer carrying four different languages. The image from each projector is focused on the rear of a polarized type screen, manufactured by Polacoat, Inc. Fig. 19-164B shows a floor plan of the projection area

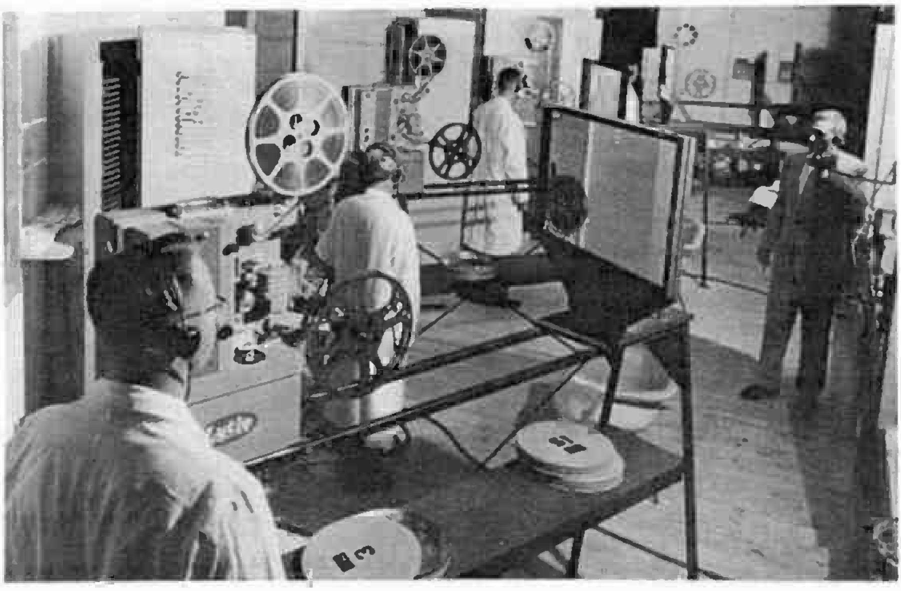


Fig. 19-164A. Projection-room area housing four folded-throw, 16-mm projectors.

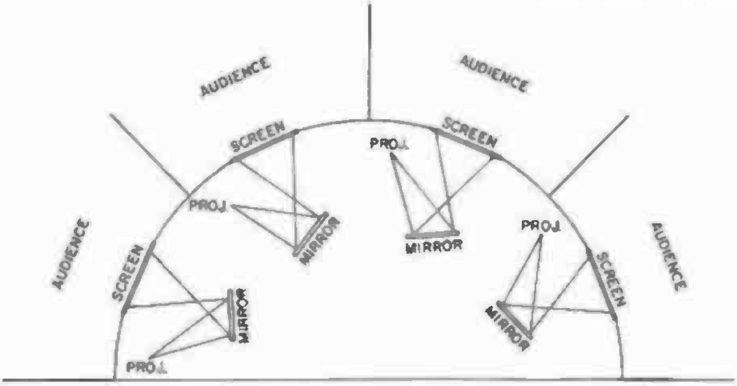


Fig. 19-164B. Floor plan of rear projection installation at the Second International Conference for the Peaceful Uses of Atomic Energy, Geneva, Switzerland 1958.

and theaters. At each seat in the auditorium is a headphone, a volume control, and a language-selector switch. This switch permits the selection of English, French, Russian, or Spanish. The system was designed by H. M. Tremaine for the Lytle Engineering Corp. The four-track recorders and reproducers were manufactured by the Magnasync-Moviola Corp. In addition to the magnetic sound tracks, the picture carried an English optical sound track; however, this track was not normally used. Sound-track placement dimensions for the four-track magnetic sound tracks are given in Fig. 19-164C.

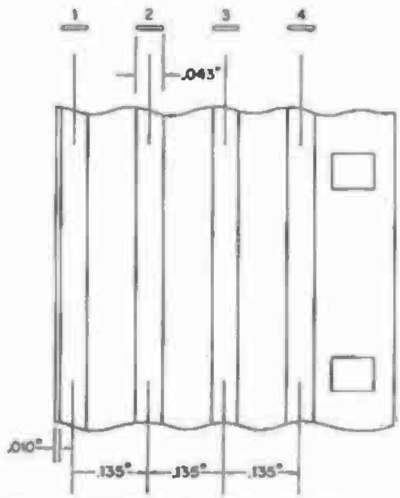


Fig. 19-164C. Sound-track placement dimensions using four tracks on a 16-mm magnetic film.

A front view of one of the four-track reproducers is given in Fig. 19-164D, and a rear view is shown in Fig. 19-164E. The block diagram for a single theater installation appears in Fig. 19-164F.

Starting with head 1 (the English sound track), the signal is amplified by preamplifier 1, and then is applied to the input of a 10-watt power amplifier. The 16-ohm output of the power amplifier is terminated in 16 ohms and feeds one leg of the several language-selector boxes (each theater has 25 boxes). The 600-ohm output circuit feeds a VU me-

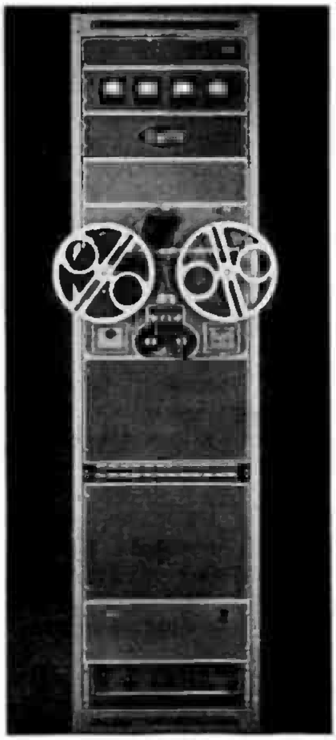


Fig. 19-164D. Front view of four-track 16-mm magnetic film reproducer. (Courtesy, Lytle Engineering Co.)

ter. The French, Russian, and Spanish sound tracks are treated in a similar manner. The block in the diagram labeled projection monitor is a motor-driven cam system that connects the projectionist's headphones across each sound track every 15 seconds to provide continuous monitoring of the four tracks.

Since the headphones and the volume controls are of high impedance fed from a 16-ohm circuit, they function as a bridging load across the amplifier output. No change in level is reflected when the headphones are changed by the audience from one language circuit to another. This not only is prevented by the use of high impedance headphones across a 16-ohm circuit, in addition a terminating resistor is substituted for the headphone load in the selector switch box. One spare preamplifier and a power amplifier are included in each rack for emergency use.

19.165 Why are high-frequency oscillator exciter-lamp power supplies used?—To provide a lightweight, economical source of exciter-lamp current.

Fig. 19-164E. Rear view of four-track 16-mm magnetic film reproducer. (Courtesy, Lytle Engineering Co.)

If 60-Hz is applied to an exciter lamp, hum modulation can occur at a frequency of 120 Hz. A rectifier supplying a source of dc can also be used, but the added weight of components eliminates it for portable equipment. If the projector is a studio type, a dc supply is recommended. The use of a high-frequency oscillator type supply raises the frequency applied to the exciter lamp beyond audibility.

19.166 Give a schematic diagram for a high-frequency exciter-lamp supply. Fig. 19-166 shows a typical high-frequency exciter-lamp supply used in 16-mm projectors. The circuit consists of a single 6V6 oscillator tube operating at a frequency of 125 kHz. Capacitors C1 and C2 are radio-frequency bypass capacitors across the heater supply terminals to prevent rf from entering the power supply circuits.

The output voltage for the exciter lamp is taken from a low-impedance winding L2 coupled to tank-coil L1. The voltage at the coupling coil is approximately 6 volts at 1 ampere.

19.167 Show the methods used for connecting water cooling systems for projectors.—Five different methods for connecting the water supply to a projector are given in Fig. 19-167. If the water supply has a high mineral content, a water recirculation system should be used, and a quantity of rust inhibitor should be added. If the water is of low mineral content, such as rain water, the recirculating system may be dispensed with. The water pressure generally ranges between 2 and 3 pounds per square inch.

19.168 Describe a radio-frequency sound system for drive-in theaters.—Such systems consist of a low power radio-frequency transmitter, operating in the broadcast band. The transmitter is modulated by the theater sound system. The antenna system is designed so that the signal is confined to the theater area and cannot be heard outside the parking area. The program material can be heard by the customer on his car radio if he so desires, rather than his using the loudspeaker normally supplied. For stereophonic production, the car radio and the theater speaker are used together. The frequency of the transmitter is crystal controlled and will not be interfered with by local broadcast stations. The power and installations of such systems come under the jurisdiction of the Federal Communications Commission (FCC). Therefore, a qualified engineering company must issue the certificate of certification stating that the installation meets the requirements of the FCC.

19.169 Describe a solid-state sound-reproducing system.—Pictured in Fig. 19-169 is the interior of a Century Pro-

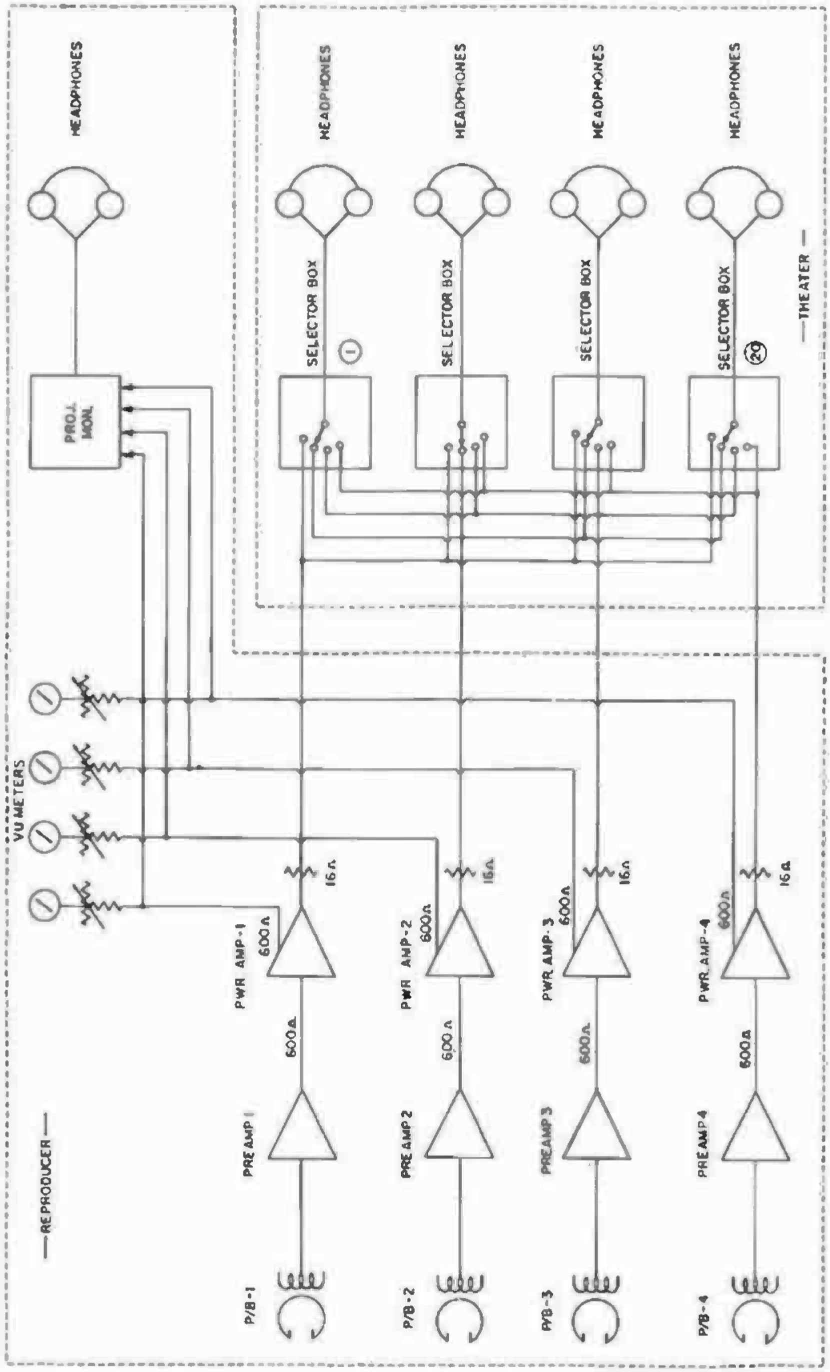


Fig. 19-164F. Block diagram for multilingual sound system.

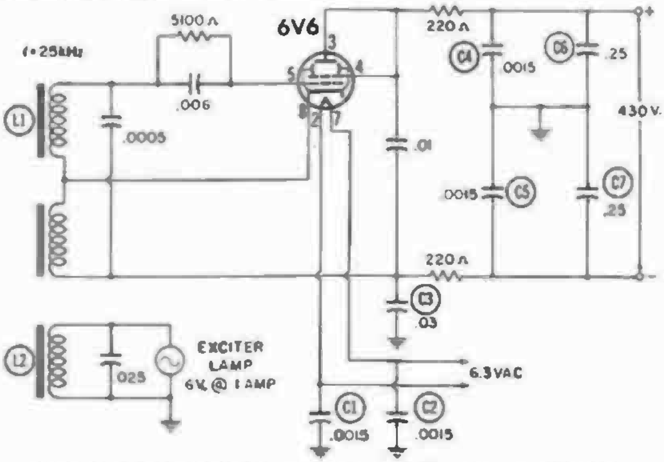
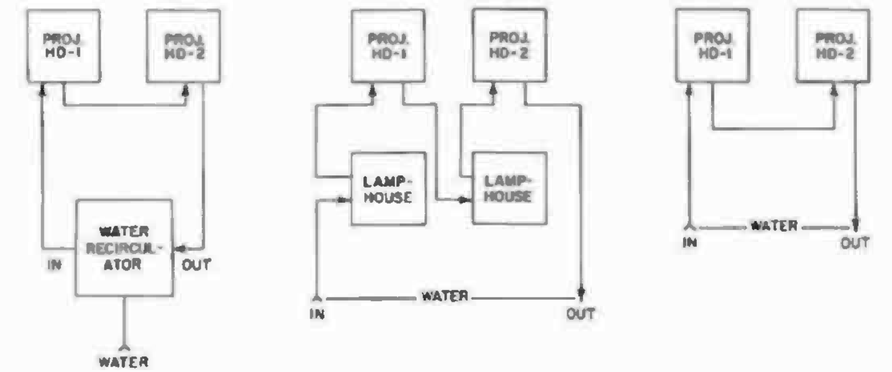


Fig. 19-166. Radio-frequency exciter lamp power supply.

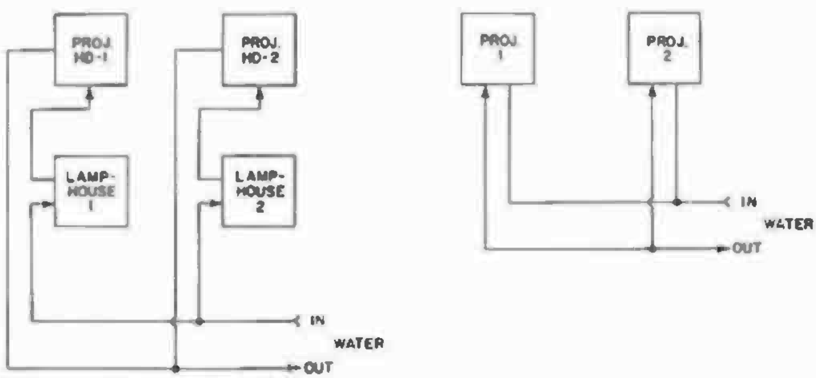
jector Corp., photographic sound-track reproducing sound head. Mounted within the impedance drum is a solid-state photosensitive device termed a ANAPFET (anamorphic photo field-effect transistor) developed for Century by L. W. Davee and E. Chisholm. This device replaces the conventional photo-



(a) *Two projectors with a water recirculation system.*

(b) *Two projector heads and lamphouses, with water supply system connected in series.*

(c) *Two projector heads only in series.*



(d) *Water system in parallel, for two projectors and lamphouses.*

(e) *Two projector heads only connected in parallel with water system.*

Fig. 19-167. Water systems for water cooled lamphouses and picture heads.

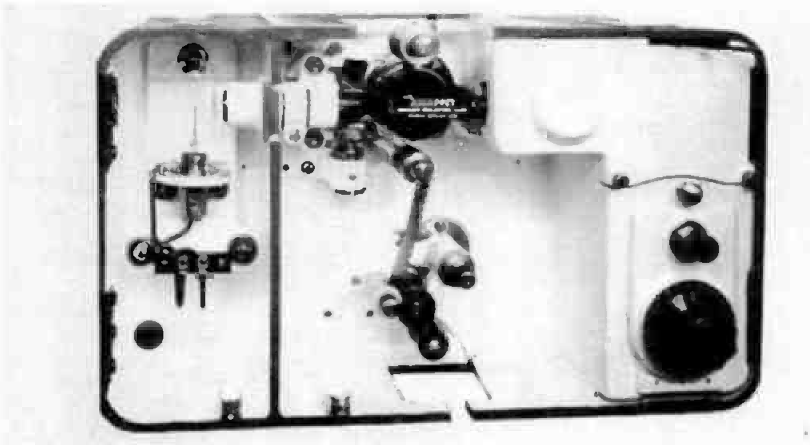


Fig. 19-169A. Interior view of Century Projector Co. sound head, showing an ANAPFET sound-reproducing element mounted inside the impedance drum.

tube assembly generally employed for the reproduction of photographic sound-track reproduction.

In the conventional optical sound-track reproducer, the light falling on the phototube light-sensitive surface is oblong and has an average dimension of .0008 × .10 inch. As this beam cannot be applied to the surface of a photofet (PFET) because of its dimensions, an anamorphic lens system is placed in the optical train for enlarging the photo-sensitive surface of the PFET in one dimension, and to permit the acceptance of the full height of the beam in the other.

Because the sound-track image is diffused, the modulation of the PFET is entirely dependant on the total-light modulation by the sound-track image. Therefore, both variable-density and variable-area sound track is reproduced in the same manner.

The ANAPFET system consists of three elements: optical, PFET, and a transistor amplifier. These three elements comprise a single unit mounted in the sound-head impedance drum as shown in Fig. 19-169A. A schematic diagram of the optical train is given in Fig. 19-169B and Fig. 19-169C, with the schematic diagram for the electronics shown in Fig. 19-169D.

The PFET used in this assembly is a special unipolar device. When radiant energy falls on its surface, a current is generated across the gate channel junction. This current flows into the gate, and out from the drain and source elements causing a variation in the gate voltage which, in turn, is amplified by the FET.

Fig. 19-169D shows that a large value of resistance is connected between ground and the gate element. The voltage developed across this resistor by the photocurrent is amplified by the PFET. The output from this device is of low impedance, and the noise level is that of the FET alone. No noise is generated by the photosensitive portion of the device.

The photosensitive characteristics of a PFET is expressed in radiometric terms rather than those relative to the visual sensitivity of the human eye.

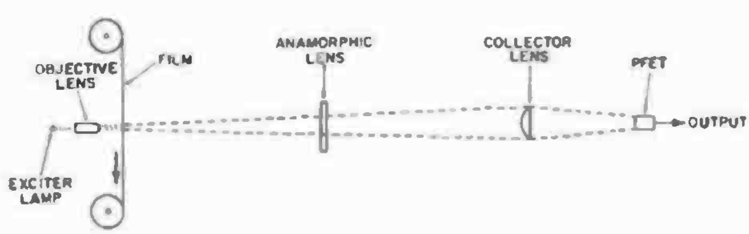


Fig. 19-169B. Side view of Century Projector Corp., anamorphic photo field-effect transistor sound-reproducing system optical train.

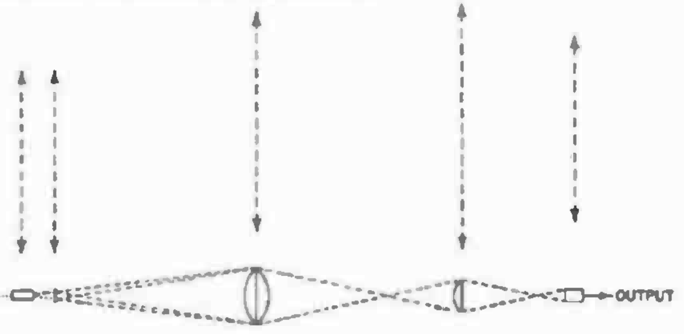


Fig. 19-169C. Top view of optical train.

Thus, light intensity is expressed in watts per square centimeter. The photocurrent increases linearly with an increase of light intensity. The signal-to-noise ratio is constant, regardless of frequency.

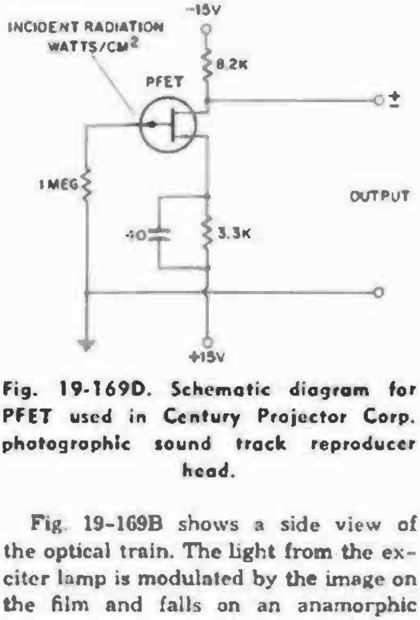


Fig. 19-169D. Schematic diagram for PFET used in Century Projector Corp. photographic sound track reproducer head.

Fig. 19-169B shows a side view of the optical train. The light from the exciter lamp is modulated by the image on the film and falls on an anamorphic lens. It is passed without modification to a collector lens, and then is projected on the surface of the PFET. The collector lens forms an image that just covers the surface of the PFET. As this light beam is not an image of the sound track, it becomes a variable-intensity source of light. Thus, the image of the sound track is eliminated.

Fig. 19-169C is a top view of the optical train. The light from the film diverges and has a width of the total sound track. However, by placing the anamorphic lens as shown, the light is made to image at the collector lens which reduces it to a size that fits the photosensitive surface of the PFET. The result is that the total modulated light from the film is made to activate the full surface of the PFET, thereby increasing the signal-to-noise ratio.

It is claimed that this system of reproduction from optical sound track obtains increased signal-to-noise ratio, dynamic range, faster response, and better linearity. Also, attendant difficulties with high-impedance devices are eliminated.

Stereophonic sound systems using two optical sound tracks side-by-side employ such optical trains as described above. (See Questions 19.104 and 11.161.)

19.170 What is a solid-state sensor? —It is a solid-state device used in optical film reproducers to replace the conventional phototube. The device operates on the solar-cell principle and thus requires no energizing potential. It is mounted in line with the optical sound track, a few thousandths of an inch from the surface of the film. The output level is quite high and requires less amplification than a phototube. The signal-to-noise ratio compared to a phototube is considerably greater. Such devices may be used with any type optical sound-track reproducer.

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Section 20

Loudspeakers, Enclosures, Headphones, and Hearing Aids

It might be said that the forerunner of the loudspeaker was the telephone receiver. This was followed by the addition of various types of horns to provide acoustical amplification. The direct radiator dynamic loudspeaker, which was developed by Chester Rice and E. W. Kellogg in 1925, set the pace to follow in this field. In 1931, Fredrick, of Bell Telephone Laboratories, demonstrated a divided-range (2-way) system. Later, Hilliard, Lansing, Shearer, and Stephens all made valuable contributions.

This section deals with the mechanical, electrical, and acoustical design of loudspeakers and enclosures, as well as their placement. Crossover networks, phasing and impedance matching, characteristics, and measnrements are discussed. Also, types and characteristics of headphones and hearing aids are reviewed.

20.1 Describe a loudspeaker of the direct-radiator type.—It is a speaker in which a diaphragm is directly coupled to the air mass and radiates directly into the listening area. Direct-radiator loudspeakers were first discussed by Lord Rayleigh in his *Theory of Sound*, Volume Two, published in 1877. This book covered the mathematical treatment of a direct radiator in an infinite baffle. Forty-six years later, the dynamic loudspeaker was invented by C. W. Rice and E. W. Kellogg. Mounting a dynamic loudspeaker in a baffle or enclosure qualifies it as a direct radiator, and this is the basis of design for the majority of loudspeakers manufactured today.

Loudspeakers are electroacoustic transducers used for the purpose of transforming electrical energy into acoustical energy, through the mechanical motion of a diaphragm coupled to the air mass. Among the direct-radiator type of loudspeakers are the electrostatic, crystal, balanced armature, and any type of speaker using a diaphragm. Any system employing only a horn is termed a projector-type system. Many designs of speaker systems make use of both horns and direct-radiator type speakers. A typical example is that of the Klipschorn, illustrated in Fig. 20-65A.

A cross-sectional drawing of a typical loudspeaker unit is shown in Fig. 20-1A. Item A is the frame or basket, supporting a diaphragm B, with a voice coil C attached to the apex of the diaphragm. The outer edges of the diaphragm are supported by a flexible edge D. The voice coil consists of several turns of small copper or aluminum wire, centered over a central pole piece E in a magnetic structure F built in the form of the letter E. The voice-coil assembly is centered over the pole piece by means of spider G. The magnetic lines of flux are indicated by the dotted lines H. The central pole piece E becomes the south pole of the magnetic field and the outer edges of the magnetic structure are the north pole of the magnetic field.

The magnetic flux can be supplied in two different manners. In the first method, the magnetic structure is made of soft iron and is easily magnetized. A field coil I is placed over the central pole piece and energized by an external source of dc current. If the field coil is energized and an audio-frequency signal is applied to the voice-coil leads, the voice coil will react to the fixed magnetic field by moving either inward

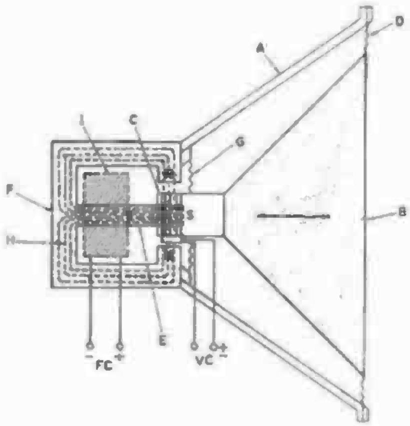


Fig. 20-1A. Cross-sectional construction view of a dynamic loudspeaker.

Fig. 20-1B. Typical single diaphragm, direct-radiator permanent-magnet dynamic loudspeaker unit.

or outward, depending on the instantaneous polarity of the applied audio signal. Such a design would be termed an electrodynamic loudspeaker. This type of design is now obsolete, having been replaced by the permanent-magnet type of speaker.

In the second method, permanent-magnet type speaker units employ a highly charged magnetic structure, rather than a field coil, to supply the magnetic flux. The structure is similar in nature to a field coil, using a central pole piece. The reaction of the voice coil in the magnetic field is the same for either design. A commercial loudspeaker of the permanent-magnet type is pictured in Fig. 20-1B. Several different types of magnetic structures are employed, depending on whether the speaker is of the single diaphragm type or additional diaphragms, such as a two- or three-way speaker unit, are involved. (See Figs 20-3 and 20-6.)

20.2 ***Can more than one speaker unit be used simultaneously?*** — Yes. Some speaker systems use up to six or more separate units, with multiple-section crossover networks to separate the units into frequency bands. It is important in multiple-speaker systems that the individual units be in phase acoustically in respect to each other. When more than one speaker unit is employed, it is termed a speaker system. Multiple-speaker systems offer difficulties in obtaining the proper balance in reproduction unless separate controls are provided for all units except the low-frequency unit. Speaker systems using up to 32 individual speaker units in a single baffle have been constructed. (See Question 20.184.)

20.3 ***Describe a coaxial loudspeaker unit.***—A coaxial loudspeaker is a single unit composed of a low-frequency and high-frequency diaphragm and driven by a single voice coil, as shown in the cutaway illustration of Fig. 20-3. Such speakers make use of a mechanical crossover network and generally divide the frequency spectrum so that the large diaphragm covers the frequency range of 30 to 4000 Hz and the high-frequency element covers from about 2000

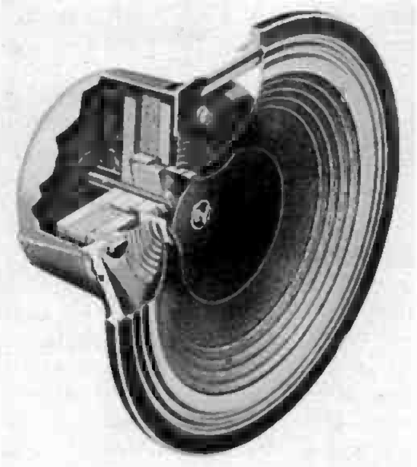


Fig. 20-3. Cutaway showing the interior construction of a coaxial dynamic loudspeaker Model SP15 manufactured by Electro-Voice Inc.

to 15,000 Hz. The crossover frequency is controlled by the size, mass, diameter, and shape of the smaller diaphragm. Low frequencies are not radiated by the smaller diaphragm because of its diameter, and the high frequencies are not radiated by the larger diaphragm because of its decoupling characteristics.

20.4 *Describe a coplanar loudspeaker unit.*—It is a low-frequency unit with two high-frequency units mounted in the center and forward of the low-frequency diaphragm. A crossover network is used to confine the low-frequency unit to a given frequency range. Such a speaker unit, manufactured by the R. T. Bozak Manufacturing Co., is pictured in Fig. 20-4. This unit, a Model B-207A, consists of a single low-frequency unit and two high-frequency units. The crossover frequency is 2500 Hz, with an overall frequency response of 40 to 20,000 Hz. The crossover is accomplished by connecting a 4-μF capacitor in series with the smaller units. The high-frequency units are angled slightly to spread the high frequencies over a wide angle.

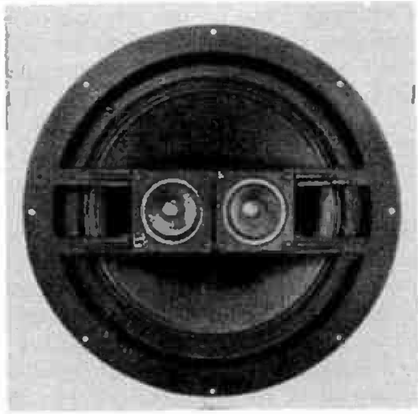


Fig. 20-4. Coplanar loudspeaker unit Model B-207A manufactured by R. T. Bozak Manufacturing Co.

20.5 *Describe the construction of a three-way speaker unit.*—Fig. 20-5 shows

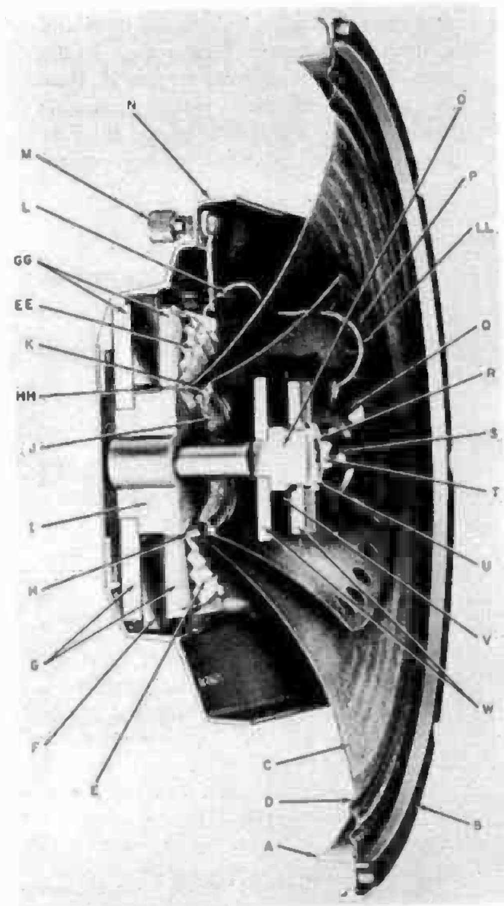


Fig. 20-5. Cutaway view of the University Audio Co. three-way loudspeaker.

a cutaway view of the internal construction of a University Audio Co., three-way speaker unit. Starting at the lower part of the illustration, at A is a die-cast frame. A mounting basket is shown at B. The frame or basket supports the outer periphery of the low-frequency diaphragm C, through a flexible edge or hinge D. The apex of the diaphragm terminates in an inner suspension or spider E. At F is a ring type permanent magnet, with two keeper plates at G. The voice coil for the low- and midrange-frequency diaphragms is shown at H, in the voice-coil gap. A central magnetic pole piece is shown at I, and a dust seal is shown at J. The voice coil with its forming ring is shown at HH. At K is a mass loading ring, which is attached to the midrange diaphragm P. Another view of the inner suspension for the midrange- and low-frequency diaphragms appears at EE. The upper portion of the keeper plates is seen at GG.

The lead at L is a tinsellike wire and runs through a hole in the low- and midrange-frequency diaphragms to the high-frequency unit diaphragm R. Item M is one of two input terminals mounted on the back cover N. Pole piece O is an extension of the central magnetic circuit I for the high-frequency unit diaphragm R. The midrange diaphragm shown at P has a series of holes around its outer edge for the relief of pressure. Tinsel lead LL is shown running up to the high-frequency diaphragm R.

Component Q is a conical loading ring for the high-frequency diaphragm R, with a spherical diffractor at S. Resonance damping is supplied at T, with the high-frequency unit shown at V and the keeper plates at W. Speaker units of this design have a wide frequency characteristic with low distortion and are capable of handling large amounts of integrated program material (IPM).

20.6 ***Describe the construction of a Triaxial loudspeaker.***—*Triaxial* is a registered trade name for a three-way loudspeaker designed and manufactured by the Jensen Manufacturing Division of the Muter Co. The unit consists of three independently driven speaker units combined mechanically into one large unit. A cross-sectional view of the construction is shown in Fig. 20-6A.

Briefly, the unit consists of three independently driven reproducing elements, each element covering a portion

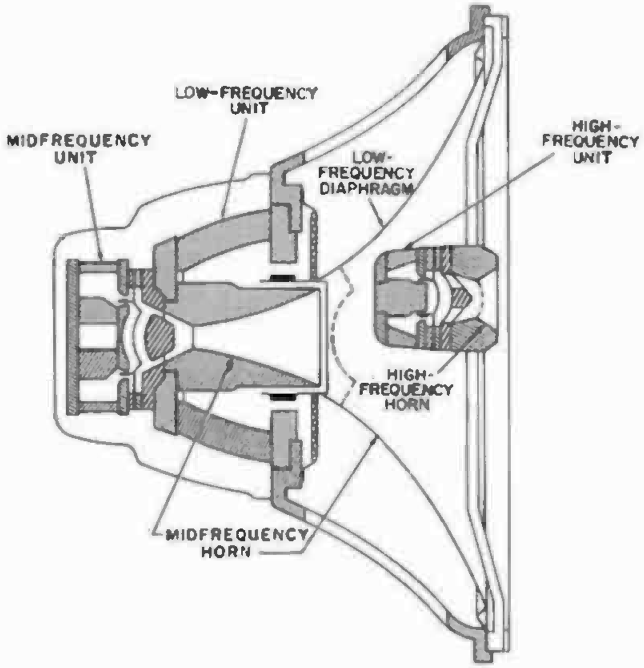


Fig. 20-6A. Cross-sectional view of Jensen Mfg. Co., GS-610 Triaxial dynamic loudspeaker showing three independently driven units.

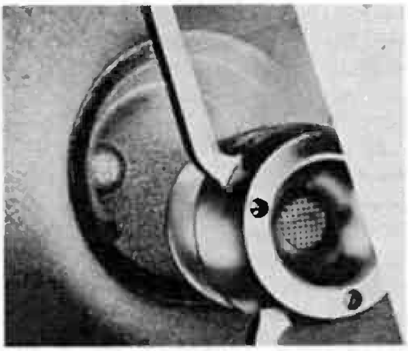


Fig. 20-6B. Intermediate and high-frequency units of the Jenson GS-610 Triaxial loudspeaker unit.

Fig. 20-6C. Jensen Model GS-610 Triaxial dynamic loudspeaker.

of the total frequency spectrum, and a crossover and control network which divides the electrical input into the proper proportions for driving the three separate reproducing elements. Low frequencies up to 600 Hz are reproduced by a heavy, curved-diaphragm, direct-radiator unit. The midfrequency range covers from 600 to 4000 Hz and is reproduced by a compression-driver unit located at the rear of the assembly. This unit feeds a small horn which passes through the center of the low-frequency unit and is designed to act as the articulated final section of the midfrequency horn. Frequencies above 4000 Hz are reproduced by a small compression-driver unit and horn combination at the front of the large diaphragm.

The crossover network and associated controls are housed in a separate unit and connected to the loudspeaker unit by means of a cable assembly. The low-frequency unit is designed to operate in a frequency range where its behavior is essentially that of a piston.

The design of the midfrequency element places the crossover acoustically at 600 Hz, a point of low transient distortion. The low-frequency unit consists of a 15-inch curved diaphragm driven by means of a 3-inch voice coil in a magnetic field of exceptionally high energy in the order of 30,000 ergs. This provides high damping and efficiency. The spider assembly is such that the linearity is maintained with large peak excursions.

The midfrequency driver unit is terminated by a Hypex formula horn (see Question 20.54) the initial section of which passes through the pole piece of the low-frequency motor, while the final section is formed by the curved low-frequency diaphragm. The horn mouth thus has the largest possible size and is limited only by the diameter of the speaker, with attendant low acoustic cutoff. The initial section is visible through the porous dust cap in Fig. 20-6B.

The midfrequency compression-driver unit employs a rigid-shaped plastic diaphragm in conjunction with a sound chamber. The high initial damping of the plastic material suppresses unwanted diaphragm vibrations, thus preventing breaking-up of the diaphragm.

The frequency range above 4000 Hz is reproduced by a compression-driver unit and small Hypex-horn combination at the front of the large diaphragm. Its placement is slightly off-center to the large diaphragm to facilitate phasing and to minimize obstacle effects which are quite important at the high frequencies.

The design of a high-frequency unit covering frequencies from 2000 to 20,000 Hz is rather difficult because of the conflicting factors of lightness, rigidity, the small sound-chamber clearances, and the fact that it must also handle high peak powers. These difficulties have been overcome in this unit by the use of a plastic diaphragm. The overall frequency range is 25 to 20,000 Hz with a peak rating of 80 watts, and a 40-watt rating for integrated program material. The magnetic structure is of Alnico 5 and weighs 6½ pounds. The impedance is 16 ohms.

20.7 ***What are nodes in a loudspeaker diaphragm?***—When a loudspeaker diaphragm is in motion and actuated by a single frequency, the diaphragm will vibrate in different areas called nodes. These nodes may be made visible by observing the motion with a stroboscope or, in some cases, by means of fine sand applied to the surface. The area of vibration is determined by the mass and area of the diaphragm and by the applied frequency.

20.8 ***What is a loudspeaker-motor mechanism?***—The moving elements of a loudspeaker, namely, the voice coil, the diaphragm, and the supports. This term is often used when referring to the loudspeaker actuating elements because of its similarity to an electric motor.

20.9 ***What is compliance in a loudspeaker?***—It is the acoustical and mechanical equivalent of capacitance. The equivalent electrical circuit for a typical horn-driver unit is shown in Fig. 20-9.

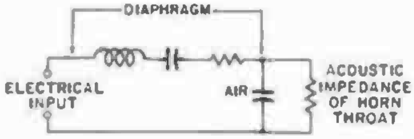


Fig. 20-9. Equivalent circuit for a typical horn-driver loudspeaker unit.

20.10 ***What is a compression-driver unit?***—A loudspeaker unit which does not radiate sound waves directly from a vibrating surface but requires acoustic loading from a horn.

20.11 ***What is horn loading?***—Coupling a loudspeaker diaphragm to the air by means of a horn. Generally, the horn uses an exponential flare starting with a small throat and expanding rapidly to a large bell.

20.12 ***What is back loading?***—A horn coupled to a low-frequency loudspeaker unit in such a manner that the rear surface of the diaphragm feeds the horn, while the front surface radiates directly into the listening area. (See Question 20.71.)

20.13 ***What is a tweeter?***—A name used in audio jargon for a high-frequency speaker unit in a multiple-speaker system.

20.14 ***What is a woofer?***—The name given to a low-frequency loudspeaker used in a multiple-speaker system. Low-frequency speaker units are generally of the direct-radiator type; however, they may also be used with a bass-reflex enclosure or a loaded horn.

20.15 ***Describe the design and construction of an electrostatic loudspeaker?***—An acoustic transducer consisting of two pieces of metallic foil separated by a sheet of dielectric. A polarizing voltage is applied to the foils to maintain a steady attraction between them. Audio-frequency voltages are superimposed on the polarizing voltage and may either add to or subtract from the polarizing voltage, thus causing the foils to move in accordance with the waveforms of the applied audio-frequency voltage. The movement of the foil causes a disturbance of the air which, in turn, generates sound waves.

It is claimed by the designers of electrostatic loudspeakers that certain basic disadvantages of the cone-type loudspeakers, particularly with respect to the propagation of acoustic energy at the high frequencies, are overcome because cone-type loudspeakers driven by a voice coil attached to the center of the diaphragm fail to act as a piston at the middle and high frequencies. Because of this breakup at the higher frequencies, the voice coil does not control the diaphragm motion, and the result is a lack of correspondence between the electrical input and the acoustic output.

One of the advantages claimed for the electrostatic loudspeaker is that it has a diaphragm which is driven equally at all points of its surface. Breakup is eliminated, and harmonic distortion and phase differences are reduced. Because of the design, the diaphragm can be made essentially massless (or extremely low) compared to the air load on the diaphragm. This permits the loudspeaker to have a good high-frequency and transient response which, for all practical purposes, is peakless.

As a rule, electrostatic loudspeakers are made to operate as push-pull transducers, because they are essentially linear in operation and free from waveform distortion, producing neither even nor odd harmonics. Electrostatic loudspeakers may be constructed in several different ways. Two of the most important are:

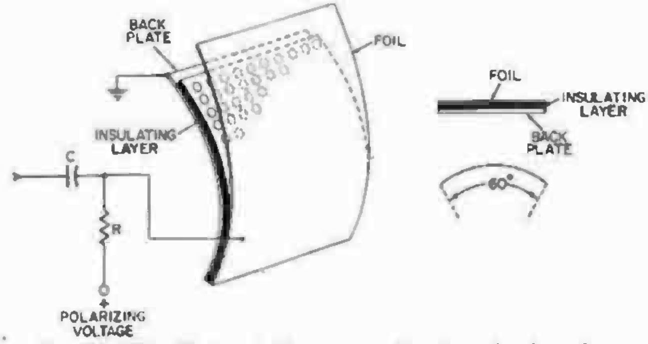


Fig. 20-15A. Electrostatic or capacitor type loudspeaker.

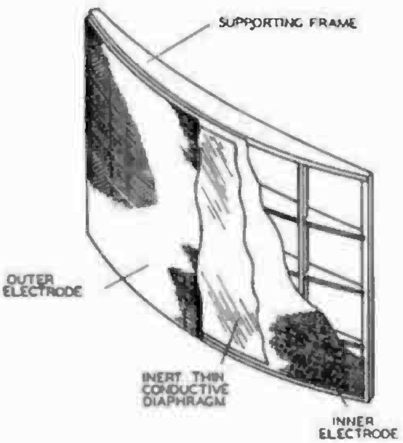


Fig. 20-15B. Cutaway showing the internal construction of an electrostatic loudspeaker.

1. Stretching the diaphragm between supports around its periphery and leaving an air gap between the diaphragm and the two electrodes, (Fig. 20-15A) or,
2. Using an inert diaphragm which is supported by a large multiplicity of tiny elements disposed across the entire surfaces of the two electrodes. These elements act as spacers to hold the diaphragm in the center between the electrodes (Fig. 20-15B).

In this latter type of loudspeaker, the diaphragm is a thin sheet of plastic on which has been deposited a very thin layer of conductive material. It is supported by a multiplicity of small elastic elements which hold the diaphragm but permit it to follow the audio-signal

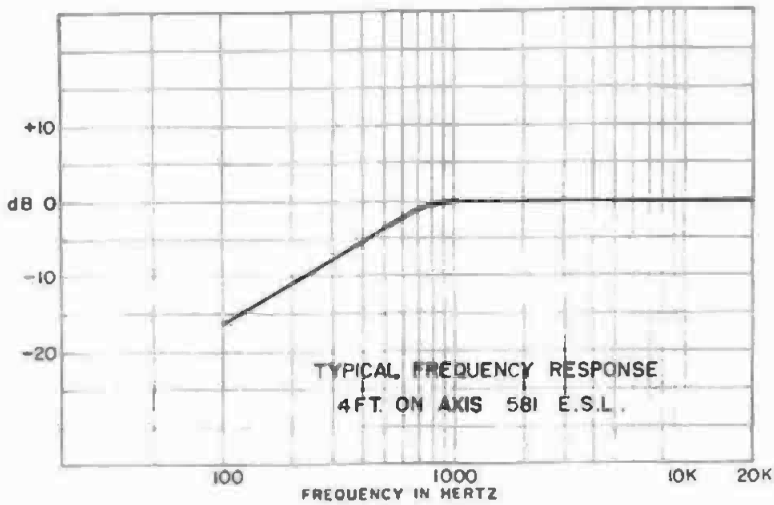
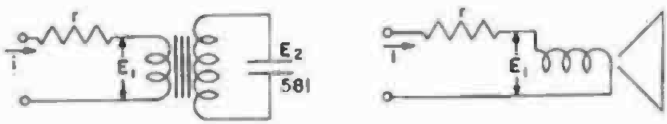


Fig. 20-15C. Acoustic output of an electrostatic loudspeaker. (Courtesy Pickering and Co., Inc.)

waveforms. The electrodes on each side of the diaphragm are acoustically transparent to avoid pressure effects from the trapped air as well as to permit the acoustic energy to move away from the diaphragm. This type of construction permits the diaphragm to be made almost any size required. The performance-per-unit area is the same for any area of the diaphragm.

The actual loudspeaker is a plane surface curved in the horizontal plane —a section of a cylinder. A surface that is large with respect to the wavelength becomes increasingly directional as a propagator at the high frequencies. A large surface, such as an electrostatic loudspeaker diaphragm, projects a large portion of this high-frequency energy outward at right angles to the plane



FREQUENCY Hz	ELECTROSTATIC SPEAKER (VOLT AMPERES)	4—12" DYNAMIC CONE SPEAKERS IN INFINITE BAFFLE (VOLT AMPERES)
700	7	14
1000	4.75	20
2000	10	6.7
3000	8	7

CHART SHOWING RELATIVE POWER REQUIRED TO PRODUCE A SOUND PRESSURE OF 3.3 DYNES PER CM^2 AT A DISTANCE OF 4 FT. ON AXIS.

Fig. 20-15D. Power comparisons of an electrostatic loudspeaker compared to four dynamic speakers mounted in an infinite baffle.

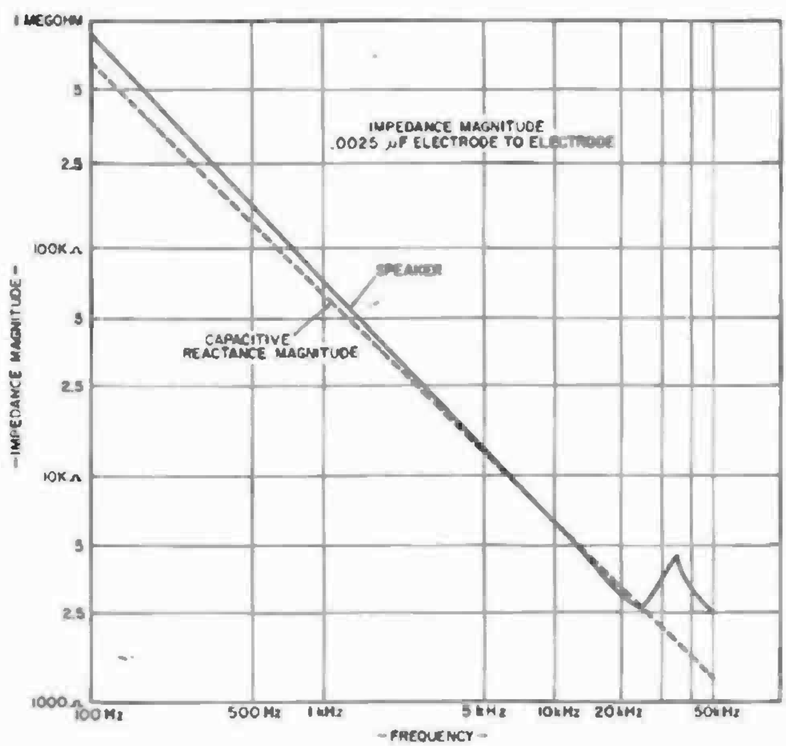


Fig. 20-15E. Impedance characteristics of an electrostatic loudspeaker. (Courtesy, Pickering and Co., Inc.)

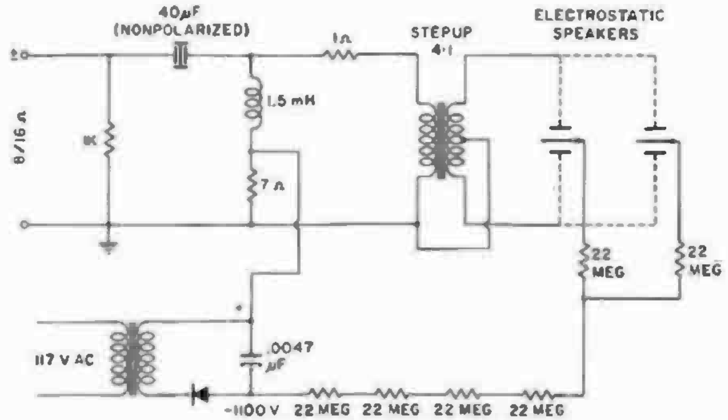


Fig. 20-15F. Typical coupling circuit and high-voltage power supply for electrostatic loudspeakers. The source impedance may be 8 to 16 ohms.

of the surface. By curving the plane in the horizontal direction, an even dispersion of high frequencies is obtained over an angle of 55 degrees.

The vertical pattern is equal to the vertical dimension of the diaphragm, since the diaphragm is a flat plane in the vertical direction. Thus, by controlling the radius of curvature in the horizontal plane and the vertical height of the diaphragm, it is possible to control the dispersion pattern over a large area.

Since an electrostatic loudspeaker is designed to couple directly, in effect, with the air resistance, the mass of the diaphragm, as mentioned previously, can be neglected. The velocity of the diaphragm is directly proportional to the electrostatic force applied, except as affected by the stiffness of the diaphragm suspension. Measurements indicate that for a constant voltage applied to the electrodes, the acoustic response is uniform (flat) to well beyond the range of human hearing. (See Fig. 20-15C.) A slight rise in the impedance curve is observed around 35 kHz. (See Fig. 20-15E.)

The frequency response at the low-frequency end is limited by the maximum linear amplitude of the diaphragm motion as determined by the spacing and the stiffness of the suspension. The power input to the speaker must be increased to overcome this and to maintain sound pressure at low frequencies with a diaphragm area large enough to move an adequate volume of air for this purpose.

The maximum power output from an electrostatic loudspeaker of a given diaphragm area is determined by the strength of the electrostatic field that can be produced between the diaphragm and the electrodes. Therefore, a dc polarizing voltage is applied to the plates. The electrostatic field is the sum of the field produced by the polarizing voltage and the peak signal voltage superimposed on the polarizing field. Polarizing voltages of 1000 to 2000 volts dc are common. A second function of the polarizing voltage is the prevention of frequency doubling. If a polarizing voltage is not employed, when a frequency of, for example, 2000 Hz is applied to the speaker, the results could be a distorted 4000 Hz. This is caused by the capacitor

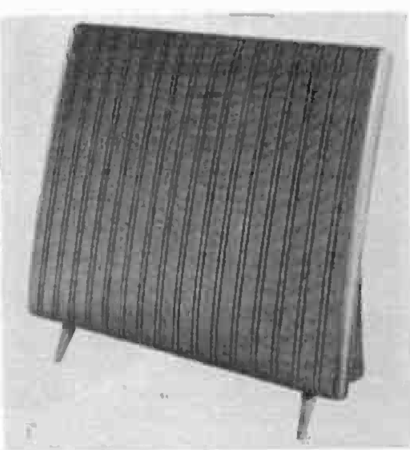


Fig. 20-15G. Quad electrostatic loudspeaker manufactured by Acoustical Mfg. Co., Ltd., England.

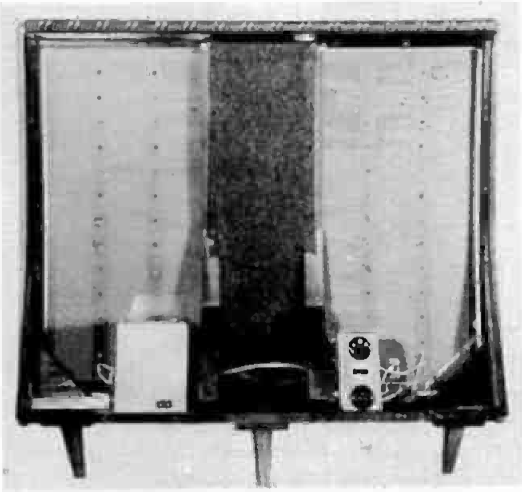


Fig. 20-15H. Rear view of Quad electrostatic loudspeaker, showing the curved plate assembly for supporting the membranes, the amplifier, and power supply. (Courtesy, Acoustical Mfg. Co. Ltd., England)

effect. If a sine-wave signal is applied to the plates without a polarizing voltage, as the difference of potential increases toward the peak of the sine wave, the movable plate is attracted to the fixed plate. With the decreasing potential on the downward side of the sine wave, the electrostatic force between the plates decreases, the movable plate returns to its original position, and the sine-wave voltage is zero. This motion of the movable plate has produced a single air motion for the positive half of the sine wave.

On the negative half of the cycle, the voltage rises to a peak again and applies a voltage of opposite polarity to the plates. The plates are again attracted to each other. The movable plate goes through the same action and eventually returns to the zero position. Thus, it may be seen that two pulses, both moving in the same direction, have been obtained for a single cycle of the applied sine wave—frequency doubling. With the polarizing voltage applied, a steady electrostatic force is created be-

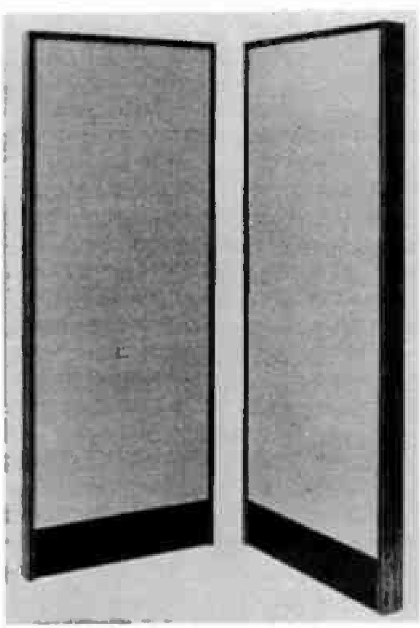


Fig. 20-15I. Acoustech-X stereophonic electrostatic loudspeakers. Each panel contains its own individual solid-state amplifier and power supply.

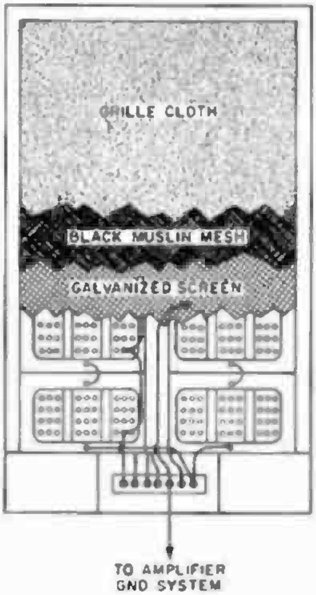


Fig. 20-15J. Cutaway showing the internal construction of an Acoustech-X electrostatic loudspeaker panel.

tween the plates, and the movable plate is slightly attracted toward the fixed plate.

Applying a sine wave (plus), the movable plate is attracted to the fixed plate beyond its position fixed by the polarizing voltage. Upon reversal of the sine wave, the electrostatic force between the plates is reduced and the movable plate returns to its zero position. On the next half of the cycle (negative), the polarizing voltage is reduced because the sine wave is of opposite polarity, and the attracting force is decreased below the polarizing voltage value. Therefore, the plate moves away from the fixed plate, completing the cycle. In this manner, frequency doubling has been eliminated and the speaker produces a sine wave similar to the conventional piston-type loudspeaker. By connecting the plates for push-pull operation (Fig. 20-15F), the signal is split between the two sides of the speaker, and distortion is further reduced.

An electrostatic loudspeaker is nothing more than a capacitor; the internal capacitance is in the order of 0.0025 μF from electrode to electrode. Thus, the impedance presented by the loudspeaker to the output of the amplifier falls off at a constant rate of 6 dB per octave as the frequency is increased. This precludes the possibility of maintaining a constant voltage at the loudspeaker input for high volume levels when using a conventional amplifier. If the matching transformer is designed for the most efficient transfer of power at the high frequencies, insufficient signal voltage will be available at the middle frequencies.

It is a well-known fact that the distribution of energy above 2500 Hz falls off at a rate of 6 dB per octave, and this fact is taken advantage of in the RIAA disc-recording curve established in 1953. (See Question 13-169.) This characteristic permits pre-emphasis of the high frequencies during recording, with a corresponding reduction (post-equalization) during reproduction, resulting in a considerable improvement in the overall signal-to-noise ratio. Thus, the reproduction falls off at a rate of 6 dB per octave, an inverse of the recording characteristic.

As the high frequencies fall off, so does the impedance of the loudspeaker, so that the current through it is maintained constant. This means that the power required to operate the loudspeaker will also fall at a rate of 3 dB per octave (power equals voltage times current), making it possible to couple the loudspeaker for the best transfer of energy around 3000 Hz. Thus, the frequencies above 3000 Hz will not overload the amplifier.

With the energy distribution as described above together with the diaphragm area and other characteristics of the design, the maximum power requirements will occur around 3000 Hz. Using this assumption, sufficient voltage drive for the loudspeaker will require approximately 44 watts of audio power for maximum acoustic outpnt. The power required to drive an electrostatic loudspeaker to full output goes up in direct relation to the area of the diaphragm. Electrostatic loudspeakers are, as a rule, designed to start operating at either 400 Hz or 1000 Hz and to be effective up to the limit of audibility.

An adapter unit is required that will supply a polarizing voltage and a dividing network for separating the audio-frequency spectrum into two parts, one for tne low-frequency loudspeaker and the other for the electrostatic unit. The adapter matches the electrostatic loudspeaker to the amplifier output and provides a means of controlling the frequency balance between the low-frequency and the electrostatic high-frequency unit. As a rule, these adapters are designed to operate from the normal 16-ohm output winding of a conventional output transformer. The low-frequency loudspeaker is fed from the adapter. The electrostatic unit is physically located above the low-frequency loudspeaker and to one side, or separated some distance—the latter will be dependent on local acoustic conditions.

Electrostatic loudspeakers do not always operate satisfactorily with all type amplifiers and may cause oscillation of the amplifier at the high frequencies. The reason for this is the reflected capacitive load of the loudspeaker to the plates of the output stage of the amplifier. This subject is discussed in Question 12.167 and in Question 23.125.

In Fig. 20-15D is shown a comparison of the power required for an electrostatic loudspeaker to produce a sound pressure from the diaphragm of 3.3 dynes per square centimeter at a distance of four feet, compared to the same pressure from four dynamic cone-type loudspeakers in an infinite baffle at a distance of four feet. The impedance characteristics of the electrostatic loudspeaker are shown in Fig. 20-15E. In Fig. 20-15F is shown the basic circuit used for coupling an electrostatic loudspeaker of push-pull design to the output of a conventional amplifier.

20.16 ***What is an ionic loudspeaker?*** —A nonmechanical loudspeaker using no diaphragm. Air particles in an acoustical chamber are agitated by an electrostatic field which is varied in accordance with the applied audio frequencies. Vibration of the air particles in the acoustic chamber is radiated through a horn coupled to the surrounding air.

The air particles are ionized by molecular collision induced by thermal agitation in an ultrasonic field. The audio-frequency signal is superimposed, as a modulation voltage, on the electrostatic field voltage. The frequency of the field voltage is approximately 27 mHz. The loudspeaker system is two-way and uses both low- and high-frequency agitating units. This loudspeaker is a development of the French loudspeaker manufacturer Audax.

20.17 ***What is an apex dome?*** — Since the greater part of the high-frequency radiation is from the apex of the diaphragm, a small dome is placed over the opening to increase the radiations at the middle and high frequencies. The curved dome spreads the high frequencies over a wide area, thus increasing the width of the polar pattern at the high frequencies. It also functions to keep dirt out of the magnetic gap.

20.18 ***What does the term "whizzer" mean?***—It is a small cone attached to the apex of a single-diaphragm dynamic speaker unit to increase the midrange and high-frequency response. It is generally used on medium-priced coaxial speaker units. An Electro-Voice coaxial speaker using a whizzer is pictured in Fig. 20-18.

20.19 ***What is a general-purpose speaker unit?***—It is a single-diaphragm-type speaker unit that can be used for the reproduction of both speech and music. As a rule, such units are mounted in a simple open-back baffle

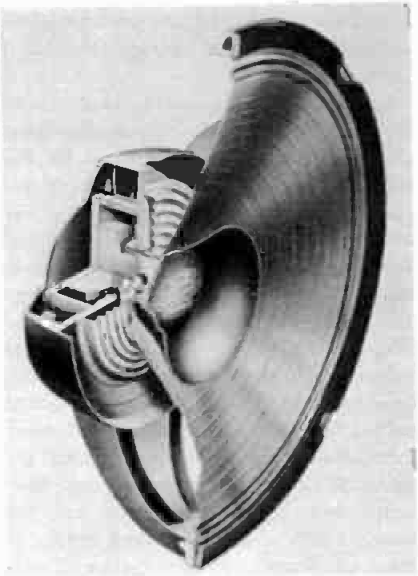


Fig. 20-18. Electro-Voice dynamic speaker unit with a whizzer attached to diaphragm to increase the midrange and high-frequency response.

and employ an 8-inch diaphragm. Although larger speakers reproduce the low frequencies better, they are lacking in high-frequency reproduction. Also, if speech is important, the voice may sound "tubby." It has been found that where a general purpose speaker is required, an 8-inch speaker, when mounted in a simple open-back baffle, has the best overall reproduction for both speech and music. If the baffle is wall mounted, a 1-inch piece of *Fiberglas* should be placed behind the speaker to reduce the effects of resonance.

20.20 ***What is a projector-type loudspeaker?***—It is a loudspeaker unit attached to a horn. The horn may be looked upon as an impedance-matching transformer which couples the heavy diaphragm of the driver unit to the air-mass at the bell of the horn. The horn-type loudspeaker is the most efficient device for the reproduction of sound (see Question 20.51). Horns may be used as individual units or in speaker systems employing other types of speaker units.

20.21 ***What methods are used for centering a voice coil?***—In a well-centered diaphragm, the voice coil is balanced, both mechanically and electrically, within the magnetic gap. The device for centering the coil in the gap is called a spider, and it holds the coil to within a few thousandths of an inch for a high-frequency unit, and about 10 to 15 thousandths for a low-frequency unit. The spider plays an important role in the operation of the speaker; it keeps the coil in the exact center of the gap. Thus the coil will move in a linear manner while maintaining its proper spacing.

Two types of spiders in general use are a flexible spiral type made of bakelite, and an impregnated flexible cloth type. In the first 15-inch dynamic loudspeaker systems used in the early RCA Photophone motion picture sound installations, the voice coil was held in the magnetic gap by three heavy threads. These threads performed remarkably well, although they required frequent recentering. (See Question 20.185.)

20.22 ***What is Alnico?***—The trade name of a magnetic alloy used in the construction of loudspeakers and other devices requiring a strong permanent magnetic field. The alloy consists principally of aluminum and nickel.

20.23 ***At what rate do permanent magnets lose their charge?***—Permanent magnets lose about 1 percent of their charge when they are first energized. After that, they lose about 0.2 percent, for a total of approximately 1.2 percent within the first year. From then on, the loss is on the order of 1 percent in the next several thousand years, assuming that the magnet is not abused. (See Question 24.85.)

20.24 ***What is the purpose of the pierced ring in the center of a loudspeaker diaphragm?***—It is a form of acoustic lens for spreading the high frequencies over a greater dispersion angle than that normally obtained from a conventional loudspeaker. Acoustic lenses are discussed in Question 20.74.

20.25 ***Describe the construction of a duplex loudspeaker unit.***—A duplex is a two-way speaker unit, consisting of a low-frequency speaker with a small multicellular horn mounted in the center of the large diaphragm for high-frequency reproduction (Fig. 20-25A). A cross-sectional view of this speaker unit, showing the interior construction, is given in Fig. 20-25B, with the principal components indicated.

At A is the frame or basket, with a 15-inch low-frequency diaphragm at B, supported by a high compliance edge C. Item D is a felt dust barrier, and E is a six-cell multicellular horn. An exponential throat F extends from the rear of the horn. A high-frequency phasing plug is seen at G, an aluminum diaphragm at H, with a 1¾-inch edge-wound voice coil shown at I. Component J is an acoustical loading cap. The magnet for the high-frequency unit is shown at K, and the low-frequency unit is at L. At M is the low-frequency voice coil, and its spider is at N.

The acoustic plug G is machined with four exponential slots to provide the

Fig. 20-25A. Altec-Lansing Model 604E duplex loudspeaker unit.

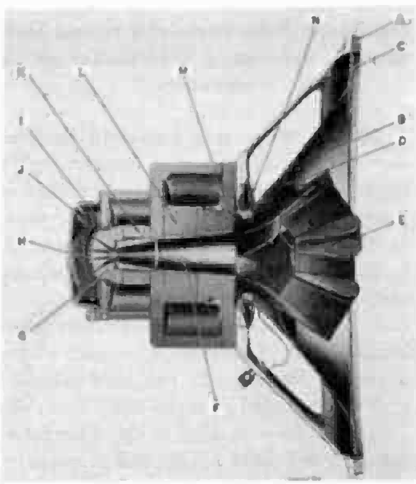


Fig. 20-25B. Cutaway showing the internal construction of Altec-Lansing Model 604E duplex loudspeaker.

proper phase relationships between the sound waves emitted from the center and outer edges of the high-frequency diaphragm, thus providing a smooth midrange and high-frequency response. Both the low- and high-frequency units are mechanically, electrically, and magnetically independent. The voice coil for the low-frequency unit is 3 inches in diameter and is made of edge-wound copper.

The magnetic structure weighs 26 pounds. The crossover frequency is 1500 Hz. The resonant frequency of the cone is 25 Hz. The speaker is designed to operate from a 16-ohm source impedance. Frequency range is from 20 to 22,000 Hz. Power capability is 35 watts continuous, with 50 watts peak.

20.26 What is a polar curve for a loudspeaker?—A circular curve plotted to show the angle of radiation of a loudspeaker with respect to frequency for a given power input at the voice coil terminals. A typical polar curve is shown in Fig. 20-26.

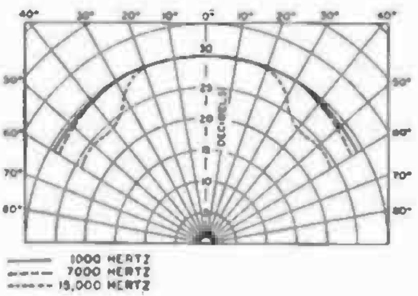
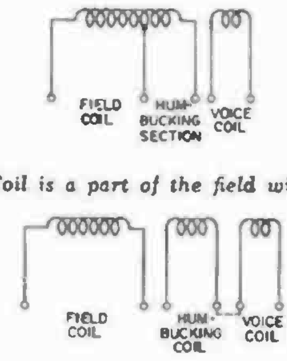


Fig. 20-26. Polar curve of a typical high-quality, wide-range loudspeaker in an enclosure.

20.27 What is a free-field pattern?—A frequency-response pattern of a loudspeaker made in open air away from reflections or in an anechoic room as described in Question 2.83.

20.28 What is a hum-bucking coil?—A coil mounted at the end of the field-coil pole piece in an electrodynamic loudspeaker. The coil is wound in the opposite direction to the field winding and is connected in series with the voice coil, as shown in Fig. 20-28. The function of the hum-bucking coil is to cancel ripple voltage induced by the dc supply to the field coil. In radio sets where the field coil is used for a filter choke, the ripple voltage may be high enough to be heard unless a hum-bucking coil is used. The coil cancels the 120-Hz ripple voltage from a fullwave rectifier circuit; however, it has no effect on the frequency response of the loudspeaker. Hum-bucking coils are not used with permanent magnet loudspeakers.



(a) Coil is a part of the field winding.

(b) Separate coil connected in series with the voice coil.

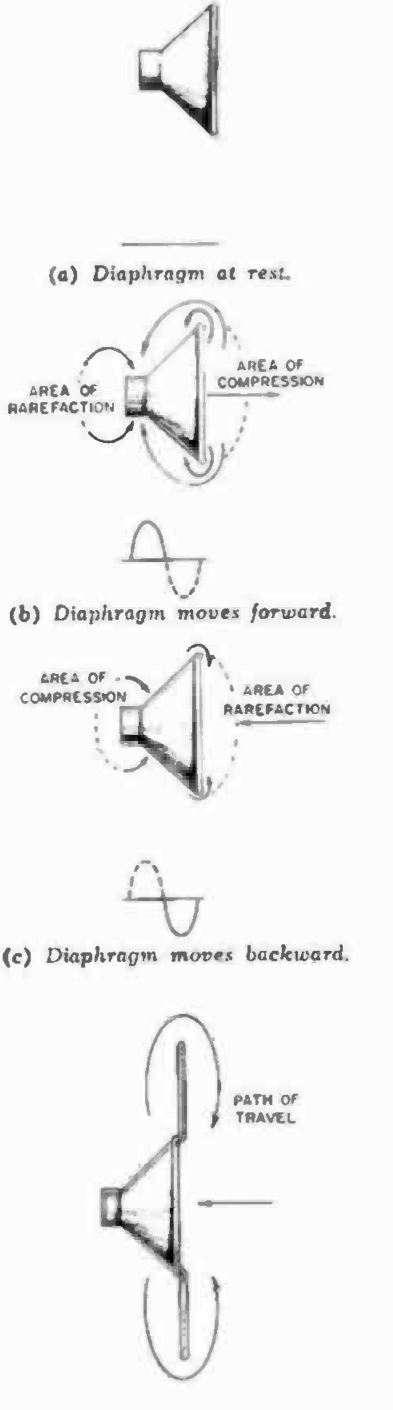
Fig. 20-28. Hum-bucking-coil circuits for electrodynamic loudspeakers.

20.29 Describe the action of a single-diaphragm, radiator-type loudspeaker.—In Fig. 20-29 is shown the action of a single-diaphragm radiator, without a baffle, being actuated by a low-frequency sine wave. At (a) the diaphragm is shown at rest. At (b) the diaphragm is caused to move forward. The air at the front of the diaphragm is compressed, and a rarefaction is set up at the rear of the diaphragm.

On the reverse half of the cycle at (c), the diaphragm moves backward. The air at the rear is now compressed and that at the front is decompressed; thus, a partial vacuum is created on one side of the diaphragm.

When the air is compressed on one side of the diaphragm, the air from that side rushes to the decompressed side in an attempt to equalize the pressures, thus cancelling the sound waves emanating from the opposite side. The result is that very little energy in the form of a sound wave is created.

At the higher frequencies, the diaphragm is moving at a very high rate, the wavelengths are short, and the air does not have time to travel from front to back; therefore, sound waves are created and little effect from cancellation is noted. At the lower frequencies (longer wavelengths), the diaphragm moves at a much slower rate, and the



(a) Diaphragm at rest.

(b) Diaphragm moves forward.

(c) Diaphragm moves backward.

(d) Speaker mounted in baffle showing the increased length of travel caused by the baffle.

Fig. 20-29. Loudspeaker operated without baffle, showing how the front and rear sound waves cancel at the lower frequencies.

air has time to travel from one side of the diaphragm to the other in any one cycle, setting up cancellation between the front and back waves. To prevent cancellation at the lower frequencies, a baffle is placed in front of the diaphragm, as shown at (d). Now, the low frequencies must traverse a much longer path to reach the opposite side. If the path is made long enough, cancellation can be eliminated entirely. The baffle may be made in the form of a flat surface, a closed box, a labyrinth, or any one of the various forms described in this section.

Cancellation of the low frequencies may be demonstrated by operating a loudspeaker without a baffle and then placing a two-foot-square flat baffle over the diaphragm. It will be noted immediately that the low-frequency response is increased.

20.30 What is the efficiency of the average dynamic loudspeaker unit?—The efficiency of a dynamic loudspeaker unit can vary from 1 to 10 percent, depending on the design. For highly damped designs, the efficiency may be on the order of 1 to 2 percent, while for other designs, it could be much higher. When several speakers are assembled in an enclosure, the efficiency of the overall design may approach 40 to 60 percent; this again depends on the efficiency of the individual units and the enclosure design. When designing an enclosure, it is advantageous to consult the manufacturer of the speaker units.

20.31 What is a transformerless loudspeaker unit?—This loudspeaker unit is now obsolete. It was designed, some years ago, for direct connection in the cathode of a power-amplifier stage. The voice-coil impedance was on the order of 500 ohms. With solid-state power-amplifier stages, the voice coil of the speaker, with an impedance ranging from 4 to 16 ohms, was, as a rule, connected directly to the output stage.

20.32 Describe the construction of a horn driver unit.—In the early type dynamic loudspeaker, horn driver units employed a dc energized field coil, similar to that discussed in Question 20.1. These units are now obsolete. Because of the design, a cross section of its construction is given in Fig. 20-32A.

A phase-correction plug is placed in the sound chamber to minimize effects caused by the differences in path length

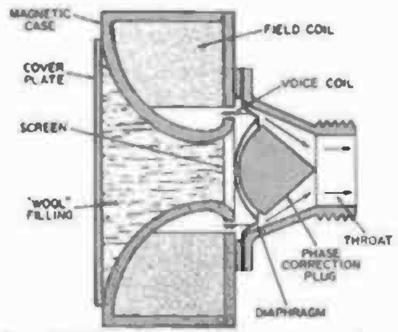


Fig. 20-32A. Horn driver unit employing a conventional diaphragm and field coil. (Field coil now obsolete.)

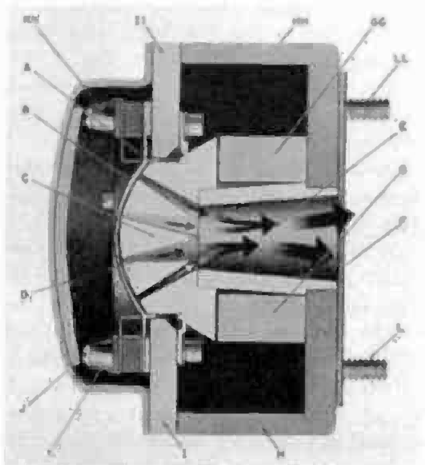


Fig. 20-32B. Interior construction of Altec-Lansing Model 802D horn driver unit.

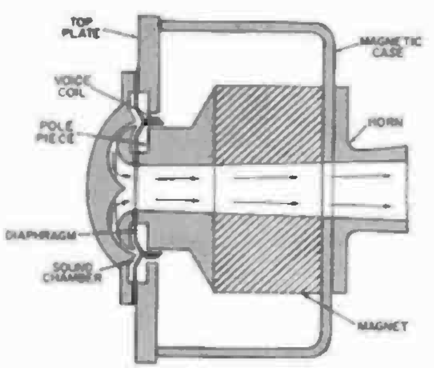


Fig. 20-32C. Cross-sectional view of horn driver unit using an annular diaphragm and a permanent magnet.

from the various areas of the diaphragm to the throat. Thus, interference occurring in the sound chamber at the high frequencies is reduced, and a smoother frequency response is obtained. The dia-

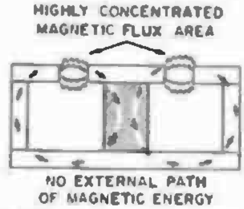


Fig. 20-32D. A permanent magnet mounted in the center of the magnetic circuit.

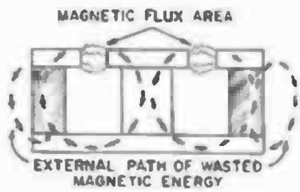


Fig. 20-32E. Magnets mounted at the outer edge of the assembly.

phragm is supported at the outer edges of its periphery, with the voice coil supported at the edge of the dome-shaped center.

Fig. 20-32B shows the interior construction of an Altec-Lansing, Model 802D horn driver unit. At A is a 1¾-inch voice coil would edgewise with aluminum wire and attached to a 2¼-inch domed aluminum diaphragm B, having tangential compliance. A mechanical phasing plug C, with four exponential slots D, is utilized to provide proper phase relationship between sound emanating from the center and outer edges of the diaphragm assembly and to insnre a smooth overall response. The sound is projected throngh an exponential throat E in the center pole piece of the magnetic structure. A protective screen appears at F. The magnetic structure is composed of items G, H, and I. Item J is a dust cover. The speaker connection terminals appear at K and KK. The unit is attached to the horn assembly at two threaded bolts L and LL. A similar, but earlier, design is shown in Fig. 20-32C.

Basic design principles used in the manufacture of horn driver units are given in Figs. 20-32D and E. Fig. 20-32D shows the magnetic circuit for a driver unit using a permanent magnet mounted in the center of the magnetic circuit. In this design, the maximum number of lines of force are obtained. In the structure shown in Fig. 20-32E, many of the magnetic lines of force are lost. In either design, the magnets generally consist of an Alnico alloy assembled as a unit, then magnetically charged. Thus, the maximum lines of force are obtained.

20.33 How are loudspeakers rated relative to power and frequency response?—The power-handling capabilities of a loudspeaker are generally based on the power of a complex waveform termed, "integrated program material" (IPM). Therefore, a speaker unit rated at 20 watts means that it will carry 20 watts of IPM without damage. If 20 watts of sine-wave power is applied to a speaker with this rating, it could severely damage the voice-coil structure.

The power rating of a loudspeaker has no bearing on the frequency response or the amount of distortion generated. At the lower frequencies, the physical dimensions of a loudspeaker unit are small compared to those required acoustically; thus, the unit becomes a low efficiency radiator with high distortion. With the development of new type enclosures, magnetic structures, cone materials, efficiency, and lower distortion, there has been a very marked improvement over the past few years. Standards for measuring the characteristics of loudspeakers are given in the EIA Standard SE-103-1949. These were reaffirmed in 1954.

A technique used by the Bell Telephone Laboratories for rating the power capabilities of direct-radiator-type units is to apply a sweep frequency of uniform amplitude from 50 to 1000 Hz to the unit at the calculated power-handling ability of the unit. If after 100 hours of operation there is no failure, the maximum power used during the test is considered to be the power rating of the unit. For horn-driver units, a frequency 100 Hz below the cutoff frequency with a high frequency of 2000 Hz is used. When making such tests, the unit must be used with the proper horn loading.

20.34 If the output power to a loudspeaker is doubled, what is the increase in acoustic output power?—The increase is 3 dB, if the acoustic output power can be assumed to be in free space.

20.35 How does the Doppler effect enter into a loudspeaker frequency characteristic?—The pitch of a given tone will rise as the diaphragm moves toward the listener and fall as it moves away. This effect is independent of the nonlinearity characteristics of the speaker involved. This effect has long been a bone of contention among loudspeaker manufacturers. (See Question 1.124.)

20.36 Define the term "Lim."—The term "lim" was derived from the word "liminal." It was used several years ago for rating loudspeaker units, relative to their frequency response based on the results of a group of critical listeners. This term is now obsolete.

20.37 What are the directional qualities of a low tone?—Low frequencies spread out from the source and bend around any object in their path of travel. This is called diffraction. (See Question 2.25.)

20.38 What are the directional qualities of a high-frequency tone?—High frequencies tend to travel in a straight line from the source in a manner similar to a light beam. The higher the frequency the sharper is the beam effect. When a high frequency encounters an object in its path, an acoustic shadow is cast, and the area behind the object becomes completely dead.

20.39 What is spatial effect?—An illusion on the part of the listener that the original program was recorded in a large auditorium. This effect may be obtained by the use of a reverberation chamber with recording, as described in Question 2.82.

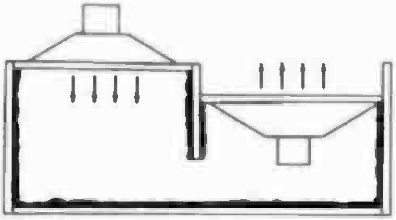


Fig. 20-40. A push-pull loudspeaker enclosure unit.

20.40 What is a push-pull loudspeaker system?—An enclosure housing two or more (even numbers) loudspeakers placed as shown in Fig. 20-40. The diaphragm movement must be so phased that the diaphragms are operated acoustically in phase and electrically out of phase.

20.41 Describe the construction of ceramic and crystal-type loudspeaker units.—The basic ingredients of ceramic materials used in loudspeaker construction are discussed in Question 14.43. A crystal transducer element generally consists of Rochelle salt. In either type loudspeaker, the diaphragm is connected by a mechanical linkage to the transducer element (Fig. 20-41A). When audio frequencies are applied to a crystal transducer, because of the piezoelectric effects of the crystal, the transducer is caused to bend or twist. This bending transmits mechanical motion to the diaphragm and to the air, thus setting up sound waves.

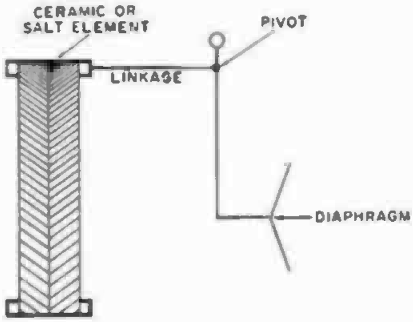
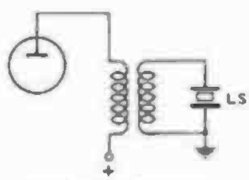
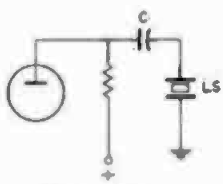


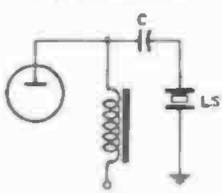
Fig. 20-41A. Drive mechanism of a ceramic or crystal loudspeaker.



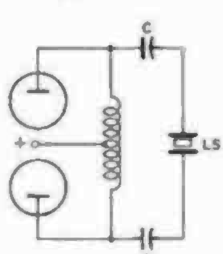
(a) *Transformer.*



(b) *Resistance.*



(c) *Choke.*



(d) *Push-pull choke.*

Fig. 20-41B. Methods for coupling a crystal loudspeaker to the output stage of an amplifier.

Ceramic driving elements do not have piezoelectric properties in their original state. The motion imparted to the diaphragm is greatly dependent on the orientation of the applied force or electric field, with respect to the axis of the material.

Because of the high impedance of the crystal or ceramic loudspeaker, these elements are generally confined to hearing-aid use, very small radio receivers, or uses where other designs of loudspeakers would be impractical. Application of large amounts of power to a crystal loudspeaker can result in the shattering of the crystal element.

Ceramic or crystal londspeakers can be coupled to the output of an amplifier by the use of an output transformer with a high-impedance winding, but this is rather impractical. A more convenient method is to couple to the element through a low-leakage capacitor, to isolate any dc flowing in the output circuit. Typical coupling circuits are given in Fig. 20-41B. (See Question 20.78.)

20.42 *What are the electrical equivalents of a loudspeaker?*—The electrical equivalents of a loudspeaker unit vary and depend on the type enclosure in which the speaker is mounted. Briggs has shown that if the speaker is mounted in an open-back baffle, it is electrically equivalent to a 20-ohm resistor in series with a 79-μF capacitor. If mounted in an enclosure of two cubic feet, the electrical equivalent is that of a 24-ohm resistor in series with 33-mH inductance. These data were arrived at by using a speaker with foam rubber suspension construction. Thus, it can be seen that when discussing the electrical quivalent of a loudspeaker mechanism, the type enclosure must also be stated.

20 43 *What is a balanced-armature loudspeaker?* — A loudspeaker unit in which the soft-iron armature is balanced and centered in a permanent magnetic field. See Fig. 20-43. The ar-

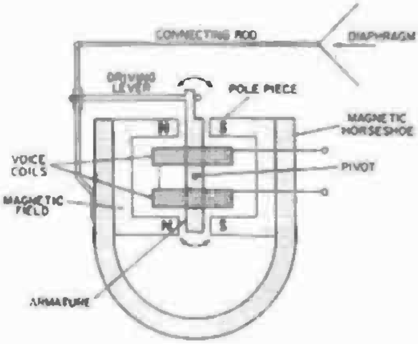


Fig. 20-43. Balanced-armature loudspeaker (magnetic).

mature is enclosed within a coil to which audio frequencies are applied.

When the audio-frequency currents are induced in the coils, the armature is either attracted or repelled by the fixed magnetic field. A rod attached to the armature is connected to a diaphragm which is moved by the action of the armature, setting up sound waves. This design was used extensively in the early-day loudspeakers before the advent of the dynamic loudspeaker. The impedance was generally 2000 ohms or greater. It is also referred to as a magnetic loudspeaker.

20.44 What is a capacitor loudspeaker?—An electrostatic loudspeaker as described in Question 20.15.

20.45 What is an infinite or flat baffle?—A large flat surface used as a separator to increase the distance between the sources of sound (the front and back surfaces) of a loudspeaker diaphragm. The dimensions of the baffle must be such that the wavelength of the lowest frequency is small in comparison to the distance of separation. However, this is not always the ideal condition because a baffle of at least 15 square feet is required for satisfactory low-frequency reproduction. (See Question 20.29.)

An example of an infinite baffle is a loudspeaker mounted in a wall and with a small room at the rear. The volume of the room at the rear is large enough to produce no effect on the resonance of the loudspeaker. The coupling of the cone to the air, however, is unsatisfactory in the low-frequency region due to the increasingly high ratio between the wavelength and the cone diameter as the frequency is lowered. This effect results in a loss of low-frequency response because of insufficient loading of the diaphragm.

20.46 What is the ratio of the dimensions for a flat baffle, relative to the lowest frequency to be reproduced?—With the loudspeaker mounted in the center of the baffle, the dimension of any one side should not be less than one-quarter wavelength for the lowest frequency to be reproduced. The wavelength of any frequency may be calculated:

$$\lambda = \frac{V}{F} = \frac{1127}{F}$$

where,

λ is the wavelength in feet,
V is the velocity of sound in air,
F is the frequency in hertz.

The velocity of sound in air is 1127 feet per second at 20 degrees Centigrade. (See Question 1.150.)

20.47 What are typical infinite- or flat-baffle dimensions?

Cutoff Frequency	*Baffle Size Each Side of Center.*
40 Hz	14.0 ft
60 Hz	9.5 ft
100 Hz	5.5 ft

20.48 What is the purpose of placing a loudspeaker off-center in a flat baffle?—To create paths of different lengths between the front and rear surfaces of the loudspeaker diaphragm.

20.49 What is a baffle?—A loudspeaker enclosure developed by H. A. Hartley, using a group of stretched, resilient, sound-absorbing screens in the interior of the enclosure.

20.50 What is an exponential horn?—A horn which has a constant rate of expansion of flare at an exponential rate. (See Fig. 20-50.) The purpose of the horn is to provide an acoustical match between the diaphragm of the loudspeaker unit and the air in the throat of the horn. A horn facilitates the transfer of electrical energy into acoustical energy and, if properly designed, will do so with a minimum of distortion.

The design of a loudspeaker horn is complex and requires careful consideration to prevent reflection of the acoustical energy back into the horn bell.

20.51 What are the factors influencing the design of an exponential

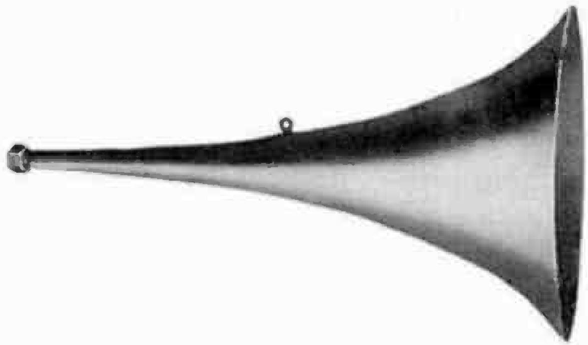


Fig. 20-50. Exponential loudspeaker horn. (Courtesy, Bell Telephone Labs.)

horn?—A horn consists of a throat, a taper or flare, and an end bell or mouth (Fig. 20-51). The design takes into considerations the flare rate between the mouth and the end bell, based on the lowest frequency to be reproduced—the cutoff frequency. Assume that a small horn cluster containing eight horns and similar to that shown in Fig. 20-60 is to be constructed for a two-way speaker system, using a crossover frequency of 800 Hz. The design is based on that of a single horn. After the eight horns have been completed, they are assembled into a cluster as shown in the illustration.

The horn cannot be designed to cut off at exactly 800 Hz but must cover approximately one octave lower in frequency to assure a smooth acoustic crossover. This horn is designed for a cutoff of 475 Hz. This cutoff frequency, although not a full octave below the electrical crossover frequency, will still provide a smooth crossover acoustically, and will hold the physical dimensions down to a reasonable size. The electrical crossover network should be designed to attenuate frequencies below 800 Hz at a rate of 12 dB per octave using a network as described in Section 7. The use of an exponential flare rate will provide a sharp rise in throat resistance at the lower frequencies for a given horn length. (See Fig. 20-55.)

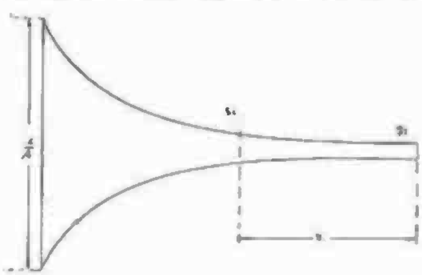


Fig. 20-51. Factors considered in designing an exponential horn.

The cross-sectional area of the horn throat at the driver end will be dependent on the area of the throat of the driver unit. For this illustration, it will be assumed the throat area is 0.140 inch. The cutoff frequency is to be 475 Hz.

$$\lambda = \frac{V}{F} = \frac{13{,}548}{475} = 28.5 \text{ inches}$$

where,

λ is the wavelength in inches,
V is the velocity of sound in air in inches per second based on 1129 feet per second,
F equals frequency in hertz.

The flare constant may now be calculated:

$$M = \frac{4\pi f_c}{V}$$

$$= \frac{4 \times 3.1416 \times 475}{13{,}548}$$

$$= 0.440$$

where,

f_c is the cutoff frequency of horn,
V is the velocity in inches per second.

To further illustrate the procedure, assume the area of the flare at a point four inches from the throat end is to be calculated.

$$S_x = S_1 \epsilon^{mx}$$

where,

S_x is the area of the flare at any given point along the flare in inches,
S_1 is the initial throat area at the driver-unit end,
ϵ is 2.718, the base of Napierian Logarithms,
mx is the rate of flare at any given point along the flare.

The value of mx may be calculated:

$$mx = 0.440 \times 4$$
$$= 1.76$$

Referring to a Table of Exponential

Functions, under column (x) find 1.76. Opposite in column ϵ^x read 5.8124.

$$S_x = S_1 \epsilon^{mx}$$
$$= S_1 \times 5.8124$$
$$= 0.140 \times 5.8124$$
$$= 0.8137 \text{ inch}$$

Since the month of a horn must be one-quarter wavelength of the cutoff frequency, the horn length must extend to a point where the bell or mouth diameter is approximately 7.21 inches.

The area for each half-inch of length is computed in a similar manner. When the horn is included as a part of a group of horns as shown in Fig. 20-72A, they do not have to extend to the cutoff frequency as required for a single horn, because the several bells have a total diameter that will extend to beyond the cutoff frequency.

The flare rate for different size horns relative to their cutoff frequency is given below.

Cutoff Frequency	*Rate of Flare Doubles Every*
64	12.00 inches
128	6.00 inches
256	3.00 inches
512	1.50 inches
1024	0.75 inches

20.52 What is an air column?—The air in the throat of a horn.

20.53 What is the length of a horn for a given cutoff frequency?

Cutoff Frequency (Hz)	*Side Dimensions of Mouth (inches)*
650	5.2
512	6.6
256	13.2
128	26.4

20.54 What are the different rates of flare used with loudspeaker horns?—The Hypex, Conical, Exponential, and Parabolic. The rate of flare determines the efficiency of the horn design. The Hypex is a rate of flare developed by the Jensen Manufacturing Co.

Different rates of flare in common use are shown in Fig. 20-54. The efficiency of a horn falls between 25 and 50 percent.

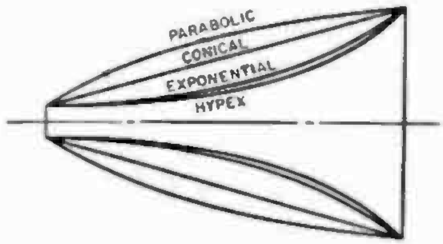


Fig. 20-54. Comparison of the different rates of flare used in the design of loudspeaker horns.

20.55 What part does the throat of a horn play in its design?—The area of the throat determines the loading on the diaphragm. If the area of the throat is small compared to the area of the diaphragm, the efficiency is increased because of the heavier loading effect. However, small throats require a longer horn, which increases the frictional losses. A horn designed to use the Jensen Hypex flare has a considerably higher throat resistance as may be seen in Fig. 20-55.

20.56 How is resonance held to a minimum in a horn-type loudspeaker?—By making the bell or mouth of the horn a dimension which is two-thirds or more of the lowest frequency to be reproduced. Horn resonance causes cancellation of certain frequencies and introduces distortion.

20.57 What is a re-entrant horn?—An exponential horn folded within itself to reduce its physical length as shown in Fig. 20-57A. The folding of the horn permits it to be designed for a lower cutoff frequency with a shorter physical

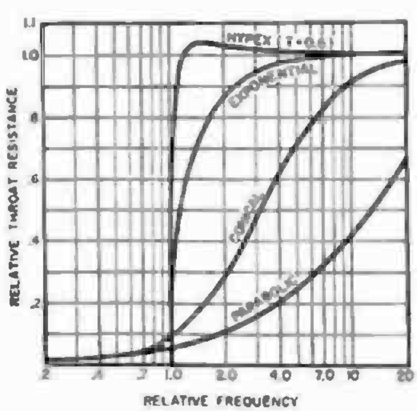


Fig. 20-55. Comparative performance of the horn flare rates that are illustrated in Fig. 20-54.

Fig. 20-57A. Jensen Hyperbolic-exponential folded horn, Model H-240.

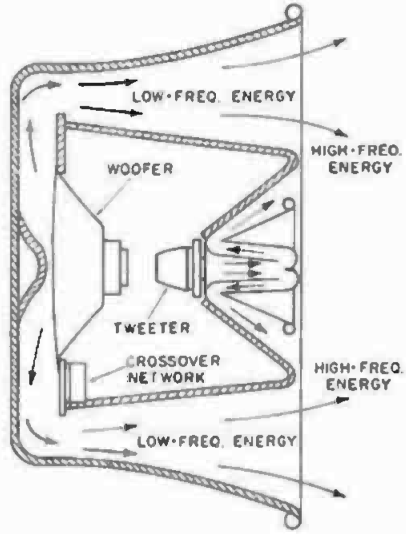


Fig. 20-57C. A two-way re-entrant type folded horn manufactured by University Sound.

length. Such horns are designed to be weatherproof and can be used either indoors or outdoors. The low-frequency cutoff is 120 Hz, with a coverage angle of 75 degrees.

Cross-sectional drawings of two more elaborate designs, by University Sound, are shown in Figs. 20-57C and D. These speakers are designed to use both low- and high-frequency units mounted inside the horn, with a simple crossover network. The arrows indicate the path of sound projection. The unit shown in Fig. 20-57C has a frequency range of 50 to 15,000 Hz, while the unit shown in Fig. 20-57D has a low-frequency cutoff of 150 Hz, with the same high-frequency response. The dispersion angle is 90 degrees for the assembly in Fig. 20-57C and 120 degrees for the one in Fig. 20-57D. The power-handling capabilities are 30 and 15 watts, respectively, and the impedance is 8 ohms.

20.58 What is a directional baffle? —A short, flared baffle placed in front

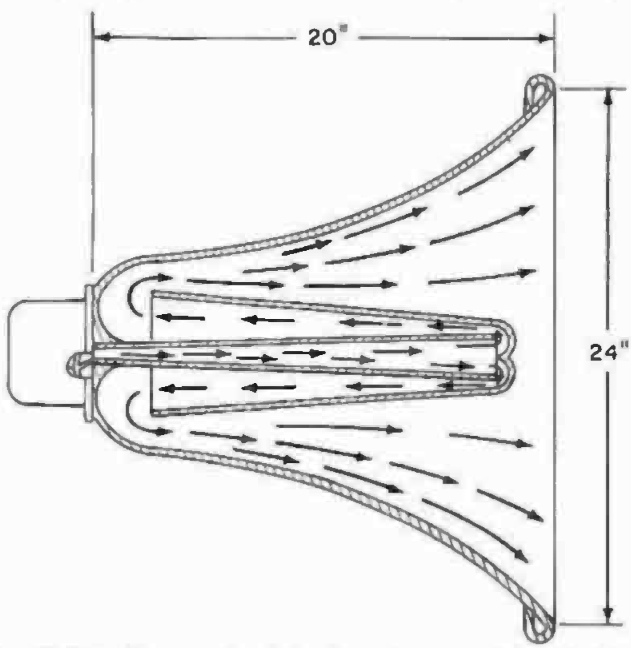


Fig. 20-57B. Cross-sectional drawing of exponential folded horn.

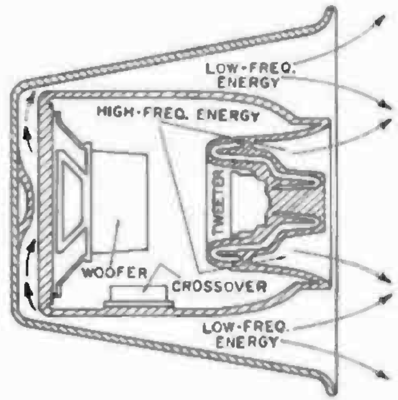


Fig. 20-57D. Another two-way re-entrant type folded horn manufactured by University Sound.

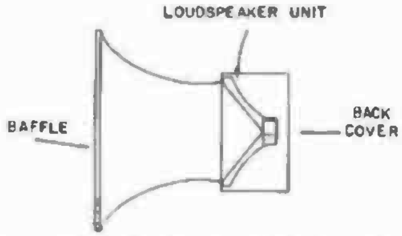


Fig. 20-58. A directional loudspeaker baffle. This baffle should not be confused with a horn. The diameter of the throat is the same as the speaker.

of a dynamic loudspeaker unit as shown in Fig. 20-58. The rear of the baffle is closed tightly and has only two small holes to relieve the pressure at the rear of the diaphragm. The baffle should not be confused with a horn. The throat of the baffle is the same diameter as the loudspeaker diaphragm.

20.59 ***What Is a curled horn?***—A horn-type loudspeaker with a rather low low-frequency cutoff and which is curled back upon itself to reduce its physical length, as shown in Fig. 20-59. These horns were used in early motion picture theater installations.

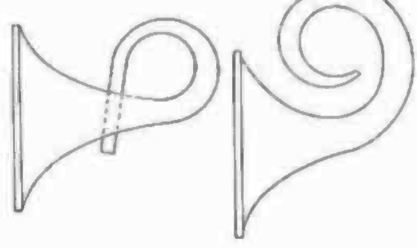


Fig. 20-59. Curled loudspeaker horns.

20.60 ***Describe the basic principles of a bass-reflex enclosure.***—The bass-reflex enclosure, as it is known today, was first introduced by Jensen in about 1937. It has become one of the most popular types of enclosures for increasing low-frequency reproduction. However, many researchers had worked on this type enclosure prior to its introduction by Jensen. Among them were Thuras of the Bell Telephone Laboratories in 1930, Voight of England, Olson of RCA, and others. They based their work on the original Helmholtz resonator, invented in the 19th century. The basic design consists of an air-tight enclosure, with a loudspeaker unit set near the center of the front panel, with a port or opening below (Fig. 20-60A).

The base-reflex enclosure is in reality a Helmholtz resonator, consisting of an enclosed volume where acoustical capacitance resonates with a mass of air enclosed within the confines of a port opening. By properly adjusting the port area or the volume of the enclosure, it is possible to tune the enclosure for a smooth, extended low-frequency response (see Question 20.29). The back wave of the speaker diaphragm is used to reinforce the frequencies below approximately 150 Hz. The acoustic impedance of the enclosure, with the help of the tuned port opening, shifts the phase of the back wave by 180 degrees, so that when it leaves the port it radiates in phase with the front wave from the speaker. Thus, the low-frequency response is increased.

The location of the port opening is not too important, since the wavelength of the frequencies at which the port is

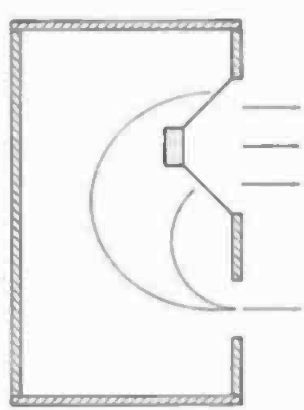


Fig. 20-60A. Cross-sectional view of a bass-reflex enclosure.

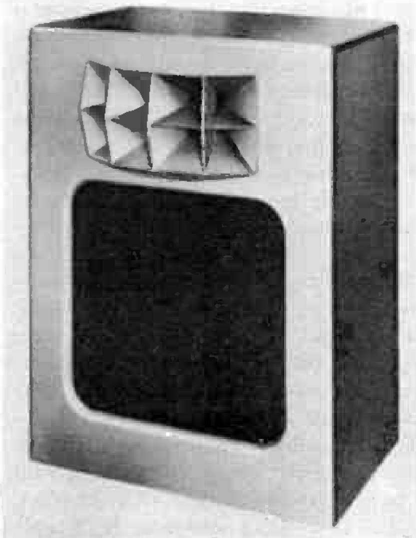


Fig. 20-60B. A bass-reflex enclosure.

effective is much longer than the overall dimensions of the enclosure. Also, the shape of the port may be varied to suit a particular design; however, the aspect ratio must not exceed 5:1. The important factor is the area of the port in square inches. The port may be circular, rectangular, or square, and may also be divided into two ports provided the total area of the openings does not exceed that for a single port. While the position of the port is not critical, it cannot be closer than 2 inches from the speaker opening. As to the overall dimensions of an enclosure, the depth must not be less than ⅓ the width, with the height and width not less than twice the diameter of the loudspeaker unit.

Since the port, in effect, closes up acoustically at the higher frequencies, the bass-reflex is used only at the low frequencies. To extend the overall frequency range, separate units must be employed for the midrange or high frequencies. A typical bass-reflex enclosure, using a 2 × 4 multicellular high-frequency horn assembly and a 15-inch low-frequency speaker unit crossing over at 800 Hz, is shown in Fig. 20-60B.

Free-air resonance for a given speaker may be determined by reference to the plot in Fig. 20-60C—cubic volume versus port opening. This is further discussed in Question 20.84.

20.61 Describe the construction of a bass-reflex enclosure.—Bass-reflex enclosure design may be approached from two different standpoints—where space is not a factor and the enclosure may be placed against a wall or in a corner and the free-air resonance of the speaker is unknown, and where the speaker unit must fit into a given space and the free-air resonance is known.

For the first method, the data given in Fig. 20-61A are used. The diameter

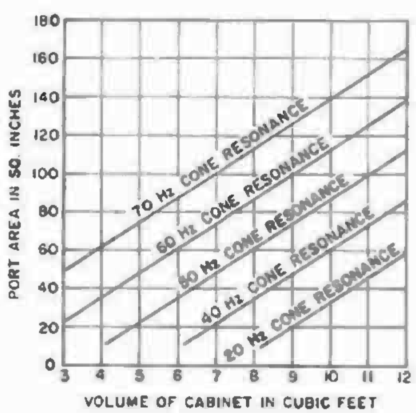


Fig. 20-60C. Relationship of enclosure volume versus port area, for a given free-air resonance of the loudspeaker diaphragm. For an enclosure above or below the limits of the above graph, infinite baffle types are recommended. (Courtesy, Altec-Lansing Corp.)

SPKR DIA	A	B	C	D	E	PORT AREA	CUBIC FOOTAGE
8"	24	16	11	7	3	1/2 AREA D	2.75 FT
10"	28	22	12	9.5	3	1/2 AREA D	4.27 FT
12"	31	24	13	11	3.5	1/2 AREA D	5.6 FT
15"	34	26	14	13.5	4	9/16 AREA D	7.17 FT
18"	40	27	14	16	4.5	5/8 AREA D	8.75 FT

AREA = πR^2 R = RADIUS OF DIAPHRAGM

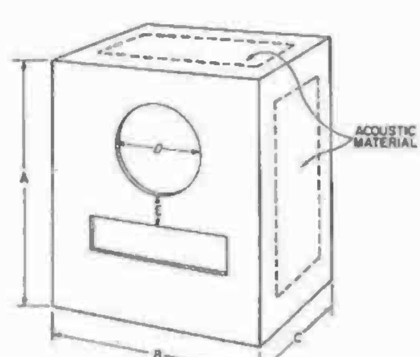


Fig. 20-61A. Dimensions for bass-reflex enclosures. The dimensions shown do not take into consideration the resonant frequency of the loudspeaker.

of the speaker unit governs the dimensions for the enclosure, which are taken from columns A, B, and C, with openings for the speaker and port taken from columns D and E. Data for the port opening and cubic footage of the enclosure are read in the last two colums. If devices such as crossover networks or horn assembly are to be included, they are placed in a separate compartment above the enclosure. The dimensions given in Fig. 20-61A do not take into consideration the free-air resonance of the speaker cone.

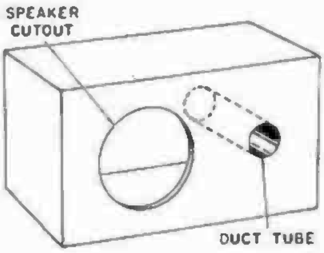


Fig. 20-61B. Bass-reflex loudspeaker enclosure using a ducted port.

The most desirable material for the construction of an enclosure is ½- to 1-inch plywood. The thickness will depend upon the size of the enclosure. Speaker enclosures are designed to be nonresonant, and not as a sounding board used in a piano. The construction must be as rigid as possible with 2 × 4-inch diagonal bracing on all interior surfaces, and corner blocks. When a panel is struck with the fist, only a dull thud should be heard. After bracing, the interior surfaces are completely lined with 2-inch *Fiberglas*, secured with tacks in combination with about 1-inch diameter fiber washers. The *Fiberglas* must not be compressed. All joints must be securely screwed down and glued to ensure an airtight enclosure. Felt or rubber stripping is employed to seal around removable panels. The hole for the passage of the speaker must be carefully sealed.

The second design approach is for a situation where the enclosure must fit into a particular space and the free-air resonance of the speaker is known. In this instance, a rectangular-shaped enclosure employing a duct-tube opening is recommended (Fig. 20-61B). The data for this type design are given in the table of Fig. 20-61C.

Because the design of the enclosure depends on a given free-air speaker resonance and the enclosure volume, the values given are net values and do not take into consideration the thickness of the absorption materials or the area taken up by additional speaker units. Using the values given, very satisfactory results can be obtained. The frequency of the speaker free-air resonance is taken from the manufacturer's data sheet. Fig. 20-61C shows that for a speaker of a given free-air resonance, the proper duct-tube diameter and length can be read under the cubic volume of the enclosure. In some instances, a rectangular' port area is given in square inches. It will be observed that at the extremes of frequency and volume range, an unvented or infinite baffle-type enclosure is recommended.

To arrive at the correct volume of an enclosure, subtract the thickness of the panel from the height, width, and depth, and multiply the remaining figures to obtain the internal volume. The shape of the enclosure is not important; therefore, it may be shaped to set on a shelf, against a wall, or for corner use. The important factors are the cubic volume of the enclosure, free-air resonance, and the size of the duct tube.

If mid- and high-frequency speaker units of the direct-radiator type are included in the enclosure, the back of these units must be boxed in with separate air-tight enclosures, using the smallest possible dimensions. This is done to avoid interference with the low-frequency unit. If in the design neither the volume nor the resonant frequency of the speaker matches the figures given, choose the closest volume or resonance values shown. If the value falls halfway between two values, move to a higher value. Inside diameters (I.D.) for the duct-tube are given below the table. There is a mistaken idea that the larger the enclosure, the better are the results. However, this is far from the truth, as there is an optimum-size enclosure for a given speaker size; therefore, the design data should be adhered to. (See Questions 20.86 and 20.89.)

After the enclosure is finished, the front panel is painted dead black to camouflage the speaker opening and is covered with an open weave grille cloth. The resonant frequency of a speaker

Free-Air Resonance	*Volume-Cubic Feet*								
	1.5	*2.0*	*2.5*	*3.0*	*3.5*	*4.0*	*5.0*	*6.0*	*8.0*
15	Closed	Closed	Closed	(T2) 11"	(T2) 8¾"	(T2) 7⅝"	(T2) 5¾"	(T2) 4½"	(T3) 8"
20	Closed	(T2) 8¾"	(T2) 6⅝"	(T2) 5¼"	(T2) 4¼"	(T2) 3⅝"	(T3) 7⅛"	(T3) 5¼"	(T3) 8¼"
25	(T2) 7¼"	(T2) 5"	(T2) 3¾"	(T2) 2¾"	(T3) 6"	(T3) 5"	(T3) 3¼"	(T5) 8¾"	(T5) 5½"
30	(T2) 4½"	(T2) 3"	(T3) 5¾"	(T3) 4½"	(T3) 3½"	(T5) 9¼"	(T5) 6½"	(T5) 4¾"	11 sq. in.
40	(T3) 5½"	(T3) 3½"	(T5) 7¾"	(T5) 5¾"	(T5) 4½"	(T5) 3¼"	13 sq. in.	18 sq. in.	28 sq. in.
45	(T3) 3¾"	(T3) 2¼"	(T5) 5½"	(T5) 3¾"	10 sq. in.	13 sq. in.	20 sq. in.	26 sq. in.	43 sq. in.
50	(T3) 2½"	(T5) 5½"	(T5) 3½"	13 sq. in.	16 sq. in.	18 sq. in.	29 sq. in.	39 sq. in.	62 sq. in.
55	(T5) 6¼"	(T5) 4¾"	11 sq. in.	15 sq. in.	20 sq. in.	25 sq. in.	37 sq. in.	61 sq. in.	88 sq. in.
60	(T5) 4¾"	11 sq. in.	16 sq. in.	20 sq. in.	29 sq. in.	35 sq. in.	50 sq. in.	75 sq. in.	Closed
65	(T5) 3¼"	13 sq. in.	20 sq. in.	27 sq. in.	36 sq. in.	45 sq. in.	69 sq. in.	Closed	Closed
70	11 sq. in.	18 sq. in.	26 sq. in.	35 sq. in.	46 sq. in.	58 sq. in.	90 sq. in.	Closed	Closed
80	17 sq. in.	28 sq. in.	41 sq. in.	60 sq. in.	80 sq. in.	96 sq. in.	Closed	Closed	Closed
90	25 sq. in.	42 sq. in.	64 sq. in.	89 sq. in.	117 sq. in.	Closed	Closed	Closed	Closed
110	51 sq. in.	87 sq. in.	132 sq. in.	Closed	Closed	Closed	Closed	Closed	Closed
115	73 sq. in.	104 sq. in.	Closed	Closed	Closed	Closed	Closed	Closed	Closed

Duct tubes T2, 2" I.D.; T3, 3" I.D.; T5, 4¾" I.D.

Fig. 20.61C. Design data for bass-reflex enclosures for a given speaker free-air resonance, and enclosure volume in cubic feet.

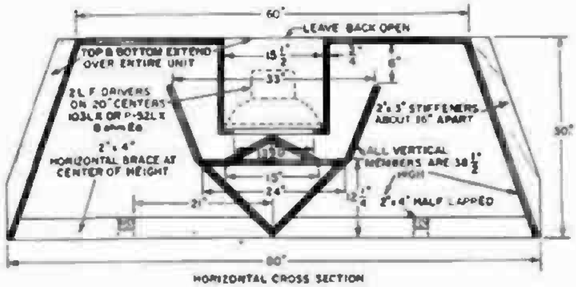


Fig. 20-63. Theater-type folded low-frequency horn (bath-tub). The vertical dimension is 38½ inches.

cone can be measured as given in Question 20.84.

20.62 ***What is a vented baffle?***—It is another name for a bass-reflex enclosure. It is also called a phase inverter.

20.63 ***Describe a folded horn.***—An enclosure designed first as a horn and then folded to reduce its physical length. A typical theater-type folded horn is shown in Fig. 20-63. A horn of the dimensions shown will reproduce frequencies down to 50 Hz, its cutoff frequency. Two 16-inch, low-frequency units are mounted at the rear. A high-frequency multicellular horn is usually mounted at the top center of the folded horn, as shown in Fig. 20-133C. The vertical opening of the horn in Fig. 20-63 is 38½ inches.

20.64 ***What is a Karlson enclosure?***—An enclosure design based on the broadbanding effect of an air column coupled to the outside air by means of an exponential slot in the front panel. A typical enclosure is shown in Fig. 20-64A, and its development from a simple pipe is shown in Fig. 20-64B.

The enclosure volume is 4.5 cubic feet, and it is claimed by the manufacturer to have a frequency range of 12 Hz to beyond audibility, with an efficiency of 40 percent. The high-frequency dispersion is 160 degrees in the horizontal plane, and 120 degrees in the vertical plane.

20.65 ***What is a Klipschorn?***—A low-frequency loudspeaker enclosure developed and patented by Paul W. Klipsch. The enclosure is a low-frequency horn so folded that it may be placed in a room corner to utilize reflections from the floor and walls to

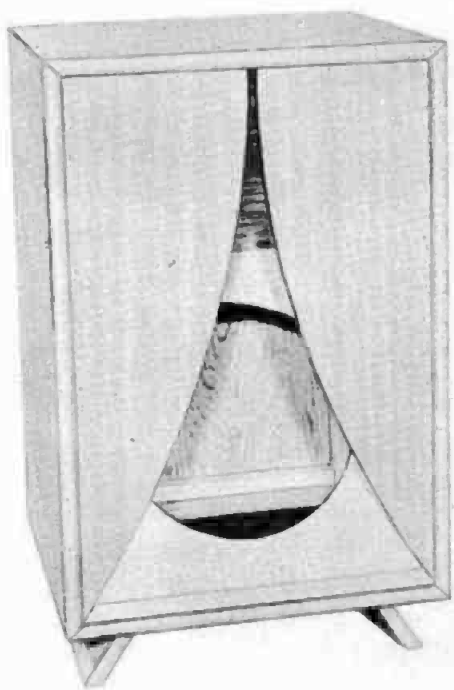


Fig. 20-64A. Karlson loudspeaker enclosure.

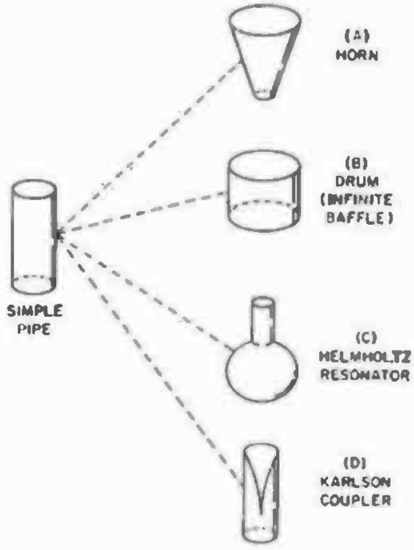


Fig. 20-64B. Origin of the Karlson loudspeaker enclosure.

improve the impedance match at the mouth of the horn and thus increase the response at the low frequencies.

One advantage of using a horn at the low frequencies compared to the use of a direct radiator mounted in a flat baffle is that the horn efficiency is 10 to 50 times greater; and because of the acoustic loading, a given acoustic power may be generated with considerably less excursion of the loudspeaker diaphragm, thus reducing harmonic and intermodulation distortion.

Fig. 20-65A shows a cutaway view of the interior of a Klipschorn enclosure. The various parts and components are: A 12-inch, low-frequency dynamic loudspeaker unit, B air chamber between the loudspeaker diaphragm and the horn throat, C air chamber behind the loudspeaker diaphragm, D and DD folded sections of the horn, E and EE side ports for letting out the acoustic energy to be reinforced by the room walls, F and FF folded-horn section, G intermediate-frequency horn, H intermediate-horn throat with its driver unit, I high-frequency loudspeaker unit, J air pocket, K back member, L front wall panel, M front wall air passage, and N side member.

Fig. 20-65B is a cross-sectional view looking downward from the top of the enclosure and showing the construction of the inner air passages, loudspeaker mounting, and the side ports.

In Fig. 20-65C is shown an interior view from the right-hand side (panel removed) and showing the air passages and loudspeaker mounting. The symbols correspond to those in Fig. 20-65A. A front view is shown in Fig. 20-65D. When this enclosure is used as a part in Fig. 20-65A, a crossover network fall-

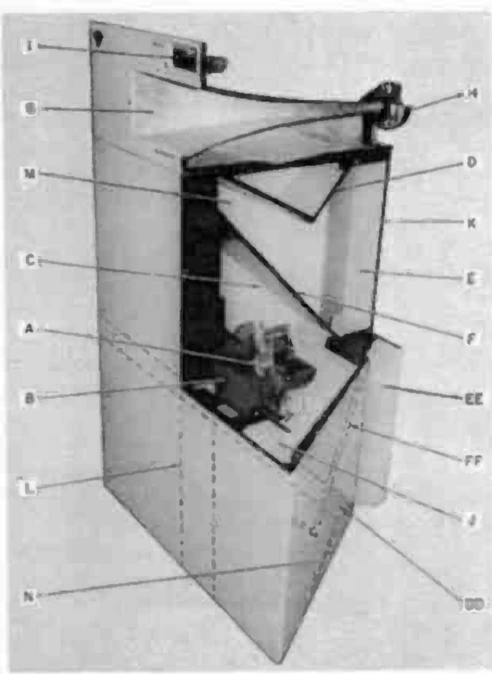


Fig. 20-65A. Cutaway view of the interior of a Klipschorn, a low-frequency horn developed by Paul Klipsch in 1940, U. S. Patent No. 2,310,243 (1943). Midrange and high-frequency loudspeakers are mounted on top of the low-frequency horn enclosure.

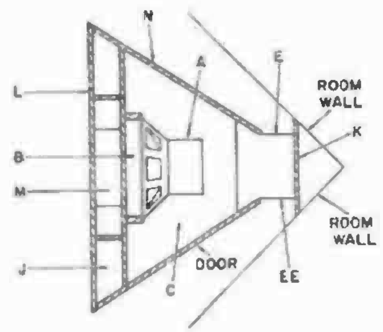


Fig. 20-65B. Cross-sectional view looking into the interior of a Klipschorn. The symbols correspond to those used in Fig. 20-65A.

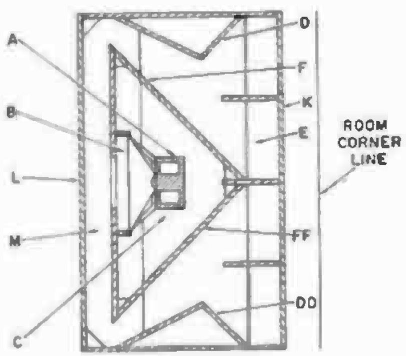


Fig. 20-65C. Cross-sectional side view of the interior of a Klipschorn (right-hand side panel removed). The symbols correspond to those of Fig. 20-65A.

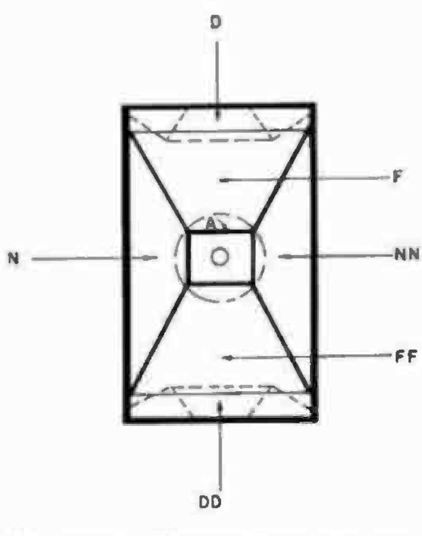


Fig. 20-65D. Cross-sectional view from the front of a Klipschorn. The symbols correspond to those of Fig. 20-65A.

of a three-way loudspeaker system as ing off at a rate of 6 dB per octave and crossing over at 400 and 5000 Hz is required to feed the intermediate- and high-frequency units.

The design of the low-frequency horn is substantially exponential in its expansion and actually comprises a series of wedge-shaped spaces which approximate an exponential rate of flare.

The frequency response covers a range of 28 to 550 Hz. Because of the short radii of bends in the horn construction, wavelengths of 18 inches, corresponding to a frequency of 750 Hz, may be reprodnced. With a 400-Hz crossover, a smooth changeover from the low-frequency horn to the intermediate loudspeaker is assured. The various sections of the low-frequency enclosure must be airtight and constructed to eliminate vibration of the sides and internal members. Otherwise, peaks in the frequency response will be generated.

The intermediate speaker G covers a frequency range from 240 to 5000 Hz. The high-frequency loudspeaker I covers from 3500 to 20,000 Hz. The overall frequency response extends from 30 to 20,000 Hz. Fundamental frequencies down to 25 Hz are radiated.

The average efficiency is over 50 percent, with frequency modulation less than 0.10 percent at 10-watts input. Maximum power-handling capability is 100 watts; impedance is 16 ohms.

20.66 ***What is an acoustic labyrinth?***—The acoustic labyrinth loudspeaker enclosure was originally developed many years ago by the Stromberg-Carlson Co. The labyrinth struc-

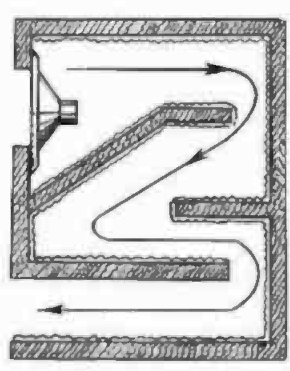


Fig. 20-66. Acoustic labyrinth loudspeaker enclosure.

ture, although appearing similar to a horn, is in reality a folded tube for loading the speaker diaphragm and can be classed as a tuned pipe. The effect is to reinforce the area above resonance, thus smoothing out the low-frequency response.

20.67 ***What is a fold-a-flex enclosure?***—An enclosure developed by Dr. Oliver Read some years ago. The enclosure is a combination of a bass reflex and folded horn. The design permits the enclosure to be operated as a folded horn, infinite baffle, or bass reflex.

20.68 ***Describe an integrated loudspeaker system.***—Such a loudspeaker system incorporates its own amplifier, and is designed to produce the best possible acoustic response from the speaker system. Matsushita of Japan accomplished this by mounting a negative-feedback coil on the speaker diaphragm and connecting it through a feedback network to the cathode of a voltage-amplifier stage. The negative-feedback voltage smooths out the frequency response, similar to the negative-feedback cutting head described in Question 14.2.

In a second method, used by EMI of England, the loudspeaker unit is connected as usual at the output of an amplifier. A negative-feedback loop having a four-section LCR network is connected from the output transformer and is fed to the input stage. Tuned circuits in the feedback loop remove the peaks and dips in the overall acoustic response. This system has also been used by Ampex and several other American speaker manufacturers.

20.69 ***Describe a stereophonic loudspeaker system using two speaker units and two sets of acoustic waveguides.***—A plan view of a stereophonic loudspeaker system, employing two speaker units and two sets of waveguides, is shown in Fig. 20-69A. This system was designed by H. F. Olson of RCA. Two

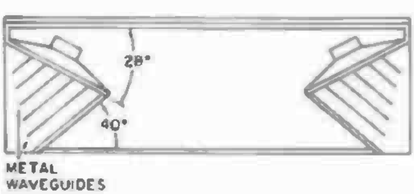


Fig. 20-69A. Plan view of stereophonic loudspeaker enclosure employing two 8-inch speaker units and a group of waveguides at each end.

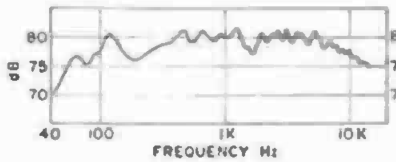


Fig. 20-69B. Frequency response of unit in Fig. 20-69A.

8-inch speaker units are mounted at an angle of 28 degrees in each end of the solid-wall enclosure with five metal waveguides. Subjective tests indicate the virtual sound source to be 3 inches beyond the ends of the enclosure. Thus, the effective separation is 36 inches between speakers. The frequency response is remarkably good, as shown in Fig. 20-69B. Dimensions for the enclosure are 30 inches long, 9½ inches high, and 10 inches deep.

20.70 ***What is a catenoid horn?***—It is a speaker system using a horn with a catenary rather than an exponential flare rate. It is claimed by its designer that when a catenary flare rate is used, the design of a folded horn is less critical. Such speaker systems employ the usual dynamic low-, intermediate-, and high-frequency speakers. Similar to other horn systems, the catenoid design also utilizes the walls to complete the horn load. (See Question 20.65.)

20.71 ***Give the basic principles of a rear-loaded folded-horn-type enclosure.***—The basic plan of construction for several different types of folded-horn enclosures, based on the work of Paul Klipsch, is given in Fig. 20-71. The enclosure, insofar as the low-frequency portion is concerned, is a folded-horn operating in a frequency range of 30 to 500 Hz. Vents are provided at the sides and rear to extend the horn by using the walls of the room as a continuation of the horn flare. The primary purpose of folding a horn is one of conserving space. Horns are antiresonant devices and should not be considered in the same category as a bass-reflex or ducted-port enclosure.

To increase the frequency range, midrange and high-frequency speaker units are added, operating in conjunction with a crossover network, which confines each speaker unit to a predetermined frequency range. Arrows on the diagrams indicate the path of the sound waves generated by the rear of

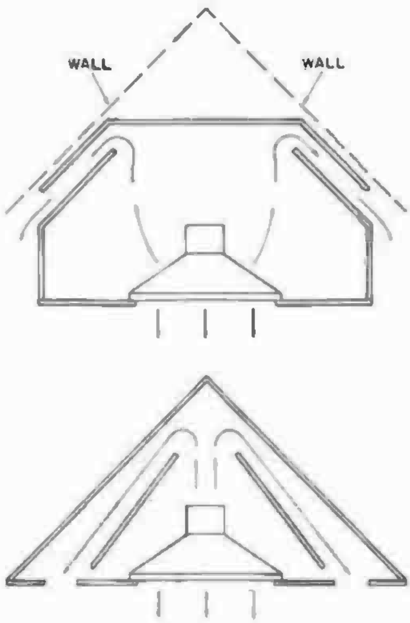


Fig. 20-71. Methods of back-loading a folded horn. The room walls are used to extend the low-frequency response.

the speaker diaphragm during its inward motion.

The advantage of a horn is its high efficiency, which is 10 to 50 times that of a flat baffle. The horn has a higher acoustic output because of the flare and acts as an impedance-coupling device between the diaphragm of the loudspeaker and the air at the mouth of the horn. This permits a given SPL to be generated with considerably less excursion of the speaker diaphragm. Consequently, there is less harmonic distortion. Thus, a horn has high power-handling capabilities with uniform frequency characteristics and low distortion.

Figs. 20-65B and C show that an air chamber is placed between the diaphragm and the throat to overcome the positive reactance at frequencies between 200 and 400 Hz, because of the multiple tapen. A second air chamber is placed behind the diaphragm to offset the mass reactance of the throat impedance at the low frequencies. (See Question 20.51.)

20.72 ***Describe a multicellular horn.*** —A multicellular horn (Fig. 20-72A) is one of the most efficient sound projectors for delivering sound over a defined listening area. Its structure consists of a number of individual exponential horns assembled in different configurations, to control the vertical and horizontal dispersion. Horns of this design have several advantages over the folded or re-entrant-type horns, as there are no bends to attenuate the high frequencies. Another advantage is that the high frequencies so important to high intelligibility of speech can be distributed over a wide angle. The beam width above the cutoff frequency and in the midhigh range to 12,000 Hz and above, is independent of frequency. This entire portion of the frequency spectrum is quite uniformly distributed throughout the full angle of the horn. Up to four driver units can be employed with a single horn assembly by using a multiple throat. Care must be taken in this type installation that the driver units are in-phase acoustically; that is, the diaphragms must all move in the same direction at a given instant. Polar patterns for a single multicellular horn are shown in Fig. 20-72B.

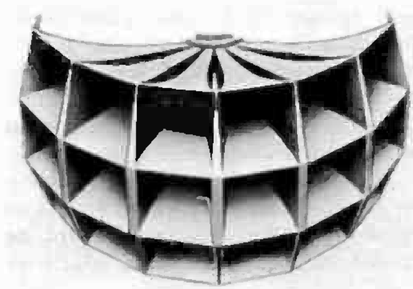


Fig. 20-72A. Multicellular horn cluster Model 1803B manufactured by Altec-Lansing.

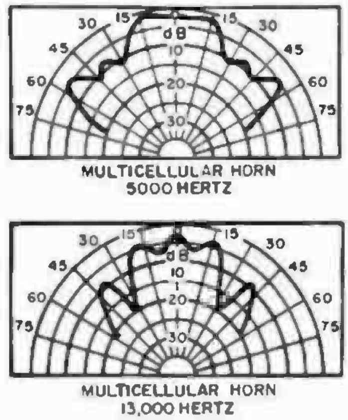


Fig. 20-72B. Polar pattern of a multicellular high-frequency horn, at frequencies of 5000 and 13,000 Hz.

These horns were developed by the Bell Telephone Laboratories many years ago and are used for sound reinforcement and for public address systems. Many motion picture theater installations use a multicellular horn for the high-frequency unit, as pictured in Fig. 20-133A.

Although the cluster in Fig. 20-72A is an 18-horn unit, a lesser number of horns can be employed. The number depends on the dispersement angles to be covered. Horn assemblies are referred to as a 2 × 3 cluster for six horns, 3 × 6 for an 18-horn cluster, etc. Clusters have been built employing up to 60 horns in a 6 × 10 array.

For installations involving only speech, to reduce reverberation the amplifier system should have a high-pass filter installed, with a cutoff frequency of 300 to 500 Hz, since the horn has an effective length greater than the physical length. (See Question 20.50.)

20.73 ***What is a multicellular baffle?*** — A single high-frequency horn with several small vanes in the bell to deflect the high frequencies over a wide area (about 105 degrees), thus reducing the directional effect of the high frequencies. (See Fig. 20-73.) It is used with multiple-loudspeaker systems.

20.74 ***What is an acoustic lens?***— A metal or plastic device placed in front of a high-frequency horn to spread the high frequencies, which are quite directional, over a wide angle of distribution.

The lens is concave in shape and pierced with a large number of small holes. This lens is transparent to sound waves because the boundary layers of the lens form a medium of increased density through which the sound passes from a high-frequency exponential horn to the listening area.

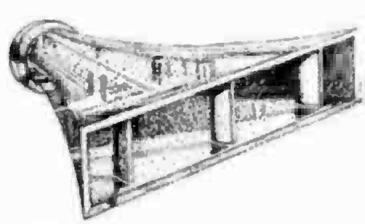


Fig. 20-73. Multicellular baffle short horn with deflecting fins.

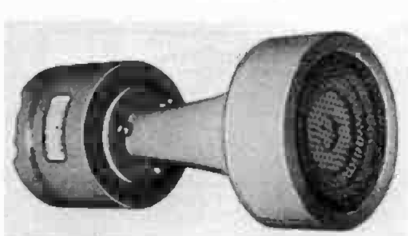


Fig. 20-74A. High-frequency acoustical lens assembly.

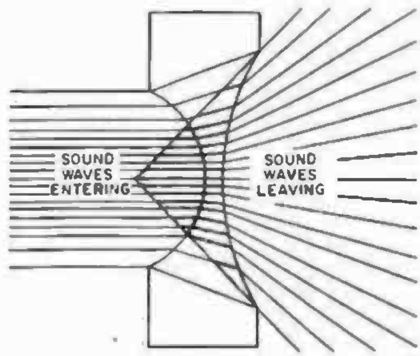
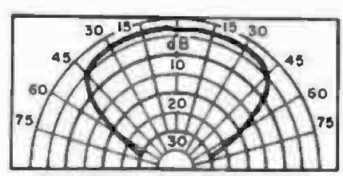
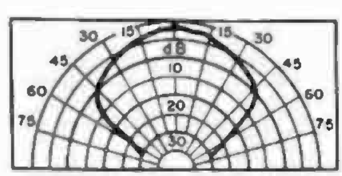


Fig. 20-74B. Ray tracing of an acoustical lens, showing the focal point and diffusion angle as controlled by the curvature and index of refraction.



ACOUSTIC LENS, 5000 HERTZ



ACOUSTIC LENS, 13,000 HERTZ

Fig. 20-74C. Field pattern for acoustical lens system.

In Fig. 20-74A is shown a high-frequency acoustical lens assembly consisting of 14 separate elements arranged in such a manner that they form a double-concave lens system deflecting the sound over an angle of 90 degrees. The index of refraction for this lens system is 1.3. A cross-sectional view of the focal point and diffusion angle controlled by the curvature and index of refraction may be seen at Fig. 20-74B. The field pattern for this lens, taken at

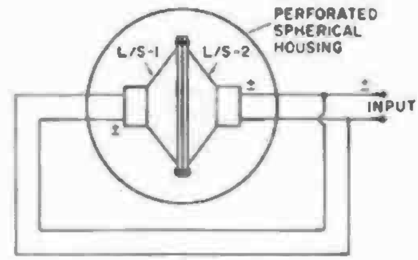


Fig. 20-76. Simulated spherical loudspeaker. The housing is perforated with many small holes. The voice coils are connected 180 degrees out of phase.

5000 and 13,000 Hz, appears in Fig. 20-74C.

20.75 What are the characteristics of an open baffle?—Open-back baffles are not used with modern high-fidelity loudspeaker systems; however, many times small extension speakers are mounted in such baffles. Actually, an open-back baffle is a flat baffle turned back on itself to decrease the size of the box. If the back is closed, the free-air resonant frequency of the speaker diaphragm is raised and the cabinet resonant effects are noted. If the baffle is made deep enough, it functions as a pipe and several points of resonance, which increase as the cabinet is deepened, will be generated. Closing the rear of the box prevents any use of the sound wave from the rear, with a resultant loss of acoustic energy. The most logical solution is to port the box; thus, it becomes a bass reflex based on the Helmholtz resonator. (See Question 20.60.)

20.76 Describe a spherical or omnidirectional loudspeaker.—To design a completely omnidirectional or spherical loudspeaker is rather difficult, as the energy from the diaphragm must cover 360 degrees vertically and horizontally. A simulated spherical loudspeaker can be designed by placing two identical speaker units face to face and fastening them securely. The voice coils are connected in parallel, but 180 degrees out of phase, as shown in Fig. 20-76. The two units are mounted in a spherical housing, perforated with many small holes. When hung from a ceiling, it functions quite well as a spherical radiator. (See Question 20.77.)

20.77 What effect does the exterior shape of an enclosure have on the overall frequency response of a loudspeaker?—H. F. Olson of the RCA Princeton Laboratories, in his work with enclosures, reached the conclusion that a direct-radiator-type loudspeaker responds according to the various shaped enclosures and that the outside configuration plays an important part in the final frequency response of the speaker system. It is possible for some shapes to have a 10-dB variation in response, because of diffraction.

Detrimental effects of diffraction may be reduced by eliminating sharp boundaries on the front of the enclosure in order that the diffracted wave will be reduced in amplitude. It will also help if the distance from the speaker diaphragm to the diffracting edges is varied to produce random phase relationships between the primary source and the diffracted sound waves.

A group of 12 differently shaped enclosures, measured by Olson, are given in Fig. 20-77A. Their frequency characteristics are shown in Fig. 20-77B. It appears from Olson's work that only three shapes least affect the final reproduction. They are the sphere A and truncated pyramids mounted on rec-

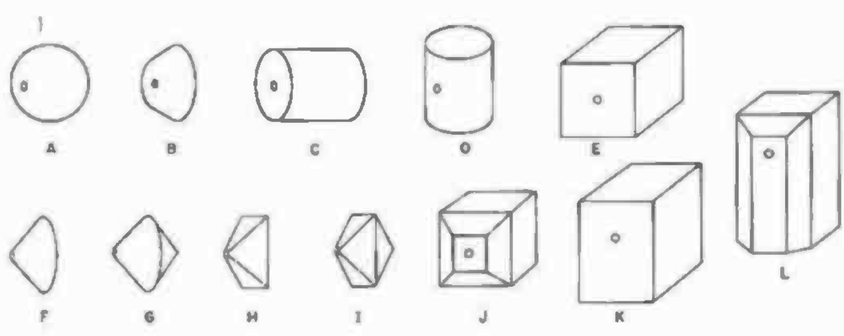


Fig. 20-77A. Twelve different enclosures tested for their effect on the final frequency response of a loudspeaker, by H. F. Olson of the RCA Princeton Laboratories.

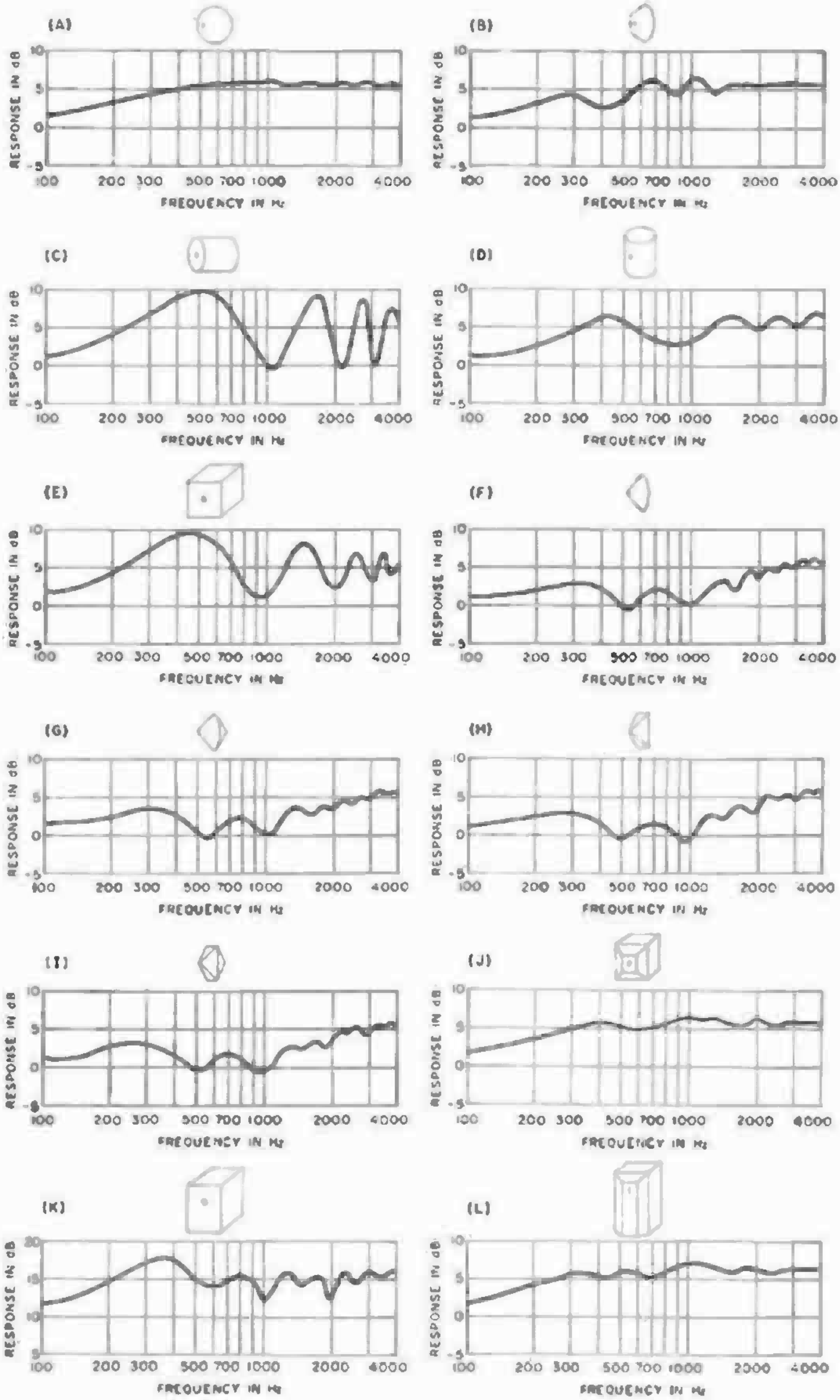


Fig. 20-77B. Frequency-response characteristics using a small direct-radiator type loudspeaker in the enclosures shown in Fig. 20-77A.

tangular enclosures J and L. The ideal shape for an enclosure appears to be the sphere.

20.78 ***Describe a solid-state high-frequency speaker unit.***—A solid-state high-frequency loudspeaker unit, utilizing the piezoelectric characteristics of a ceramic rod, coupled to an aluminum diaphragm is shown in the cross-sectional drawing of Fig. 20-78A. By ap-

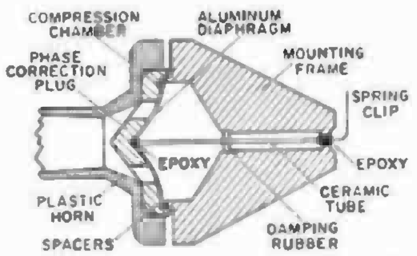


Fig. 20-78A. Cross-sectional view of a solid-state high-frequency loudspeaker unit. (Courtesy, Motorola Semiconductor Products Inc.)

plying the audio signal directly to a ceramic tube, the diaphragm is driven without benefit of either a voice coil or magnetic field. The driving member, 2.5 inches in length and consisting of lead zirconate and lead titanate, is a tube which expands and contracts longitudinally, depending on the instantaneous polarity of the applied signal. The tube has an outside diameter of 0.050 inch with a wall thickness of 0.010 inch, and is nickel-plated inside and out and poled through its thickness. The maximum expansion is on the order of 150 microinches. One end of the rod is epoxy cemented to a die-cast frame with the free end attached to the diaphragm. The compression chamber between the phasing plug and the diaphragm is 0.010 inch.

The power-handling capability is 10 volts per 0.001 inch. The SPL taken on axis 18 inches from the horn throat varies from 99 dB to 116 dB at 11,000 Hz. The steep front of the frequency characteristic (Fig. 20-78B) caused by the falling impedance characteristic, serves as a built-in crossover network. The distortion characteristics are comparable to those of an equivalent dynamic unit, except for a slight rise at the extreme ends of the frequency range. This unit was developed by Hugo Schafft of Motorola Semiconductor Products Inc.

20.79 ***What type grille cloth should be used for wide-range loudspeaker system enclosures?***—A loosely woven cloth or plastic with hard thread. A closely woven cloth will reduce the high-frequency response by several dB.

20.80 ***How is the volume of a loudspeaker enclosure calculated?***—

$$\text{Volume} = (H \times W \times D) - n = V_{cu\,in}$$

$$\left(\frac{H \times W \times D}{1728}\right) - n = V_{cu\,ft}$$

where,

H is the height in inches,
W is the width in inches,
D is the depth in inches,
n is any device mounted within the enclosure.

20.81 ***How may the resonant frequency of a loudspeaker enclosure be damped?***—By completely lining the interior surfaces of the enclosure with a highly absorbent material such as rock wool. The resonant frequency of the panels may be damped by the use of diagonal braces and by filling unused spaces with sand.

20.82 ***How may a loudspeaker enclosure be tested for undamped resonant frequencies?***—By applying white noise to the loudspeaker system, then picking up the acoustic output by means of a capacitor microphone, and finally by observing the signal on an oscilloscope. The sweep frequency of the white-noise generator is varied to indicate the resonant frequencies of the enclosure. (See Questions 1.140, 22.56, and 23.135.)

20.83 ***How may the low-frequency response in a bass-reflex enclosure be smoothed out?***—The port of a bass-reflex enclosure is considered to be properly tuned for the best low-frequency response when the bass response has been equalized and spread over as wide

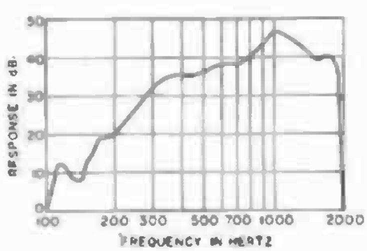


Fig. 20-78B. Frequency response for solid-state high-frequency loudspeaker unit of Fig. 20-78A.

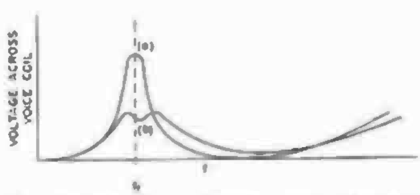


Fig. 20-83. Generalized curves showing how the resonant frequency of a loudspeaker in a bass-reflex enclosure may be smoothed out by adjustment of the port area. (a) Before adjustment. (b) After adjustment.

a range as possible. Excessive boominess in the low frequencies is generally an indication of an improperly dimensioned port area. If any doubt exists as to the correct port dimensions, an audio oscillator is connected to the speaker system, and an ac voltmeter is connected across the voice-coil leads.

A wooden shutter is placed partially over the port area, and the oscillator is varied over a frequency range of 50 to 200 Hz. The shutter is adjusted for a meter reading with a minimum of peaks. It will not be possible to secure a smooth voltage across the voice coil at all frequencies because of the variation in the impedance of the voice coil with frequency. However, with the proper adjustment of the shutter, the port may be tuned to a point where the voltage peaks are more or less uniform. The largest peak will be found at the point where the voice coil rises to its maximum impedance. The effect of tuning the port is shown in Fig. 20-83.

20.84 ***How is the free-air resonant frequency of a loudspeaker unit measured?***—The speaker is hung in free air, away from reflecting surfaces, and is connected to an amplifier as shown in Fig. 20-84. The oscillator is swept from

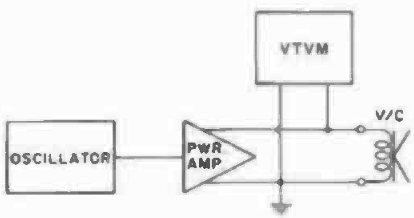


Fig. 20-84. Circuit for measuring the free-air resonant frequency of a loudspeaker unit.

below 20 Hz to 200 Hz. At the resonant frequency, the excursion of the diaphragm will be maximum as read on the meter. When measuring small speakers, care must be taken that the diaphragm is not damaged by driving it beyond its normal excursion.

20.85 ***Describe the principles of a diffraction horn.***—Diffraction horns are narrow-mouth horns. The flare provides the proper cutoff frequency, while the narrow mouth provides a slit source operating in a vertical plane. The device works on the principle that sound leaves a narrow slit that is small in

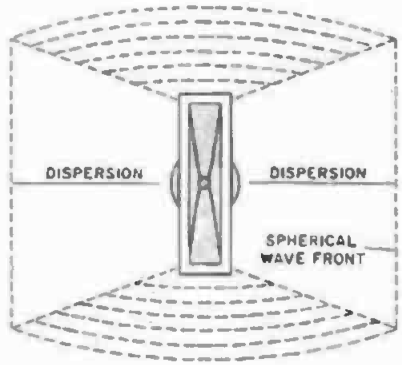


Fig. 20-85A. Diffraction horn Model TW-35 manufactured by Electro-Voice, Inc. Wavelengths that are large compared to the dimensions of the slit, emerge as a point source of sound and diffract into a cylindrical waveform.

comparison to the wavelength of the emitted sound and acts as a point-source of sound. Consequently, the sound will emerge, flowing around the sides of the horn, into a cylindrical waveform (Fig. 20-85A). The term diffraction stems from the fact that the waveform diffracts out of the narrow opening. Since the mouth dimensions are small in the horizontal plane, the flare must increase quite rapidly in the vertical direction to provide the proper cutoff frequency.

The fact that the horn flares in the vertical plane does not mean that the

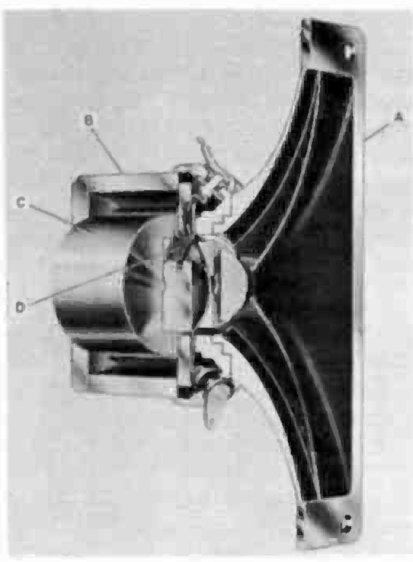


Fig. 20-85B. Diffraction horn Model TW-35 manufactured by Electro-Voice, Inc.

dispersion is greatest in the vertical plane, which varies with frequency. Because of the relationship of the slit width to wavelength, as the frequency is increased, less diffraction takes place because of the opening up of the slit. Generally speaking, such horns have a smooth angular response.

Fig. 20-85B is a cutaway view showing the interior of a diffraction horn manufactured by Electro-Voice Inc. At A is the bell, at B and C, the magnetic structure, and at D, the diaphragm and voice coil. This type speaker unit is used with multiple-speaker systems.

20.86 What is the average free-air resonance of dynamic loudspeakers?—

Diameter	*Frequency*
8 inch	60 to 150 Hz
12 inch	30 to 85 Hz
15 inch	25 to 55 Hz
18 inch	20 Hz
30 inch	15 Hz

The free-air resonance can be measured by using the method described in Question 20-84.

20.87 What is the blocked impedance of a loudspeaker?—The impedance of the voice coil when the voice coil is blocked to prevent its movement.

20.88 What is the motional impedance of a loudspeaker? — Because a loudspeaker generates a counter emf due to the movement of the voice coil in the magnetic field, the generated emf opposes the signal voltage applied to the voice coil.

The impedance measured while the voice coil is in motion is different from that when the voice coil is blocked.

Motional impedance is the impedance measured when the voice coil is in motion, minus the blocked impedance of the voice coil.

20.89 Give a typical impedance characteristic for a dynamic loudspeaker unit.—Fig. 20-89 shows a plot of a typical impedance curve for a dynamic speaker unit, taken in free air. It will be observed that there is quite a wide variation in impedance between zero and 10,000 Hz. It is not uncommon for the impedance at the resonant frequency to measure five times or more than the rated impedance, then drop off, then again rise to that amount or more at the high frequencies. The impedance is a complex value of dc resistance, inductive reactance of the voice coil, and frequency.

In Fig. 20-89, A is the dc resistance of the voice-coil winding, B the free-air resonant impedance, C the rated impedance, and D the rising impedance caused by the inductive reactance of the voice-coil winding. The EIA Standard specifies that the rated impedance is to be taken at the minimum value at C, following the resonant peak B. This impedance is sometimes referred to as trough-impedance. For modern speakers, this falls around 400 Hz and is quite narrow.

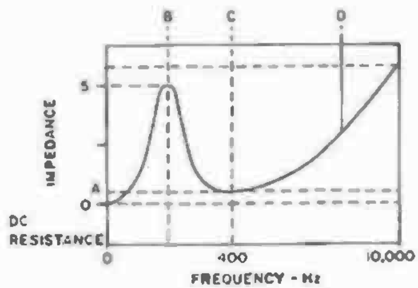


Fig. 20-89. Impedance curve for a typical dynamic loudspeaker.

As the impedance rises and approaches the resonant frequency, the counter emf (generated by the motion of the voice coil in the magnetic field) rises to a maximum at resonance, then drops off for a small portion of the frequency spectrum, and again rises as the inductive reactance of the voice coil increases. This clearly indicates that the rated impedance of a loudspeaker is only in a very narrow band. Free-air impedance can be measured as discussed in Question 23.140.

20.90 What effect does the loudspeaker impedance have on the output stage of an amplifier?—Because a loudspeaker does not present a constant reflected load to the output tubes of an amplifier, it is customary to use tubes having a very low plate resistance or beam-power tubes with large amounts of negative feedback. Variations in the loudspeaker impedance cause a nonuniform frequency response and a considerable change in distortion characteristics of the amplifier from those measured under resistive load conditions.

20.91 What effect does loudspeaker impedance have on the output stage of

a transistor amplifier?—Unlike the vacuum-tube amplifier, transistors cannot be operated at the peak of their power load curve because of the internal heat generated in the transistor. Therefore, for satisfactory operation of solid-state amplifiers, attention must be given to the magnitude and character of the load impedance.

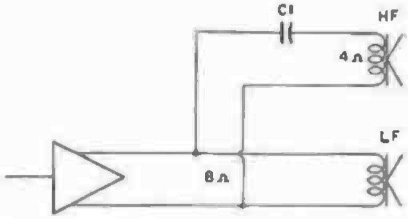


Fig. 20-91A. Simple crossover network with a capacitor in series with the high-frequency unit.

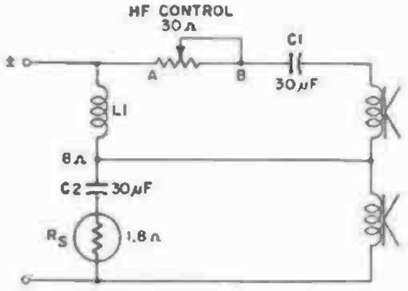


Fig. 20-91B. Simple series network with a high-frequency control and a resistor in the lower section to prevent the impedance falling below a given value.

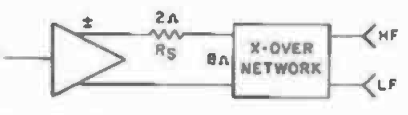


Fig. 20-91C. Series resistor in the output of a solid-state amplifier to protect the output stage.

The simple crossover network (Fig. 20-91A) employing a low- and high-frequency speaker, with the smaller speaker fed through capacitor C1 presents the worst condition. Here, when only the low-frequency speaker is operating, the impedance is near the rated impedance. However, when operating with the high-frequency unit, its impedance is in parallel with the low-frequency speaker impedance which presents a lower than rated load impedance. In effect, the minimum impedance seen by the output stage is the dc resistance of two voice coils in parallel, assuming that the frequency is high enough to make the reactance of series capacitor C1 negligible. The impedance of the high-frequency section will vary with frequency, as the reactance of the series capacitance changes with frequency.

When the voice-coil reactance for either speaker rises (see Question 20.89) and the low-frequency unit is operating near its resonant frequency, the total impedance presented to the amplifier output is higher than the rated impedance. Thus, it can be seen that the amplifier load impedance varies over a wide range. In a vacuum-tube amplifier, this will not damage the output stage, but this condition could cause considerable damage to the output stage of a transistor amplifier; therefore, protective circuits are included for its protection.

Designers of multiple-loudspeaker systems have given considerable thought to this problem, and are using protective methods in the crossover networks to avoid such a condition where a low value of speaker impedance might cause damage to the output stage.

A simple series-connected crossover network is shown in Fig. 20-91B, where a 30-ohm control is connected in the high-frequency section. If the control is at minimum (out of the circuit) position A, the impedance of the network is within the rated load impedance of 8 ohms. Turning the control to position B inserts the full 30-ohms in series with the high-frequency unit. Now the circuit consists only of choke L1, capacitor C2, and the impedance of the low-frequency speaker voice coil, which could be less than 5 ohms. A remedy suggested by one manufacturer of loudspeakers for an 8-ohm speaker network, is to insert a resistor R_s of 1.8 ohms in series with capacitor C2. An alternate method is to insert a resistor of 2 ohms in series with the amplifier output and loudspeaker system, as in Fig. 20-91C. The insertion of the resistor R_s in the network will not affect the characteristics of the low-frequency speaker, but will prevent the impedance from falling below a value that could cause damage to the output stage. (See Question 20.103.)

20.92 ***Why does the voice-coil impedance of a dynamic loudspeaker unit generally fall between 4 and 16 ohms?***

—Two factors are involved, mechanical and electrical. Mechanically, the voice coil must be suspended in a very small gap; therefore, its mass must be low. Electrically, the wire must be sufficiently heavy to carry a large current, with as few turns as possible to keep down inductive reactance. The number of turns must be such that the voice coil will react to the magnetic field sufficiently enough to drive the diaphragm. Therefore, voice coils are wound to have an impedance between 4 and 16 ohms, using either copper or aluminum wire. Voice-coil windings will vary from a few tenths of an inch in diameter for a high-frequency unit, to over 3 inches for a low-frequency speaker.

In certain early designs, the voice coil consisted of a single turn brass strip, having a few hundredths of an ohm resistance. Low-frequency speaker units designed for theater sound systems are generally 32-ohms impedance and are connected in parallel; thus, the total impedance for the two speakers is 16 ohms. (See Question 20.133.)

As a rule, speaker units designed for the high-fidelity reproduction fall in the 4- to 16-ohm category. However, impedance values are obtainable at 3.2, 10, 20, 45, and 100 ohms.

20.93 ***What is a bass energizer?***—It is a passive network inserted between the output of a power amplifier and a speaker system, to increase the low-frequency response below 150 Hz. It was developed by Alexis Badmaieff of Altec-Lansing. He used, as a basis of design, the Fletcher-Munson 80-dB curve of Fig. 1-76A. The network is designed to approximate this curve from 150 Hz to 50 Hz, then to flatten out to 20 Hz, as shown in Fig. 20-93.

Basically, the device is a T configuration and therefore has a 6-dB insertion loss. Such devices should be used only

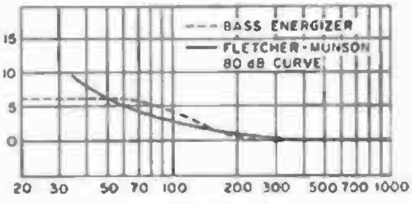


Fig. 20-93. Altec-Lansing bass-energizer passive network as compared to an 80-dB Fletcher-Munson curve.

with high-efficiency speakers or with an amplifier that can supply the necessary power output without overloading at the lower frequencies. For a 16-ohm speaker, the network has a 6-dB insertion loss at 400 Hz; for an 8-ohm speaker, it has a 10-dB loss. Such networks can be described by using the data given in Question 6.49.

20.94 ***How can the resonant frequency of a dynamic speaker be lowered?***—If the resonant frequency is too high, it can be lowered by using a method devised by D. E. L. Shorter. In this method, the edge of the diaphragm material is impregnated with a hygroscopic agent which will retain moisture and thus maintain the compliancy of the cone. A convenient wetting agent is Eastman Kodak Photo Flow diluted in a ratio of one capful to eight ounces of water. While the diaphragm is wet, apply Kodak Print Flattening solution (without dilution) over the Photo Flow. The solutions are applied only to the edge or hinge of the diaphragm, and are permitted to stand at least 24 hours before putting the speaker to use. Before applying the solutions, the resonant frequency should be measured for comparison after the solutions have been applied. It will be observed that the resonant frequency will be decreased considerably. Additional applications will tend to further decrease the resonant frequency until a point is reached where no further reduction is noted with additional applications.

20.95 ***What is variable damping?***—It is a method introduced some years ago for matching the output impedance of a vacuum-tube amplifier to the impedance of a loudspeaker. This is accomplished by the use of negative voltage and current feedback control in the amplifier. The method is now obsolete. This subject is further discussed in Question 12.242, and Question 23.139.

20.96 ***What are the intermodulation characteristics for a typical high-quality dynamic loudspeaker?***—In Fig. 20-96 is shown a typical intermodulation-distortion curve for a 12-inch direct-radiator type loudspeaker. The test was made by applying frequencies of 60 and 2000 Hz in a ratio of 4:1 to the voice coil, with the lower frequency 12 dB lower in amplitude than the higher frequency. Intermodulation distortion varies considerably with loudspeaker de-

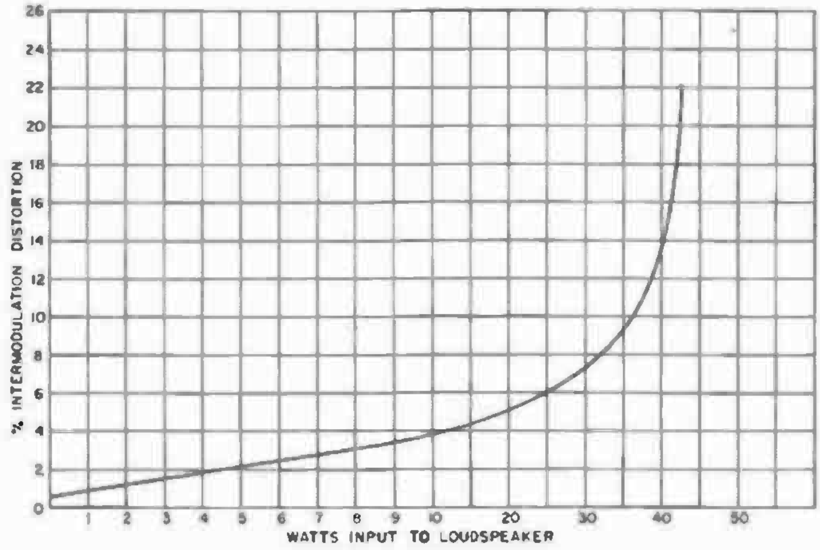


Fig. 20-96. Intermodulation distortion characteristics of an average high-quality dynamic loudspeaker.

sign, enclosures, and frequency range. No set figures can be given; however, it goes without saying that the intermodulation must be held to a minimum. Loudspeakers are available with less than one-half of 1-percent intermodulation distortion, while others may run up to several percent. This is particularly true for frequencies below 200 Hz. To date, no standards using the intermodulation method of measuring distortion have been adopted for testing loudspeakers.

20.97 ***How does intermodulation distortion affect the listening qualities of a loudspeaker?***—Frequencies which are not harmonically related to the original frequencies are generated, resulting in a distorted reproduction appearing as a fuzziness of the higher frequencies and causing listener fatigue.

20.98 ***What are the harmonic-distortion characteristics of a high-quality, wide-range, direct-radiator-type loudspeaker?***—In Fig. 20-98 is shown a distortion measurement made on a high-

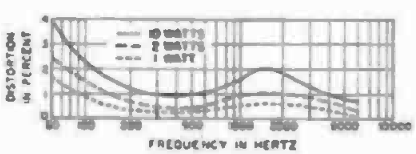


Fig. 20-98. Harmonic distortion of a typical high-quality wide-range dynamic loudspeaker.

quality loudspeaker of the direct-radiator type, for powers of 1, 2, and 10 watts at the voice-coil terminals. The frequency range covered was 60 to 6000 Hz and includes harmonics up to 18,000 Hz.

20.99 ***What is the cause of intermodulation distortion in a loudspeaker?***—Intermodulation distortion is generated in single diaphragm radiators when the diaphragm is reproducing a high and a low frequency simultaneously. If the low frequency is distorted in any manner, it distorts the high frequency by flattening off the peaks. Thus, the high frequency is modulated by the lower frequencies.

The theory of intermodulation distortion is covered in Questions 23.117 to 23.127. The distorted waveform in a loudspeaker appears quite similar to that for an amplifier and is also referred to as frequency-modulation distortion.

Multiple-loudspeaker systems generally have much lower intermodulation distortion than does a single diaphragm speaker, as each speaker unit is separate and cannot be modulated by others. Speaker systems using a low-frequency horn, midrange-, and high-frequency units have very low intermodulation distortion.

Overdriving the voice coil can cause it to leave the magnetic-gap area, thus generating sum and difference frequencies, which is another form of intermodulation distortion.

In horn-type projectors, intermodulation and harmonic distortion are caused by nonlinear compression of the air in the throat of the horn. At 3000 Hz, a horn with a cutoff frequency of 150 Hz using a ⅞-inch throat can generate up to 17 percent second harmonic distortion, with a power of 1 acoustic watt from the driver unit. Since the distortion is proportional to the number of wavelengths passing through the throat, it increases with an increase of frequency. In compound-type horns, distortion is reduced because of the shorter throat length at the higher frequencies.

20.100 How can the intermodulation distortion be reduced in a wide-range loudspeaker system?—By using individual low-, intermediate-, and high-frequency speaker units fed from a suitably designed crossover network to separate the frequency response of the speakers into bands. The several units are mounted in a common enclosure. If a single unit is to be employed, it may consist of a duplex, coaxial, or a three-way unit mounted in an enclosure. Units shown in Figs. 20-5, 20-6, and 20-25 are typical examples of speaker units having low distortion. However, the distortion for a single unit is not as low as that which can be obtained by a properly designed multiple-speaker system.

20.101 What is a tone burst?—A method used for testing transient response of a loudspeaker or system. Loudspeakers are electroacoustic transducers and achieve their purpose by converting electrical pulses into mechanical motion and then into acoustical pressure. It is in the mechanical moving system where much of the transient distortion is generated.

Whenever a signal is applied to a mechanical system at rest, there is a time delay before the system responds to the stimulus. Furthermore, when the signal is removed there is a short time delay before the system returns to rest. Tests indicate that the loudspeaker with the smoothest frequency response will have the smallest starting transient and the least hangover.

Transient measurements are not too difficult to make and can be accomplished by the connection and use of the equipment given in the block diagram of Fig. 20-101. Here is shown an audio oscillator feeding a tone-burst generator, the output of which is applied to a power amplifier of low distortion which feeds the loudspeaker under test in an anechoic chamber. The signal from the speaker is picked up with a microphone of known frequency characteristics and fed to a dual-trace oscilloscope. The trigger output from the tone-burst generator is connected to the trigger circuit of the oscilloscope. The grounding should be carried to a common point. It is not absolutely necessary that the measurement be made in an anechoic chamber. If a room that is fairly well damped is available and the microphone is placed about 2 feet in front of the speaker, a satisfactory measurement can be made. The procedure is to first make a frequency-response measurement and then to choose a point in the midrange response where it is smoothest and away from peaks or dips. The results are then displayed on the oscilloscope or plotted on an automatic recording device.

In using a tone-burst generator, the sine-wave oscillator is set to the desired frequency and keyed by the tone-burst generator, which lets a given number of hertz of a preselected frequency through the power amplifier in short bursts and selected intervals. Thus, a

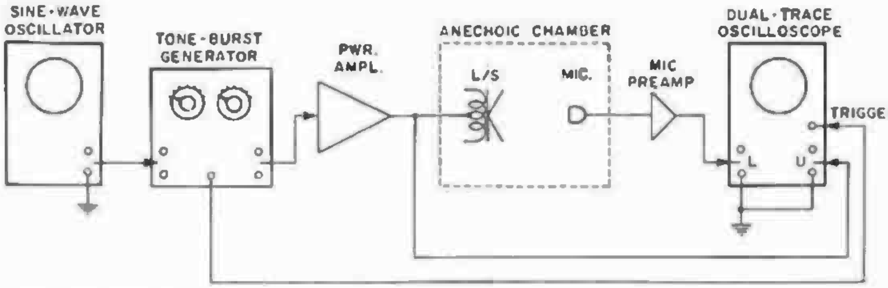


Fig. 20-101. Block diagram for making loudspeaker transient measurements using a tone-burst generator, and a dual-trace oscilloscope.

1000-Hz signal from the audio oscillator might be keyed at a rate of 2, 4, 8, 16, 32, 64, or 128 times per second.

20.102 How can a loudspeaker system be tested for transient distortion without using test equipment?—A rough test can be made with the use of an fm receiver. The receiver is tuned to a spot removed from a station carrier frequency, so that a strong hissing sound is heard. This noise is similar to white noise and covers the entire frequency spectrum. If the speaker system under test has pronounced peaks and valleys, the high frequencies appear as singing tones, and the low frequencies manifest themselves as torchlike sounds, similar to those made by a blow torch.

20.103 What effect does amplifier damping have on a loudspeaker?—The subject of amplifier damping and its effect on the performance of a loudspeaker is a controversial one, like pentodes versus triodes. It has been the practice to state the damping factor of an amplifier in terms of the ratio of the internal output impedance to the load impedance. Thus, for a loudspeaker of 8-ohms impedance and an amplifier with an internal output impedance of 0.5 ohm, the damping factor is 16. Amplifiers have appeared, stated to have a damping factor in excess of 100. Actually this is not true, as the dc resistance of the voice coil must be taken into consideration, since it is the limiting factor as to the damping factor. Using a value of 8 ohms for the impedance of a speaker and 6 ohms for the dc resistance, computing for different values of internal output impedances the damping factors would appear as given in the table below. As it will be noted, using an output impedance of 0.5 ohm and a load impedance of 8 ohms, the damping factor is 16. Adding in the voice-coil resistance, the actual damping factor is:

$$\frac{Z_{LS}}{Z_{out} + R_{vc}} = \frac{8.0}{0.5 + 6} = 1.23$$

where,

Z_{out} is the internal impedance of the amplifier,

R_{vc} is the dc resistance of the voice-coil,

and Z_{LS} is the rated impedance of the loudspeaker.

It has often been stated that the ideal driving impedance for a loudspeaker would be zero impedance. However, it will be observed that with zero output impedance, the damping factor is 1.33, because the dc resistance of the voice coil is still in the circuit. It appears that a maximum damping factor of 20 is a practical value. There is now doubt that many loudspeakers sound better as the damping factor is increased, because of peaks in the reproduction, which in some instances rise from 10 to 12 dB because of impedance variation, as discussed in Questions 20.89 and 20.91.

20.104 What is the force factor of a loudspeaker?—The ratio of the effective force to the current flow in the voice coil.

20.105 Where is the most linear portion of a loudspeaker movement, and how is its efficiency calculated?—The most linear portion of a loudspeaker is where the cone displacement is proportional to the driving force. Its efficiency is the difference between the acoustic output and the electrical input power level, expressed in dB. As this is an involved formula and the test is to be

Ampl. internal impedance	*Damping factor*	*True damping factor (VC = 6.0 ohms)*
8 ohms	1.0	0.57
4 ohms	2.0	0.80
2 ohms	4	1.0
1 ohm	8	1.14
0.5 ohm	16	1.23
0.25 ohm	32	1.28
0.125 ohm	64	1.30
0.050 ohm	160	1.32
0.025 ohm	320	1.33
0.0125 ohm	640	1.33
0.0000 ohm	infinity	1.33

Fig. 20-103. Computing damping factors.

made under rigid couditions, the reader is referred to EIA Standard SE-103, April, 1949, reaffirmed March, 1954.

20.106 What is the relationship of diaphragm movement to the radiated power in a loudspeaker?—For a constant radiated power at the low frequencies, the voice-coil excursion must be quadrupled per octave.

20.107 What is the diaphragm movement in the average loudspeaker for 1 acoustic watt output?—
Typical values are given below.

Diameter	*Peak excursion*	*Resonant frequency*
8 inches	0.19 inch	60 to 150 Hz
12 inches	0.10 inch	30 to 85 Hz
15 inches	0.50 inch	25 to 55 Hz

20.108 What is acoustic elasticity?—The elasticity provided by the air (in a speaker enclosure) which is compressed when the cone moves in a backward direction.

20.109 What does the term "acoustical suspension" mean?—It is a loudspeaker so designed that the diaphragm can be moved with a very small amount of power at the voice coil. This is accomplished by supporting the outer periphery of the diaphragm with a very flexible supporting material. Such suspensions are used in small book-shelf-type enclosures. As an example, a conventionally supported diaphragm, having a free-air resonance of 70 Hz, when installed in an enclosure may increase the resonant frequency by 100 Hz or more. Thus, when placed in a small enclosure the low-frequency response falls off quite rapidly.

Increasing the compliance of the diaphragm lowers the free-air resonant frequency. However, if carried too far, the speaker can be driven into distortion quite easily. Therefore, the speaker is mounted in an airtight enclosure completely filled with an acoustic absorption material, which back-loads the diaphragm and returns it to its original resonant frequency. The resonant frequency can also be increased or decreased by changing the interior volume of the enclosure. Such designs are termed acoustic suspension systems, and can be applied to both low-frequency and midrange speaker units. Acoustic suspension systems are of low efficiency, generally 0.5 to 2 percent. (See Question 20.183).

20.110 Define the term "speaker directivity index."—It is the ratio, expressed in decibels, of the power which would be radiated if the free-space axial sound pressure were constant over a sphere, to the actual power radiation. The directivity index is measured by applying a constant input to the speaker mounted in an enclosure and to a microphone placed at a constant distance on the axis of the speaker, for different angles of radiation. The procedure for calculating the results is quite involved; therefore, the reader is referred to EIA Standard SE-103-1954.

20.111 What is the purpose of a shading ring in an electrodynamic loudspeaker unit? A shading ring is a heavy copper ring about ⅛-inch thick placed on the forward end of a field coil in an electrodynamic loudspeaker. The purpose of the ring is to act as a shorted turn, causing a heavy circulating current in the ring. This current bucks the flux from the field coil at 60 and 120 Hz, thereby reducing the tendency of the field coil to induce hum frequencies into the voice-coil circuit.

20.112 What are the factors which affect the design of direct-radiator type loudspeakers?—The high-frequeucy response is restricted by the mass reactance of the diaphragm. The low-frequency response is affected by the small radiation resistance. Increasing the high-frequency response by one or two octaves decreases the angle of radiation, the beam becoming quite narrow as the frequency increases.

As additional octaves are added to the frequency range, the nonlinear distortion is also increased. The enclosure in which the loudspeaker is housed plays an important part in the final frequency response.

The overall sensitivity may be increased by intensifying the density of the magnetic flux. However, this also increases the damping.

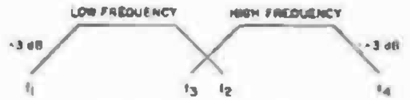
20.113 What is the effect of a loudspeaker connected across the output of an amplifier?—When a loudspeaker is connected across the output of an amplifier, it has the effect of connecting a capacitance or inductance in parallel with the output of the amplifier. Because the loudspeaker does not present a constant load impedance, the distortion characteristics of the amplifier may be affected to a considerable extent.

The average loudspeaker presents a capacitive load of 0.0025 μF or an inductive load of 1000 millihenries or greater, depending on the design. For a complete measurement, the interconnecting cable capacity should be included. Average cable presents about 40 to 80 pF of capacity per foot.

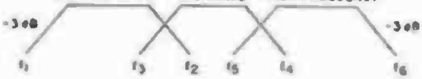
20.114 What is the axial sensitivity of a loudspeaker?—It is the acoustic output power, expressed in dB (referred to the threshold of hearing), that will be produced at a given distance for a stated power to the loudspeaker voice coil.

20.115 Show the crossover characteristics for multiple-loudspeaker systems.—The actual design of such networks is discussed in Section 7. However, crossover networks generally have crossover characteristics as shown in Fig. 20-115. It will be noted that for the two-way network at (a), the low- and high-frequency speakers are crossed over at a point 3 dB down from the uniform portion of the frequency band. The same is true for the three-way network of (b), showing two crossover points. The crossover frequency is dependent on the characteristics of the speaker units employed. The rate of attenuation may be 6, 12, or even 18 dB per octave. As a rule, 12 dB per octave is the one most generally employed.

General design considerations for crossover networks require that the crossover frequency should start becoming effective before the frequency response of the loudspeaker falls off and the loudspeaker becomes nonlinear. Loudspeaker units designed for low-frequency use in multiple installations



(a) *Two-way system.*



(b) *Three-way system.*

Fig. 20-115. Frequency characteristics for loudspeaker crossover networks.

are especially designed and seldom have much response above 1000 Hz.

The frequency response of a high-frequency unit must be restricted to frequencies whose wavelengths are such that the excursion of the loudspeaker diaphragm will not exceed the displacement as recommended by the manufacturer. If this precaution is not taken, serious damage to the diaphragm and voice coil will result.

It has been found from experience that for large commercial installations, the two-way system is the most satisfactory from a phase-shift standpoint, as any irregularities caused by phase shift are reflected in the reproduction. The preferred crossover frequencies for two-way systems are 500 to 800 Hz, with 500 Hz the preferred frequency. For three-way systems, the preferred frequencies are 500 and 5000 Hz.

20.116 Give graphically the effect of crossover-network insertion loss.—The insertion loss of a crossover network is quite important, particularly when used with an amplifier system of considerable power. In Fig. 20-116 has been plotted graphically the effect of network-insertion loss versus power-output loss for an amplifier system of 100 watts. It will be observed that the loss of power for an insertion loss of only 0.5 dB is 11 watts, and for 1 dB it is approximately 22 watts. Therefore, it is quite important that the insertion loss of a crossover network be kept to an absolute minimum. This can be accomplished by using large wire to reduce the dc resistance of the inductances and using capacitors of high quality. The line feeding the speaker system must also be of low dc resistance.

20.117 Where should a crossover network be connected?—The conventional method is to connect the network between the output of the power amplifier driving the loudspeaker system and the loudspeaker units of the system. Electronic crossover networks as described in Question 20.121 are connected between the output of a preamplifier and the input of two power amplifiers, one driving the low-frequency loudspeakers and the other driving the high-frequency loudspeakers.

20.118 What is the effect of a fast rate of cutoff in a crossover network?—Usually no effect is noted up to 18 dB per octave, unless there is a wide varia-

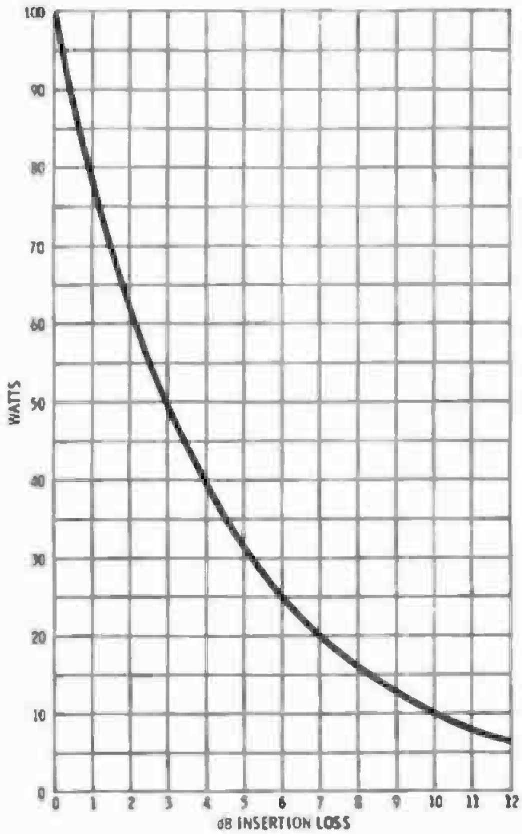


Fig. 20-116. Crossover insertion loss versus power-output loss as plotted for a 100-watt amplifier system.

tion of impedance match, which can cause ringing thus adding distortion to the reproduction. This is particularly true for the 18-dB-per-octave network.

A check for ringing can be made by connecting an oscilloscope to the output of the network in question, using a resistive termination. The frequency spectrum is swept by an audio oscillator, while keying the oscillator off and on. The ringing will appear somewhat like the waveforms in Fig. 1-37. Therefore, the same test should be made using the speaker system as the load, rather than a resistive load.

20.119 Are the phase relationships of a complex signal affected by a crossover network?—Phase shift is affected about two octaves above and below the frequency of crossover and will vary from 90 degrees for a 6-dB-per-octave crossover network to 180 degrees for an 18-dB-per-octave network.

20.120 How critical is the source impedance feeding a crossover network? —The output impedance of the amplifier should be terminated by a network that is equal to its rated load impedance, even though the amplifier internal output impedance may be considerably lower than the rated load impedance. However, this is only true for a constant-resistance type network. For an m-derived network, the impedance of the amplifier and network must be matched. It should also be kept in mind that a loudspeaker does not reflect a constant load impedance; this is discussed in Question 20.89.

20.121 What is an electronic crossover network and how does it function? —An electronic crossover network consists of two amplifiers each having a variable RC network at the input portion of the circuit. One network has a high-pass frequency characteristic; and the other, a low-pass characteristic. One section of the crossover network feeds the power amplifier driving the low-frequency loudspeakers, and the other section feeds the power amplifier driving the high-frequency loudspeakers. The crossover frequency is obtained by setting the controls of the networks to frequencies such as 400 Hz for the high-pass channel and 400 Hz for the low-

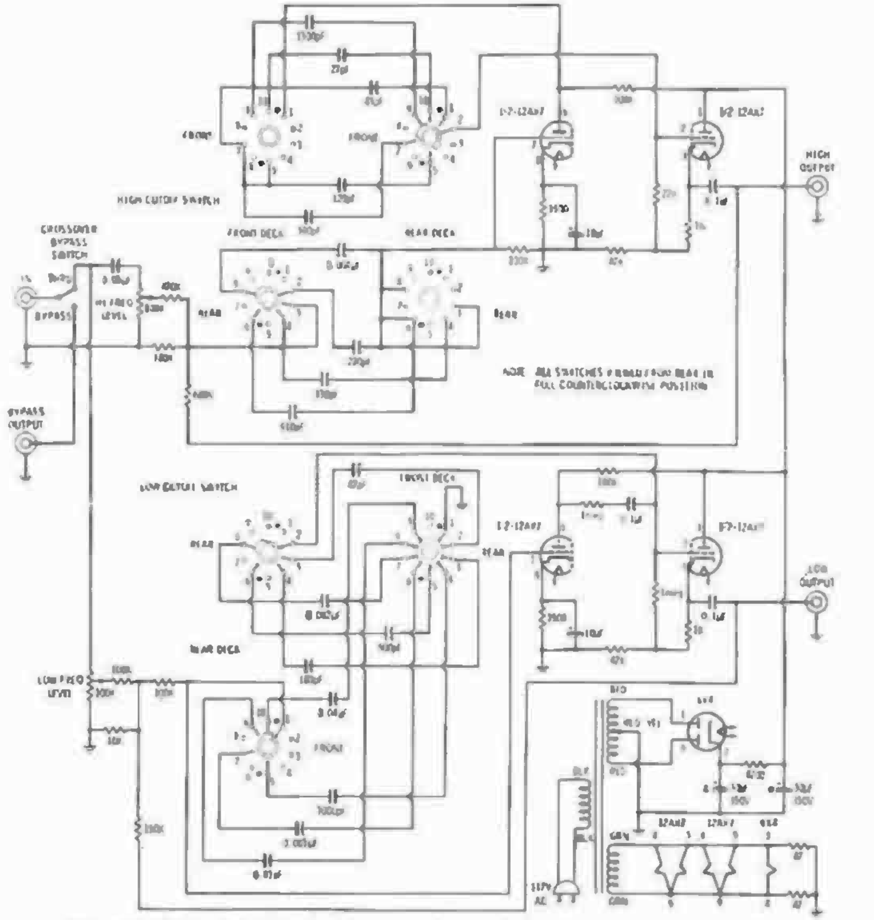


Fig. 20-121A. Schematic diagram of an electronic crossover network.

pass channel. This would result in a 400-Hz crossover frequency. The usual rate of attenuation in either section is 12 dB per octave.

Conventional crossover frequencies are 100, 200, 400, 700, 1200 200 and 3500 Hz. These frequencies will provide the proper crossover frequency for most loudspeaker systems.

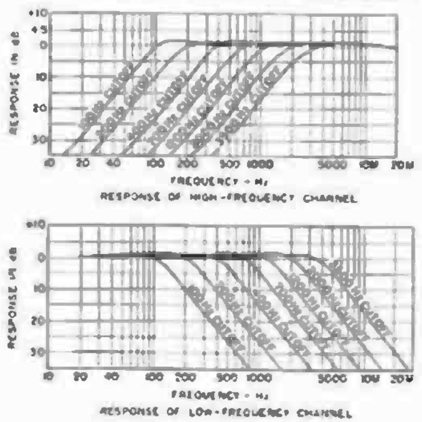


Fig. 20-121B. Frequency characteristics of RC networks in the electronic crossover network shown in Fig. 20-121A.

In Fig. 20-121A is shown an electronic crossover network developed for kit construction. In Fig. 20-121B is shown the frequency characteristics of the RC networks in the amplifier sections.

Referring to the schematic diagram, the crossover network consists of two amplifier stages composed of two 12AX7 dual triodes. The input signal is applied through a high- and low-frequency level control and through an isolating resistor to the input of each channel. The 12AX7 in each channel is utilized so that one section functions as a gain stage, and the other as an output cathode follower. Approximately 14 dB of negative feedback is employed around each channel from the output of the cathode follower to the control grid of the input stage.

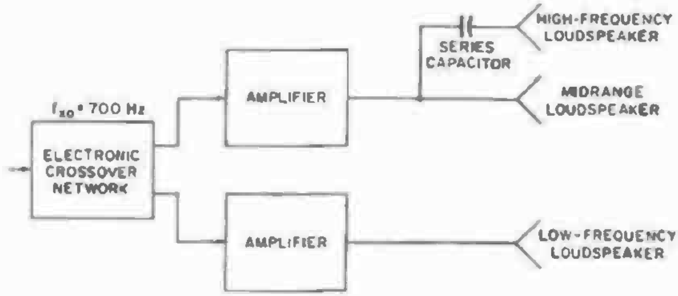


Fig. 20-121C. Three-way loudspeaker system using an electronic crossover network.

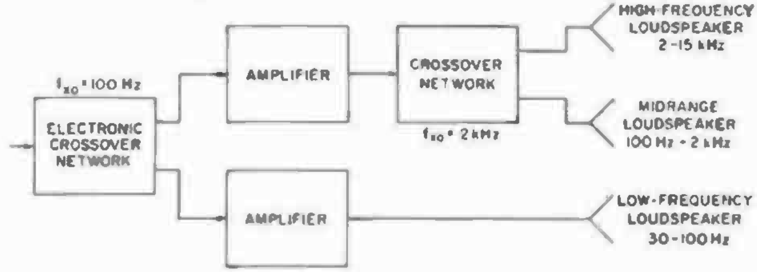


Fig. 20-121D. Three-way loudspeaker system using an electronic crossover network and a conventional network.

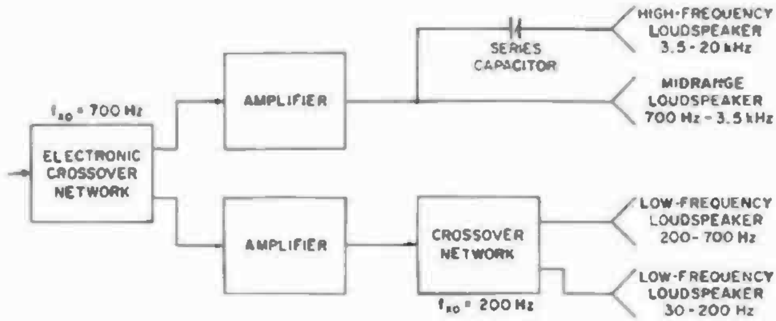


Fig. 20-121E. Four-way loudspeaker system using an electronic crossover network and a conventional network.

The purpose of the negative feedback is to modify the shape of the response curve in the region of cutoff, thus obtaining a sharper knee in the curve than would be obtained without negative feedback. This makes it possible to maintain a flat frequency response up to and including the cutoff frequency at a rate of 12 dB per octave. The use of negative feedback also reduces harmonic distortion in the passband of the amplifier stages. The amount of feedback employed reduces the network to a no-gain device, as no gain is required in this unit. Figs. 20-121C to E illustrate how the network may be connected for use with different types of loudspeaker systems. Although the electronic crossover network appears to be an ideal manner in which to obtain different crossover frequencies at will, it has certain drawbacks which must be considered seriously. Since the network is connected ahead of the power amplifiers, it leaves the high-frequency unit open to damage and possible burn-out. If a low-frequency pulse should leak through to the high-frequency unit, it can be severely damaged.

The use of this type network does not eliminate the need for a wide-range power amplifier, because to reduce phase shift in the operating bandwidth, the amplifier must extend several octaves above and below the operating range as discussed in Question 12.231.

Fig. 20-121F shows the schematic diagram for a three-channel transistor

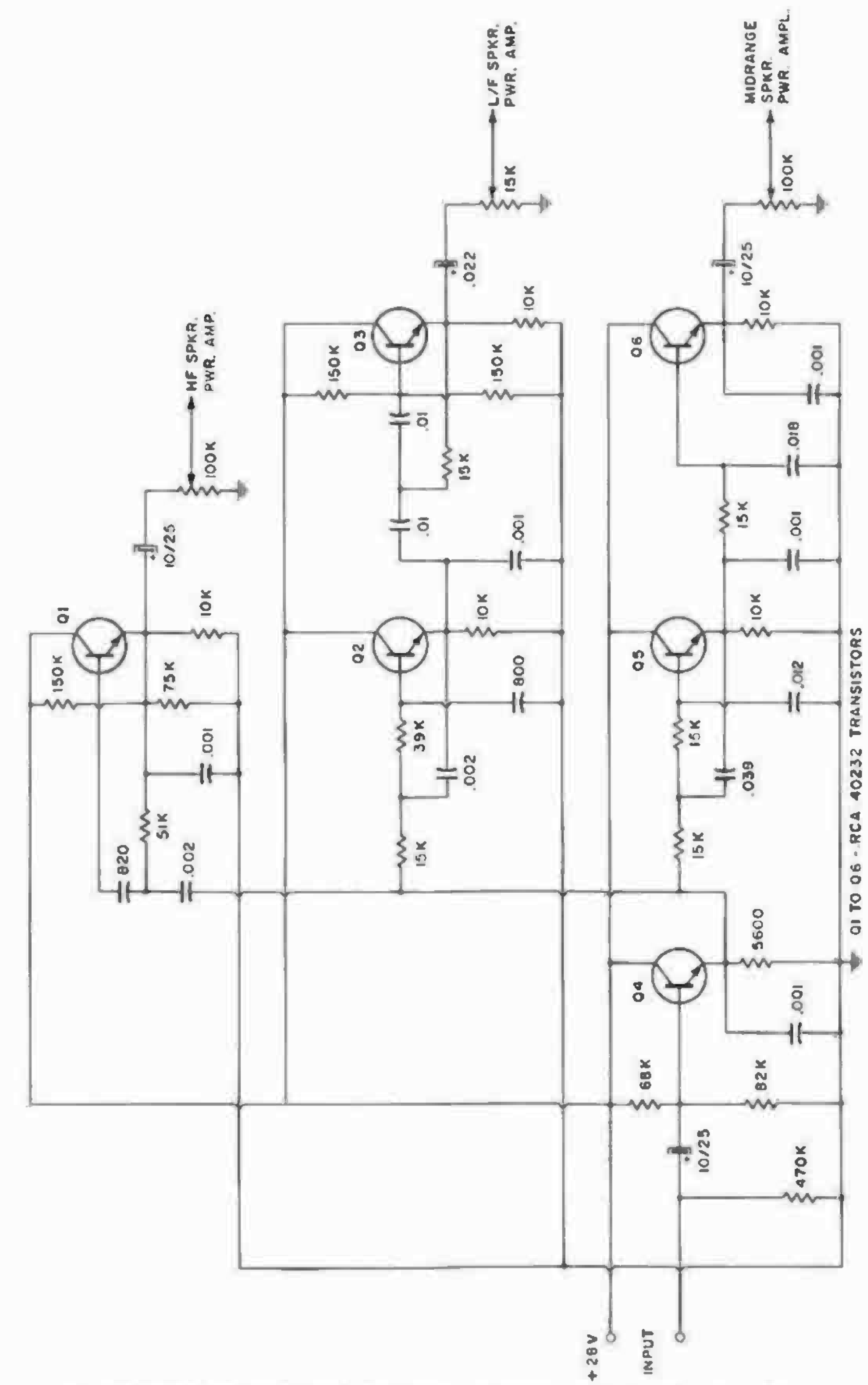


Fig. 20-121F. Three-channel electronic crossover network using transistors.

electronic crossover network designed by C. G. McProud. For stereophonic reproduction, two complete amplifiers are necessary. Three individual power amplifiers are used to drive the high-frequency, midrange, and low-frequency speakers. The power amplifier must be capable of developing the necessary high power for driving the low-frequency speaker unit. The crossover frequencies are fixed at 500 and 5300 Hz.

20.122 What is the procedure for phasing a multiple loudspeaker system? —To properly reproduce program material with a loudspeaker system employing two or more loudspeaker units, it is essential that the diaphragm of each loudspeaker unit in the system be in acoustic phase with each other. That is, the diaphragms must all move in the same direction at a given instant.

Phasing of multiple loudspeaker systems is difficult, and it is quite easy for the listener to become confused as to when the units are in phase. As a rule, when a system is out of phase, it will have a good low-frequency response as well as a good high-frequency response, but the overall response will be lacking in presence. This is particularly true of male voices when using a crossover frequency around 800 Hz. The voice will appear to jump or move from one speaker to the other.

The phasing of a multiple-speaker system must first be checked by ascertaining that the speaker units in the system are in phase electrically. Unmarked terminals must go to a common terminal of the network, and the plus or minus speaker terminals must go to the plus or minus terminals of the network, making sure they are connected to the proper section of the network. A signal of the same frequency as that of the crossover point to be checked is then applied to the input of the system. The exact center of the crossover point is found by rocking the oscillator over a small band of frequencies above and below the crossover frequency. The listener then moves in front of the areas of the high-frequency units. If they are in phase, the crossover will be smooth from one to the other. If the units are out of phase, a null point will be noted. To detect the null point, the volume will have to be carefully adjusted for a listening level.

Out of phase conditions are corrected by reversing the connection to one speaker unit only, not both. The same procedure is used for checking other units in the system, by comparing their phases to those known to be in phase. When all are correctly in phase, the signal appears to have a uniform level across all units.

If the system consists of a low-frequency unit and a multicellular horn, after completing the phasing, as above, the horn is moved forward and back, from a position where the front edge of the horn is in line with the low-frequency unit. Generally, a position will be found where a considerable improvement is noted in the quality reproduction, particularly in the crossover range.

20.123 What frequencies are used for theater crossover networks?—Usually, 400 or 500 Hz; however, a number of different frequencies have been used over the years. Some theaters are using a frequency of 250 Hz.

20.124 What effect does out-of-phase operation of loudspeakers have on the reproduction?—In a multiple-loudspeaker system, a lack of presence will be noted. In a public address system in which the loudspeakers are separated some distance from each other, no particular effect will be noted. However, if the loudspeakers are used in a cluster, they must be phased with reference to each other or dead spots will appear on the zero axis, as shown by the polar curve in Fig. 20-124.

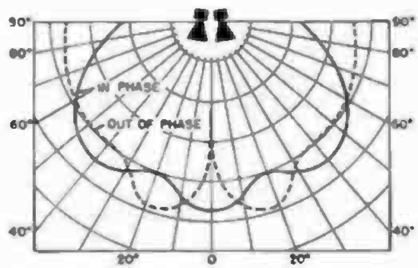


Fig. 20-124. Polar characteristics of two loudspeakers in and out of phase.

In the out-of-phase condition, the angle of sound distribution is increased.

20.125 How may two loudspeakers be phased electrically with respect to each other?—If the system does not employ a crossover network, they may be phased electrically by momentarily connecting a single flashlight battery to the terminals of each voice coil and noting the direction of the diaphragm deflections for a given polarity of the battery, and the voice coils. The voice coils are then connected across the output of the amplifier in a way that results in the same direction of deflection at a given instant for each loudspeaker.

20.126 How is a three-channel stereophonic theater system phased?—Before any attempt can be made to phase the loudspeakers, it must first be ascertained that all three channels are in phase electrically, beginning at the sound head and including the amplifiers and any other equipment connected in the circuits feeding the loudspeakers. Once this has been determined, the overall phasing of the loudspeaker system is made by creating a common connection between one wire of each loudspeaker system at the input to the dividing network. With equal and like signals from the three sound heads, measure the voltage between this common connection and the remaining wires of the networks 1 and 2 and then 1 and 3. If the phasing is correct, the voltages between the networks will be zero or a very small amount compared to the signal voltage at the input of any one network. Such measurements should be made at a frequency of 50 or 100 Hz to avoid indiscriminate phase shifts.

After phasing the overall loudspeaker system, the high- and low-frequency sections of each loudspeaker system must be phased for the best sound reproduction. To accomplish this, a signal of suitable level and of a frequency close to the crossover frequency is applied to the channel under test. From a listening position in the auditorium, compare the loudspeaker system output with the high-frequency wiring connected alternately one way and then the other. Do this for various physical positions of the high-frequency horn and select that physical position which gives the greatest cancellation of signal when the system is connected in one phase and the greatest loudness when connected in reverse order.

The connection giving the greatest output should sound smooth and relatively free from harmonics. This connection is the proper one. This adjustment should be checked for other positions in the auditorium to eliminate the effect of standing waves because of acoustic response.

The best positions for such tests are usually in the front and middle third of the seating area. The effect of correct phasing is to increase screen presence of the sound and smooth out the response, resulting in a more pleasing reproduction.

In the preceding tests, it is important that the crossover network of each loudspeaker system be connected for equal acoustic power from the high- and low-frequency loudspeaker sections before they are adjusted for auditorium acoustic response. Amplifier phasing is discussed in Question 23.104.

As a rule, theater equipment is phased by the manufacturer and the phasiug is indicated by a plus/minus mark on one terminal.

20.127 If the diaphragm of a high-frequency driving unit is not visible, how is its direction of motion checked?—Cover the throat with tissue paper. Apply a single flashlight cell to voice-coil connections in a given direction. When the diaphragm moves outward, the paper will be moved outward by air pressure within the unit. The unit is then connected to move in the same direction as the other units in the system.

20.128 Describe a phasing device for stereophonic recording and reproducing systems—The phasing (polarity) of microphones and loudspeaker monitor systems in a stereophonic recording system becomes extremely critical when a multiplicity of microphones are in use. Some method of quickly determining the phasing polarity of each channel from the microphone to the monitor speakers while the orchestra is rehearsing, is an absolute necessity. This is not only true for stereophonic recording, but also true for monophonic recording, when several microphones are being used to form a composite electrical sum-signal.

To avoid cancellation of the signals, it is necessary to ascertain if an unintentional phase-reversal exists between any group of microphones or channels.

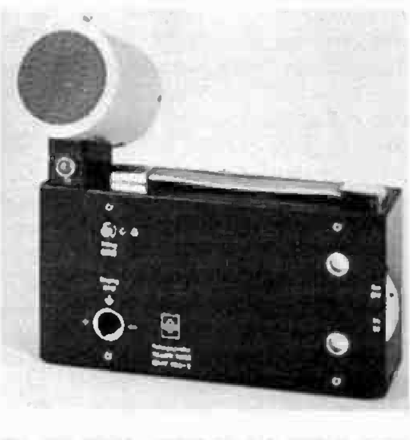
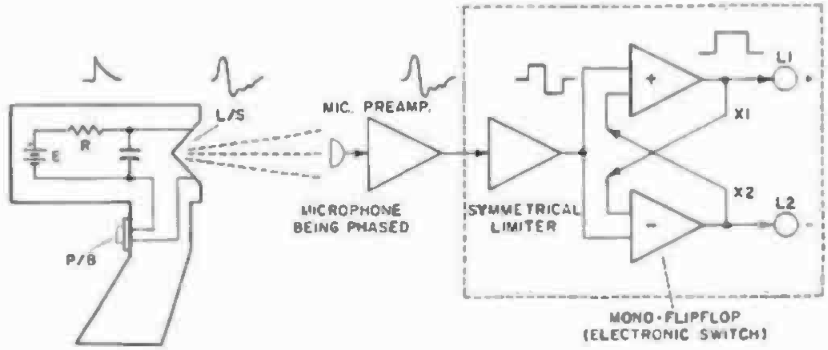


Fig. 20-128A. EMT Model 160 Polarity Tester. (Courtesy, Gotham Audio Corp.)



(a) *Pulse generator.* (b) *Polarity indicator.*

Fig. 20-128B. Block diagram for EMT-160 polarity indicator.

To accomplish this by ordinary means is not fast enough or entirely satisfactory, as an inadvertent phase reversal can take place in a cable to a microphone. Other items to be considered are amplifiers, equalizers, junction boxes, and loudspeaker systems. As a rule, recording systems are installed permanently in phase; thus, the most likely place for a phase reversal to occur is in microphones and their cables. A simple system of checking individual microphone phasing was discussed in Question 4.85. However, for a large recording setup, this is not completely satisfactory.

To speed up phasing and as a final check after the microphones have heen placed in position, a device developed by Institut für Rundfunk-technik of Hamburg, West Germany and manufactured by Elektromesstechnik Wilhelm Franz, also of West Germany, is shown in Fig. 20-128A, with its basic block diagram shown in Fig. 20-128B. The pulse generator at the left produces a single acoustic pulse of a definite polarity, by discharging a capacitor through a small loudspeaker unit mounted in a pistol-grip housing. The discharge causes the production of a positive pressure wave. The shape and duration of the pulse applied to the speaker unit is shown above the block diagram. The block diagram for the indicator unit is shown at the right, with a microphone and its preamplifier connected for a polarity test. An oscilloscope display of the acoustic waveform emitted by the loudspeaker is shown in Fig. 20-128C.

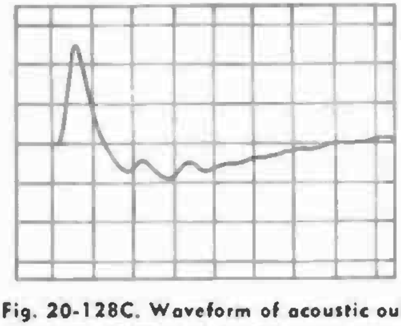


Fig. 20-128C. Waveform of acoustic output from EMT Model 160 pulse generator display on an oscilloscope.

Referring to the block diagram again, the acoustic pulse emitted by the generator (left) is changed into an electrical pulse by means of the microphone and its preamplifier, and then applied to the input of the polarity indicator. The pulse is passed through a symmetrical limiter, then to a flip-flop or electrical switch. The upper portion of the flip-flop (+) responds to only a positive pulse, and the lower section (—), to only a negative pulse. The positive peak pulse of the waveform triggers the upper section of the electronic switch, lighting indicator lamp L1. In this condition, the lower circuit is blocked through line X1. Therefore, the lower circuit cannot respond to a negative pulse. This quasi-stable state lasts for several seconds. During this time, the upper positive indicator lamp L1 remains in the on position, and the lower section remains blocked.

After the lapse of the predetermined time lag, the upper section returns to its rest position, and the blocking of the lower section is canceled. The resulting indication is always determined by the polarity or phasing of the first half of the test pulse. As an example, if a phase reversal (negative) occurs at tbe output of a preamplifier, the lower section of

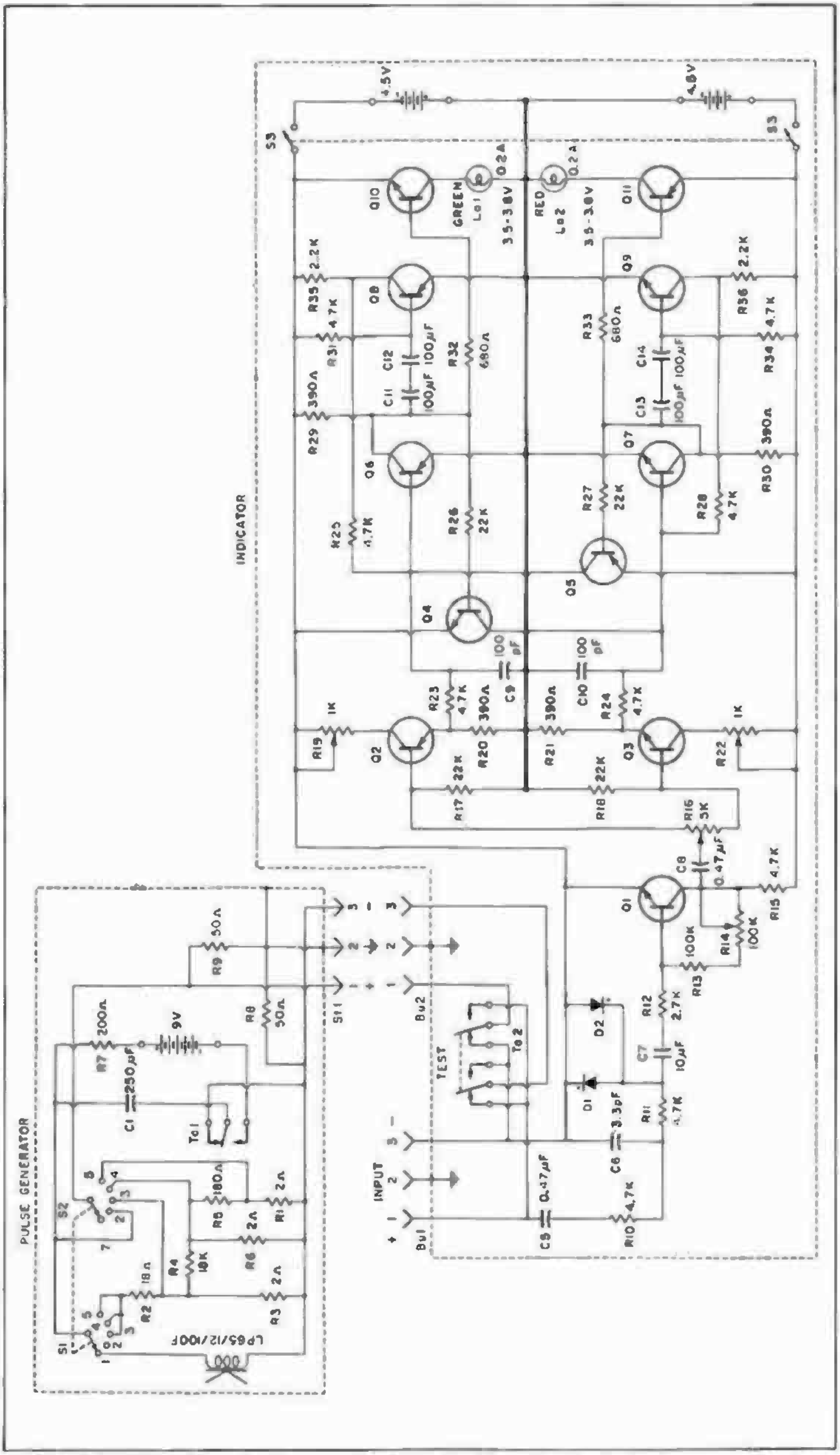


Fig. 20-128D. Schematic diagram for EMT Model 160 polarity indicator and pulse generator. (Courtesy, Gotham Audio Corp.)

the electronic switch will trigger first, lighting lamp L2, indicating a negative pulse, and at the same time block the upper section of the switch through circuit X2.

The pulse generator section is also equipped with several electrical outputs of varying magnitude for the direct testing of amplifiers, loudspeakers, etc. For loudspeaker testing, a pulse is taken directly from the generator, and a microphone of a known polarity is placed in front of the speaker. The indicator unit is then used to indicate the acoustic polarity of the speaker unit.

During a recording session when the microphones are placed in their final positions, the generator gun is placed in front of the microphone and the indicator unit is placed at the monitor speakers of a given channel. Thus, the entire system can then be checked in a matter of seconds. The most likely place for a phase reversal to take place is in a microphone or microphone cable. A complete schematic diagram of the polarity indicator appears in Fig. 10-128D.

The characteristic of the acoustic pulse is a positive pressure wave of 300 microbars at a distance of 2 inches from the microphone diaphragm. For the direct testing of loudspeakers and other devices, 9 volts are available for impedances between 4 and 64 ohms, with lower voltages for other type testing. The indicator unit requires a signal voltage of 1.55 volts, or plus 6 dBm. The switching and blocking time is less than 1 microsecond. The unit is completely transistorized. When the instrument is not in use or in storage, the pulse generator (gun) is placed in a slot at one end of the indicator unit, as shown in Fig. 20-128A. Several different uses for the polarity tester are shown in Fig. 20-128E.

20.129 What is the best height above floor level for a loudspeaker?—For persons sitting down, approximately 40 inches above floor level, because this is the average ear level.

20.130 What are surround speakers?—Loudspeakers placed at the sides and rear of an auditorium to create sound perspective.

20.131 Where should loudspeakers be placed for sound reinforcement?—Above the sound source, if possible, as this results in a more realistic reproduction.

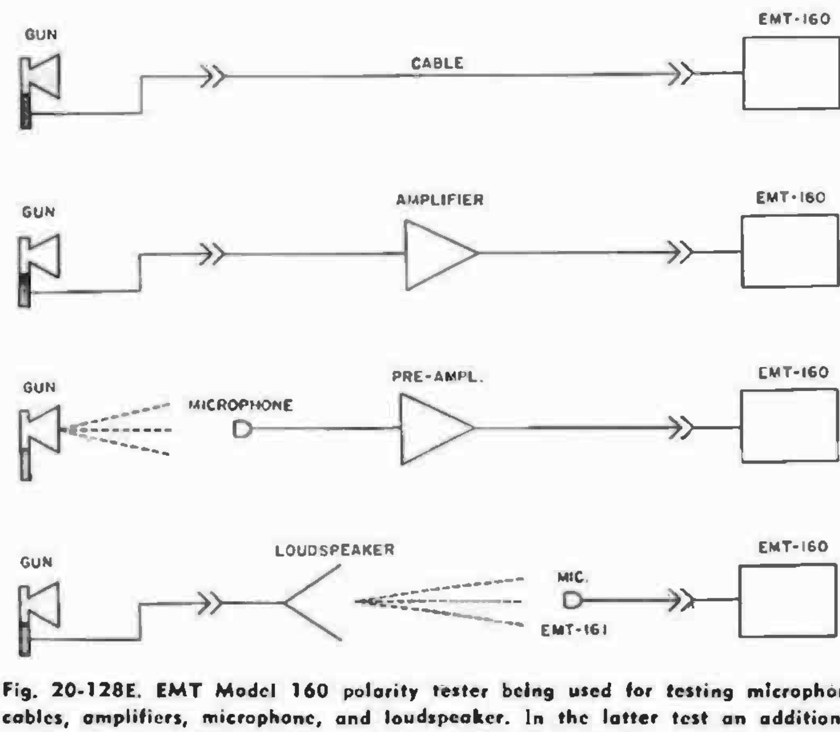


Fig. 20-128E. EMT Model 160 polarity tester being used for testing microphone cables, amplifiers, microphone, and loudspeaker. In the latter test an additional microphone and preamplifier are required.

20.132 Describe the phasing procedure for a two-channel home installation. —The first step in phasing is to check the polarity marking on the individual speaker enclosures to ensure that the plus or minus terminals are connected to the plus or minus terminals of the amplifier system, and that the left and right sides are properly established (the left speaker is the one to the left when facing the enclosure). The tone controls on the amplifier system are set to their flat position. A monophonic record (a single vocalist is best) is played in the stereophonic position. The sound levels from the right and left amplifiers are adjusted for an equal level from both speaker systems. The stereo balance control is now adjusted until the sound appears to be coming from a position between the two speaker systems. This may require some adjustment of the enclosure angles, relative to the listening area. To a person walking from side to side, the sound should appear smooth without a hole in the center. (See Question 20.160.)

If the sound in the listening area appears to have a hole in the center, reverse the leads to one speaker (not to both). Upon readjustment of the stereo balance control, the hole should disappear. These tests are based on the assumption that the leads from the two sides of the stereophonic pickup unit are properly phased. If a stereo control center is being used, turning the pickup reverse switch to reverse should deteriorate the sound considerably. This assures the system is now in correct phase. Special test records are available for phasing, for adjusting the center-balance and sound-level controls.

20.133 Describe a theater loudspeaker installation.—The majority of motion picture theater loudspeakers consist of a two-way system employing a low-frequency section, which crosses over at 400 to 500 Hz.

It appears that the first multiple-speaker systems experiments were performed by Rice and Kellogg in 1923. In their system, three horns were used, and each horn covered a certain portion of the frequency spectrum. A high-frequency unit was described by Bostic in 1930. However, nothing was done with these systems commercially until the introduction of sound motion pictures.

The first commercial systems used the curled horn (Fig. 20-59). Later, low-

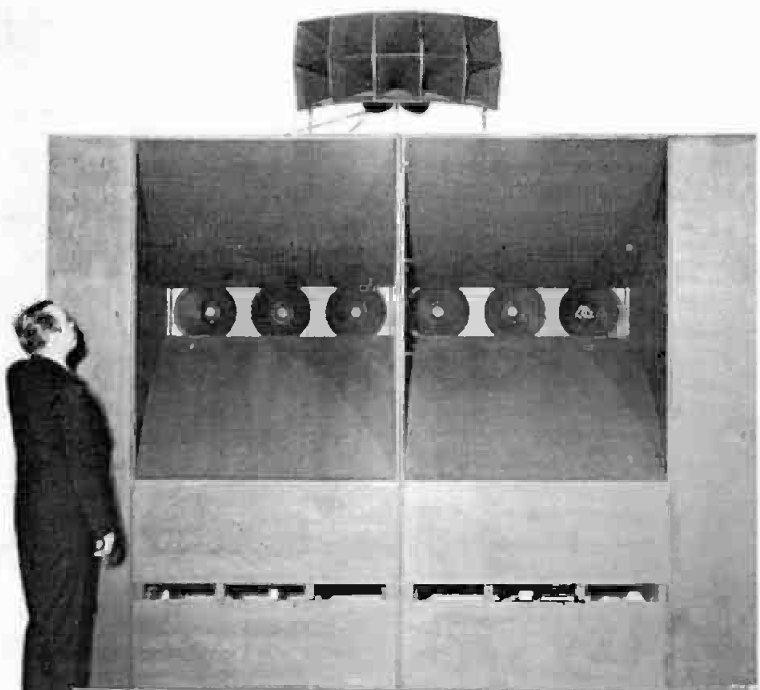


Fig. 20-133A. Altec-Lansing theater type, two-way loudspeaker system.

frequency dynamic speakers in a baffle were added and below the horns, with the addition of two high-frequency units which used a small horn about 4 to 6 inches in length. The first two-way folded horn system as used today was developed and built by Shearer and Hilliard in 1934 at the MGM Studios in Culver City, California. Generally speaking, the basic design has undergone little change, except for the improvement of the flare rate by Olson and the design of the speaker units by various manufacturers. Electrical Research Products Inc. (ERPI now Westrex) and RCA Manufacturing Co., in collaboration with J. B. Lansing Co., contributed much to the early development of multiple-speaker systems.

A theater system, manufactured by the Altec-Lansing Corp. and typical of systems nsed in the larger motion picture theaters, is shown in Fig. 20-133A. Its construction consists of two horn-type baffles, and each baffle has three low-frequency speaker units and a 2 × 5 multicellular high-frequency horn, driven by two driver units attached to a Y-shaped throat, placed above the two low-frequency assemblies. The two slots at the lower portion of the system are parts, similar to a bass-reflex enclosure, for increasing the low-frequency response, as discussed in Question 20.60. The system uses an LC network crossing over at 500 Hz, with a cutoff rate of 12 dB per octave. A cross-sectional view is given in Fig. 20-133B. For smaller theaters, only one low-frequency unit is used.

Fig. 20-133C shows the front view of an RCA PL 304 speaker assembly, which is also used quite frequently in rerecording (dubbing) stages for monitoring purposes. At A is a low-frequency horn (commonly called a bath tub), with two 15-inch low-frequency speaker units B, feeding the throat of the horn. At C and CC is shown the curved mouth of the horn, which disperses the sound several degrees in the vertical plane. At D and DD are ports for increasing the low-frequency response. Above the low-frequency horn is a high-frequency horn E, driven by a single driver unit (not shown). The crossover network is mounted on the rear of the lower horn. The overall width of the assembly is 47 inches;

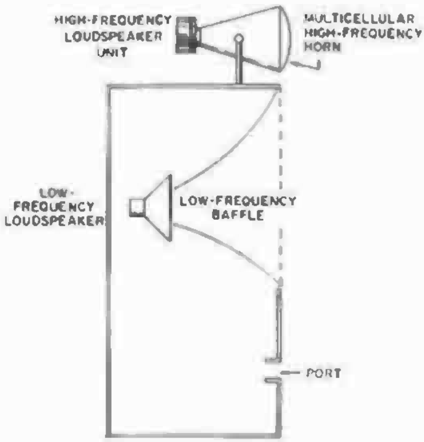


Fig. 20-133B. Cross-sectional view of Altec-Lansing two-way theater loudspeaker system shown in Fig. 20-133A.

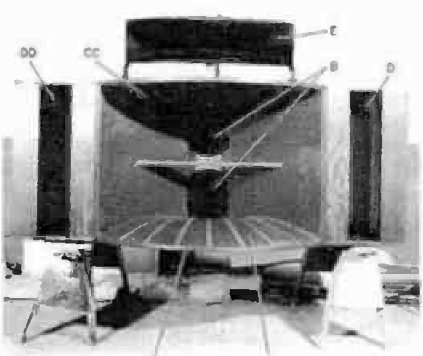


Fig. 20-133C. RCA theater loudspeaker system type PL304.

Fig. 20-133D. RCA monitor speaker system LC9A.

Fig. 20-133E. Altec-Lansing A7 loudspeaker system. This system is a smaller version of the larger systems, and is especially suitable for small theaters, auditoriums, motion picture review rooms, and rerecording stages.

height is 40 inches; depth is 32½ inches. The low-frequency units are each 32 ohms impedance and are driven in parallel. The impedance of the high-frequency unit is 16 ohms.

A smaller, two-way speaker system, the RCA LC9A, is pictured in Fig. 20-133D. The system is a somewhat smaller version of the one shown in Fig. 20-133C. The three speaker systems mentioned are of wide-frequency response and are capable of carrying high power with low distortion.

The loudspeaker shown in Fig. 20-133E is a Model A7 system, manufactured by Altec-Lansing, especially designed for small theaters, auditoriums, and motion picture review rooms and rerecording stages. The enclosure A houses a single low-frequency unit B, coupled to a short horn C. A slot D at the bottom acts as a bass-reflex enclosure. The multicellular baffle E spreads the high frequencies over a 120-degree arc. The crossover frequency is 500 Hz. The same speaker obtained in a suitable enclosure for use in locations where appearance is also essential is known as Model A-7-500.

Speakers designed as pictured in Figs. 20-133C, D, and E are termed "sort horns." The horn is designed with a large throat area, with a rapid flare rate, and a cutoff frequency below 100 Hz. Using this horn in combination with a bass reflex enclosure and a high-frequency horn results in a wide-frequency response. The frequency response for such a design is given in Fig. 20-133F and is typical of such designs. This basic design also applies to the theater speaker system of Fig. 20-133A.

It should be recognized that theater speaker systems are not required to have as wide a frequency response as

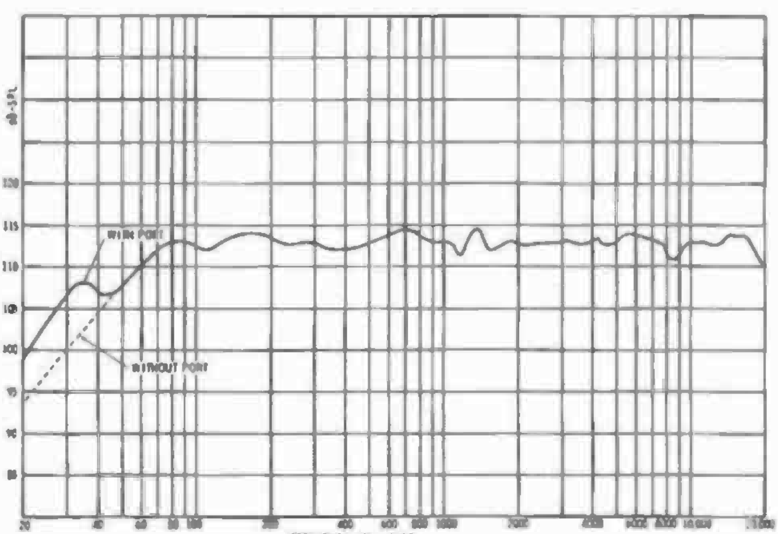


Fig. 20-133F. Typical frequency response for a short horn loudspeaker operating in conjunction with a bass-reflex enclosure and high-frequency multicellular baffle.

those used for hi-fidelity reproduction in the home, although for the most part they do have quite a broad frequency response. Generally, the frequency range is 30 to 10,000 Hz. As a rule, a low-pass filter is used in the reproducing equipment to limit the response, as discussed in Question 19.78.

20.134 ***What is the vertical and horizontal spread in degrees for a theater-type, high-frequency multicellular horn?***—About 105 degrees in the horizontal plane and 40 degrees in the vertical plane.

20.135 ***Describe the precedence effect.***—This effect was discovered in 1933 by F. K. Baker of the Bell Telephone Laboratories and applies to the reproduction of stereophonic sound. When a single sound is reproduced from two loudspeakers and the sound from one speaker is delayed by several milliseconds, the listener will hear the sound as if it came only from the loudspeaker where he first heard it. The listener also will judge the second speaker to be silent.

20.136 ***What is a loudspeaker presence equalizer?***—A resonant circuit equalizer used in a theater sound reproducing system to increase the frequency response of the midrange frequencies. Geuerally, it is adjusted for a frequency of 2700 Hz, using a rather broad peak. The amount of equalization required will depend on the type of high-frequency units and the acoustic response of the auditorium. (See Question 6.69.)

20.137 ***What is a voice filter in a theater loudspeaker system?***—A parallel-resonant circuit connected in series with the line feeding the loudspeakers. Its purpose is to remove the tubbiness of the male voice. The frequency of resonance is adjusted somewhere between 125 and 300 Hz. The amount of filtering will depend on the auditorium acoustics and the loudspeaker system. The circuit diagram and method of connecting the filter in the line is shown in Fig. 20-137.

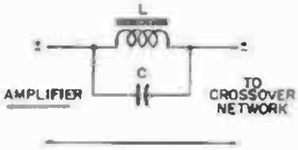


Fig. 20-137. Voice filter for removing the tubbiness of the male voice in a theater loudspeaker system.

In large auditoriums or theaters where the acoustic peaks may approach large proportions at several different frequencies, the peaks may be reduced or eliminated by the use of several filters, and the acoustic gain of the auditorium increased. This is discussed in Question 2.117.

20.138 ***Describe a loudspeaker protective circuit.***—When operating a loudspeaker system with high-powered amplifier systems (particularly home-type equipment), the high-frequency speaker unit can be easily damaged by the sudden surge of output power or by an oscillation of the amplifier system. A network to prevent such damage is shown in Fig. 20-138A.

At the low frequencies, the network is practically an open circuit because of the high impedance of capacitor C. As the frequency is increased, the reactance of the capacitor decreases. If the amplifier should break into oscillation, the capacitor reactance drops to a very low value (almost zero) and leaves only the resistor in the circuit to act as a load across the output of the amplifier and to prevent damage to the high-frequency loudspeaker unit.

Such networks should always be connected across the output of an amplifier when an electrostatic loudspeaker is used, unless it is known that the amplifier will not oscillate with capacitive loads. (See Question 12.167.) Average values of R and C for a 16-ohm output impedance are 47 ohms in series with 0.10 μF.

Such networks may also be used with transistors systems, but they offer little protection, since at very low frequencies the input impedance of a loudspeaker system is at its lowest and acts as a short circuit to the output of the transistors. Because of the fast rise tine of a transistor as compared to a vacuum tube, a faster-acting protective circuit is required. In this instance a pair of di-

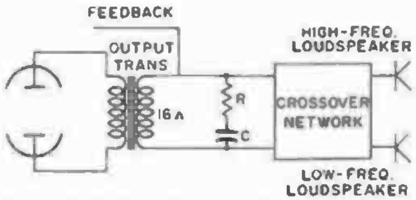


Fig. 20-138A. Method of connecting a tweeter-saver network across the output of an amplifier.

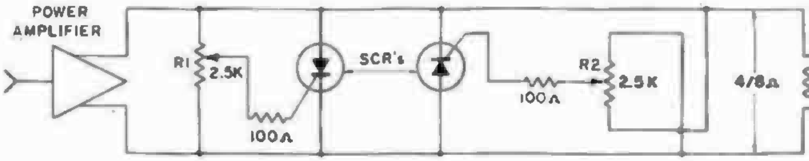
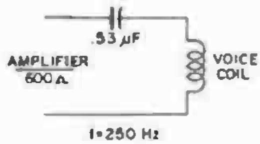


Fig. 20-138B. Loudspeaker protective circuit using two silicon controlled rectifier (SCR) units.

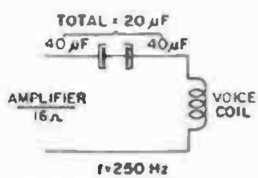
odes connected back-to-back or silicon controlled rectifiers SCR's may be used. Several manufacturers of high-power transistor amplifiers have included elaborate protective circuits in their amplifier circuitry. Such a circuit is shown at the lower right in Fig. 12-256A.

A protective circuit using two SCR's is shown in Fig. 20-138B. The firing point of the two SCR's is adjusted by potentiometers R1 and R2. It will be noted that R2 is reverse-connected to permit the two potentiometers to be ganged to move in the same linear direction. The firing point of the SCR's is dictated by the peak power output of the amplifier stage. The principal drawback to the use of SCR's (commonly used for rectification) is their limited frequency response. However, there are available such devices with wide frequency response, and these are the type that should be used. When this circuit is adjusted properly, it provides a fast-acting protective circuit.

20.139 ***What is the recommended frequency response for background music?***—If the ambient noise level is low, from 80 to 6500 Hz. A volume-limiter amplifier should be included to prevent overloading. If the ambient noise level is high, the frequency response will require tilting upward at the low and high frequencies.



(a) *600-ohm line.*



(b) *16-ohm line.*

Fig. 20-140. Capacitors connected in series with a loudspeaker to reduce the low-frequency response.

20.140 ***How may the low-frequency response of a loudspeaker be reduced?***—A convenient method of reducing the low-frequency response of an individual speaker unit for speech is by connecting a capacitor in series with the voice coil (Fig. 20-140). Paper or oil-filled capacitors are preferred; however, if these are not available, two electrolytic capacitors may be connected back-to-back, as shown in (b) of Fig. 20-140.

The value of the capacitor for a given cutoff frequency may be calculated by:

$$C = \frac{79{,}600}{f \times Z}$$

where,

C is the capacity in microfarads,
f is the frequency of cutoff in Hz,
Z is the line impedance in ohms.

A typical value for a 16-ohm line with a cutoff frequency of 250 Hz would be 19.9 µF.

20.141 ***How are L-type controls connected to a loudspeaker line?***—They are connected as shown in Fig. 20-141, with the total resistance toward the network or amplifier, and the variable portion of the L-type control toward the speaker voice coil.

20.142 ***What type volume controls are recommended for loudspeakers?***—Either the L or T types. In the former type, the impedance is constant in only one direction. In the latter, the impedance is constant in both directions.

20.143 ***Is it permissible to connect loudspeakers in series?***—The connection

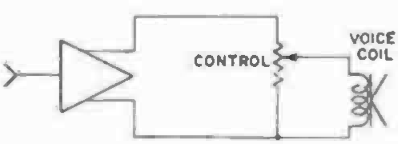


Fig. 20-141. Proper connections for a potentiometer or L-type attenuator in a loudspeaker transmission line.

of loudspeakers in series is to be discouraged because of resonant peaks in the speaker characteristics. At the resonant peak, the greatest amount of power is required; and the impedance of the loudspeaker is at its highest value, and the current is at its lowest.

As a rule, no two loudspeakers have the same peak resonant frequency; therefore, if two or more loudspeakers are connected in series, several resonance points are developed. This means that for a given power level, the power will differ for each loudspeaker and will depend on the frequency of the peaks.

Under such conditions, the loudspeakers will not have the same acoustic output level. Reproduction from such a connection will result in heavy resonant peaks which can become quite objectionable.

Connecting a 4-ohm and an 8-ohm speaker in series will result in the same current flowing through both voice coils simultaneously. Since power is equal to I^2R, the 4-ohm speaker will receive only half the power that the 8-ohm unit receives. However, since the impedance of the two speakers will not vary in the same relationship, this does not hold true at all frequencies. When a group of loudspeakers are to be driven from a common source, they are connected in parallel, and care should be taken that the impedance match is satisfied. (See Question 24.70.)

20.144 *What is the frequency range recommended for a high-quality paging system?*—Because most of the intelligibility of speech lies between 250 and 4500 Hz, it is unnecessary to reproduce above or below these frequencies.

20.145 *What is the approximate ratio of dc resistance to the ac impedance of a loudspeaker voice coil?*—A rule of thumb method of determining the impedance of an unknown voice coil is to measure the dc resistance of the coil and increase the value by 15 to 20 percent. Thus, a voice coil with a measured dc resistance of 6.5 ohms would have an impedance on the order of 8 ohms.

20.146 *Are special speaker units required for musical instruments?* — Yes, very rugged speakers are required to handle the high power transients produced by such instruments, particularly the guitar, bass guitar, string bass, and electronic organ. Because of the practice of playing these instruments at high levels and intentionally introducing transient distortion to obtain certain effects, the average high-fidelity loudspeaker unit designed for home use will not stand up under the continuously high-power output demanded. Thus, speakers using a diaphragm of special design, materials, and suspension are required. The average range of the speaker unit is 40 to 8000 Hz since the above-mentioned instruments fall within that range.

20.147 *What wire sizes are recommended for loudspeaker transmission lines, with respect to the frequency response?*

Wire Size	*Maximum Length in Feet (Pair)* 3 kHz	5 kHz	7.5 kHz	10 kHz
10	5000	3800	3000	2700
12	4000	3100	2500	2000
14	3000	2400	2000	1700
16	2500	1900	1500	1350

20.148 *When is a line considered to be of low impedance?*—When the feed lines are connected from the output of the driving amplifier directly to the loudspeaker voice coil, they are called low impedance. Generally, the output impedance of the amplifier lies between 4 and 30 ohms.

20.149 *When should a high-impedance line be used?*—When the distance between the driving amplifier and the loudspeakers is greater than indicated in the wire table of Question 20.151, or when the power losses are 15 percent or greater. Under these conditions, lines of higher impedance are desirable. As a rule, a high-impedance line is between 125 and 600 ohms. The signal is transmitted at high voltage and low current. Impedance-matching transformers are used at the loudspeaker end to match the low impedance of the voice coils to the line impedance.

20.150 *What factors determine a line impedance?*—The impedance of the line is determined by the source and terminating impedances. In audio-frequency work, the surge impedance of a transmission line is generally neglected.

A transmission line is said to have an impedance of 600 ohms when the source and terminating impedances are 600 ohms.

20.151 What are the wire sizes and maximum lengths recommended for low-impedance lines?

Wire Size	*Load Impedance* 4 Ohms	8 Ohms	16 Ohms
14	125 ft.	250 ft.	450 ft.
16	57 ft.	150 ft.	300 ft.
18	50 ft.	100 ft.	200 ft.
20	25 ft.	50 ft.	100 ft.

20.152 What are the wire sizes and maximum lengths recommended for high-impedance lines?

Wire Size	*Load Impedance* 100 Ohms	250 Ohms	600 Ohms
14	1000 ft.	2500 ft.	5000 ft.
16	750 ft.	1500 ft.	3000 ft.
18	400 ft.	1000 ft.	2000 ft.
20	250 ft.	750 ft.	1500 ft.

20.153 What variation in frequency response may be expected from a loudspeaker?—Loudspeakers, like any other commodity, vary in their quality depending on the design, price, and the manner in which they are used. For the average-priced unit, the frequency response will vary from 4 to 6 dB. However, in the better quality units the variation will be considerably less. In a well-designed, two- or three-way system, the variation may be within 2 dB, depending on the type units, crossover network, and the enclosure employed. The latter item has a considerable effect on the final response.

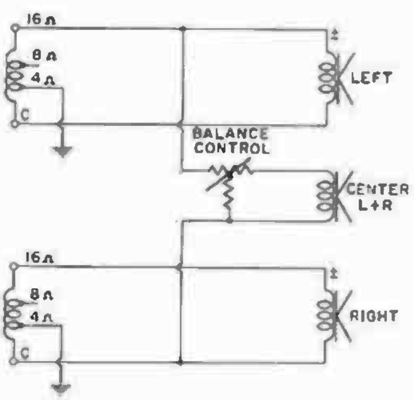


Fig. 20-154A. Center-derived channel. In this circuit the ground connection in the amplifier is returned to the four-ohm tap rather than to the common terminal. The center channel is the *sum* of the left and right channels.

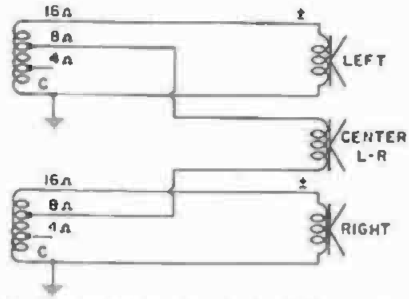


Fig. 20-154B. Center-derived channel. In this circuit the grounds are connected to the common terminal of the output circuit, resulting in a signal that is the *difference* between the left channel and the right channel.

20.154 Show the method for deriving a center-channel speaker from a two-channel stereophonic speaker system.—In a well-designed two-channel stereophonic reproducing system, a center speaker is usually unnecessary. However, if a center channel is required, it may be derived as shown in the circuits of Figs. 20-154A, B, C. The most desirable circuit is one that results in a *sum* signal of the left and right channels (Fig. 20-154A). It will be observed for this circuit that the ground connections of the output transformer have been moved from the common terminal to the 4-ohm connection. This will, however, necessitate some readjustment of the negative-feedback loop components. In some instances, a *difference* signal (Fig. 20-154B) may be quite satisfactory, since this method eliminates the need

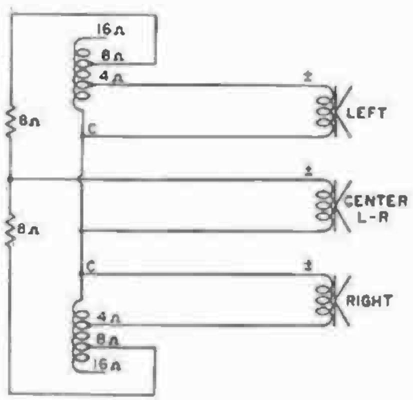


Fig. 20-154C. Center-derived speaker connections using conventional amplifiers. The signal at the center speaker is the *difference* between the left and right sides (after Klipsch).

for rewiring the output transformer circuit, and readjusting the negative-feedback loop. A center-derived circuit suggested by Paul Klipsch is given in Fig. 20-154C. The output of this circuit also is a difference signal.

After installing any of these circuits, the sound level of the three speaker systems must be carefully balanced and phased

20.155 ***Describe a multiunit loudspeaker cluster.***—Because of the directional characteristics of the high frequencies, some means must be taken to spread them over the listening area. This may be accomplished by the use of multicellular horns, a multicellular baffle, or a cluster of small high-frequency speaker units, as pictured in Fig. 20-155. Here is shown a high-frequency cluster, Model B200YA, manufactured by R. T. Bozak Manufacturing Co. It consists of eight individual units, having a frequency response from 1500 to 20,000 Hz. The recommended crossover frequency for these units is 2500 Hz. The high frequencies are spread over an arc of about 130 degrees. Each unit employs an 8-ounce Alnico-V permanent magnet. The units are connected in series-parallel to present an 8-ohm impedance.

20.156 ***Does a loudspeaker generate subharmonics?***—Yes, at one-half, one-third, and one-fourth the fundamental frequency. However, only the subharmonic at one-half the fundamental frequency is of any consequence.

20.157 ***What is radiation resistance?***—It is the measure of effective air load into which the loudspeaker system dissipates its acoustic power, thus producing sound waves.

20.158 ***How can reverberation in an auditorium be reduced without acoustic treatment?***—There is no substitute for acoustic treatment. However, the effects of reverberatiou may be diminished by the following methods:

a. If the auditorium is quite large, a great number of loudspeakers may be installed around the side walls or in front of the sound source. In either method, the speakers must be operated at a level that will not cause reflections from the opposite wall. This method depends on the distributiou of the sound rather than on the power of the sound.
b. A group of loudspeakers with directional baffles may be hung from the ceiling and pointed to a position where the sound is reflected from both the floor and the side walls. Absorption obtained from the audience in this manner reduces reverberation.
c. A third method is to reduce the response of the loudspeaker system at the low frequencies. This may be accomplished by altering the frequency response of the amplifier system or by connecting a capacitor in series with the transmission line or certain loudspeakers.
d. A fourth method, although somewhat costly and involved, is to reduce the aconstic peaks by tuning the enclosure, as discussed in Questions 2.117 and 17.136.

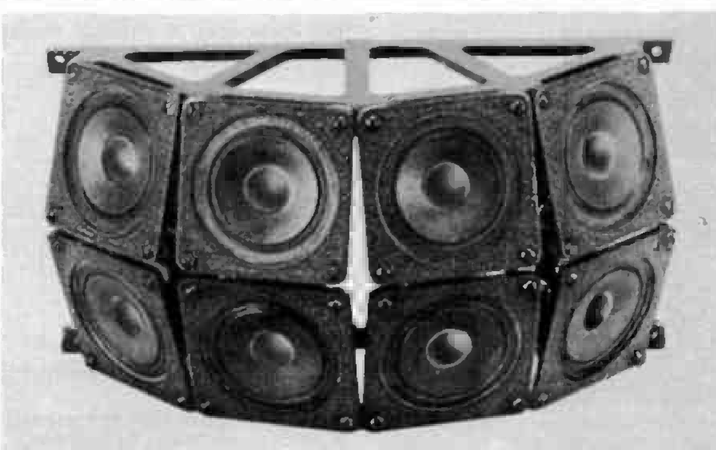


Fig. 20-155. High-frequency loudspeaker cluster Model B200YA manufactured by R. T. Bozak Manufacturing Co.

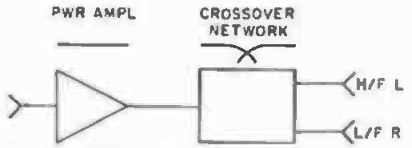


Fig. 20-159. Monophonic reproducing system connected for pseudostereophonic reproduction.

20.159 *Describe a pseudostereophonic reproducing system.*—Pseudostereophonic reproducing systems are not true stereophonic systems, but only simulate the stereo effect. A monophonic reproducing system may be made to reproduce a pseudoeffect as shown in the diagram in Fig. 20-159.

A monophonic reproducing system is shown driving two loudspeakers from a simple speaker crossover network (6 dB p/o) to limit the frequency range of the speakers. It will be noted that the speaker on the left is limited to frequencies above the crossover frequency, and the one on the right is limited to frequencies below the crossover frequency. The crossover frequency may be from 400 to 800 Hz, depending on the speakers employed.

Although the circuit shown will give a fairly good imitation of stereophonic reproduction when playing monophonic records, it can be easily detected as not being true stereo by an experienced stereophonic listener. The network can be designed from the data given in Section 7. The speakers are phased as for any stereophonic reproducing system.

20.160 *How are loudspeakers placed for two-channel stereophonic reproduction?*—For the proper placement of loudspeaker enclosures in a living room, several factors enter into the problem—the dimensions of the room, its configuration, available wall space, furniture, and the location of the listening area. To attain true stereophonic reproduction, it may take some experimenting in the adjustment and placement of the enclosures. Fig. 20-160A shows two corner speakers, with the approximate listening area. Here the angle of the enclosure is fixed, and the distance of the listening area will depend on the separation of the enclosures.

In Fig. 20-160B, the angle of the enclosures may be varied. The listening area depends on the separation of the enclosures, and their angle with respect to the listening area. Assume that two enclosures are to be placed in a room 12 feet × 15 feet. In this instance, the listening area is on the order of 10 feet in front of the speakers for a separation of approximately 12 feet. For a room of approximately 14 feet × 23 feet the listening area is about 18 feet in front of the enclosures. A guide to the approxi-

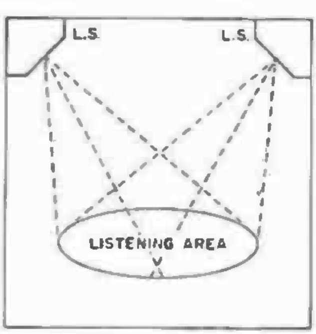


Fig. 20-160A. Loudspeaker position versus the listening area for corner speakers with fixed enclosure angles.

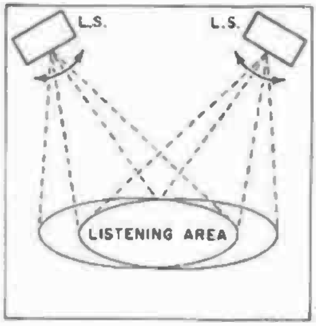


Fig. 20-160B. Loudspeaker position versus the listening area for bass-reflex and similar type enclosures that may have their angles varied.

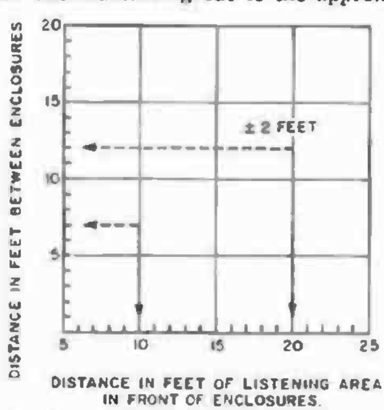


Fig. 20-160C. Chart for determining the approximate separation of loudspeaker enclosures relative to the listening area.

mate separation of the enclosures for a given listening area distance is shown in Fig. 20-160C.

After the initial placement, the speakers should be varied a few degrees one way or the other to attain the proper angles. In some instances, this angle may be quite critical and will require considerable experimentation to find the exact angle. Both speakers must be set to the same angle, and have an unobstructed path to the listening area, since objects in the path of the speakers will affect the high-frequency response. Each room is an individual installation. When the speakers are placed correctly, a person walking across the listening area should not have a feeling that there is a sound void at any point between the enclosures. Before any attempt is made to place the enclosures in their final position, the speaker systems should be phased, a temporary balance for sound level made between the two enclosures, and the stereo-balance control set to its center position. As a final adjustment, both the sound levels and the stereo balance are readjusted, as discussed in Question 20.132.

20.161 What size power amplifier is recommended for stereophonic reproduction?—The subject of how much power should be used to drive a stereophonic reproducing system is a controversial one. While it is true that the average power required is only a few tenths of a watt, to reproduce a record with a wide dynamic range with realism requires an amplifier system of considerable power.

It should be remembered that as the signal level on the record increases, the amplifier output level must increase. If the output level does not increase, the amplifier may overload and cause distortion before the full dynamic range of the record has been reached. As an example, for an increase of 20 dB on the record, the power output of the amplifier must increase 10 times. Thus, if the average power output is 0.5 watt, on peaks the power output is 5 watts.

The power of the amplifier system is somewhat dictated by the circumstance of the location. If it is in an apartment where it is always played at a low level, 10 to 20 watts (each side) will suffice. However, in an area where the sound level is of no consequence, and assuming the speaker system is designed for high-level operation, 50, 75, and 100 watts per channel should be considered.

In a system employing 50 or more watts of power, the amplifiers operate at extremely low distortion. When called upon to deliver near its full power output, the program material is not restricted, and the amplifier adds little distortion to the program material. It is the trend to employ 40 to 50 watts minimum for the average systems, and for larger systems, 75 to 100 watts or more per channel. Such systems have no restrictions as to their dynamic range and will reproduce with realism. On an average, the amplifier delivers a few tenths of a watt, but on heavy peaks, the power demand could reach 60 to 100 watts. If the speaker system is of the low-efficiency type, higher power amplifiers are essential.

20.162 Where is the best location for a monophonic loudspeaker in a medium-sized room?—A corner location, if practical, is preferred. For the best reproduction, the distance between the loudspeaker enclosure and the closest facing wall should not be less than one-half wavelength of the lowest frequency to be reproduced, or about 17.5 feet, which corresponds to a half-wavelength of 32 Hz.

Placing the loudspeaker enclosure in a corner helps to load the speaker by increasing the coupling to the air.

20.163 What is the effect of cavity resonance in the mounting hole of a loudspeaker enclosure?—When a loudspeaker is mounted in an enclosure, as shown in Fig. 20-163A, variations in the frequency response may be noted because of reflections and refractions from the circular walls of the mounting hole

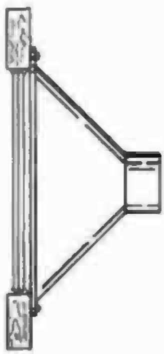


Fig. 20-163A. Cavity resonance caused by reflections from the edges of the loudspeaker mounting hole.

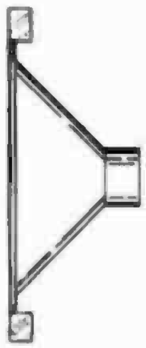


Fig. 20-163B. Cavity resonance eliminated by mounting loudspeaker flush with front of baffle.

This is particularly true if the front wall of the enclosure is over one-half inch in thickness.

Remounting the loudspeaker as shown in Fig. 20-163B, with the mounting ring of the loudspeaker mechanism flush with the front surface of the front wall, improves the frequency-response characteristics, as compared to the conventional method of mounting the loudspeaker on the rear surface of the front wall.

20.164 What is a drone cone?—An undriven loudspeaker cone mounted in a bass-reflex enclosure. Its purpose is to smooth out the frequency response and overcome the deficiencies of the enclosure. Measurements made of enclosures using drone cones indicate that a wider frequency range with greater output is obtained than when using the conventional rectangular port at the bottom of the enclosure.

Drone cones are also used in drive-in theater sound systems. The drone cone is made of a waterproof material and spaced about one-half inch in front of the regular loudspeaker mechanism. The drone cone is activated by air pressure produced by the movement of the regular cone. The sound heard comes from the drone-cone diaphragm. The basic purpose of the drone is to provide a waterproof enclosure for the conventional loudspeaker.

20.165 If two loudspeakers of identical design are close-coupled in a baffle, how does the efficiency vary?—At the low frequencies, the efficiency increases as the square of the area of the cone. Thus, the efficiency is increased four times that of a single radiator. The loudspeaker units should be spaced so that the distance between them is not a multiple or a submultiple of the diaphragm radius.

20.166 What is a distributed-port enclosure? — A loudspeaker enclosure similar to that shown in Fig. 20-166. This enclosure belongs to the reflex class of loudspeaker enclosures and has, instead of the usual port at the lower portion of the front panel, a group of small holes in the front wall. Both the impedance and frequency response are controlled by the addition of a given amount of acoustic resistance.

Interference between front and back radiation has been eliminated because the reflex action has an inherent cutoff of high-frequency back radiation. Good low-frequency response with low harmonic distortion is claimed for this enclosure.

20.167 How is a magnetic headphone constructed? — A cross-sectional view of a typical two-pole, watchcase-type headphone is shown in Fig. 20-167. Basically, it consists of two pole pieces A, energized by a circular permanent magnet B. Around the pole pieces are two small coils C, consisting of several hundred turns of small wire. A thin, soft-iron diaphragm D, is supported at its edges over the pole pieces by a nonmagnetic case E. The edges of the diaphragm are held in place by an ear cap F, which is also made of a nonmagnetic material. When the ear cap is tightly screwed down, an air gap of approximately 0.015 inch exists between the upper ends of the pole pieces and the underside of the diaphragm.

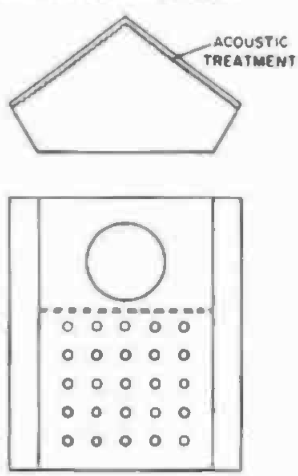


Fig. 20-166. Distributed-port enclosure.

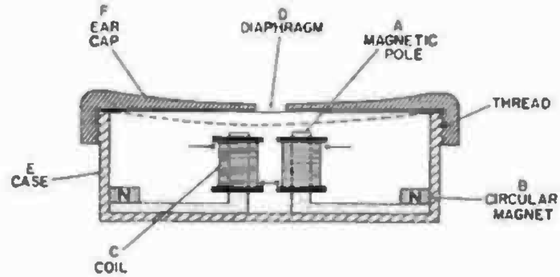


Fig. 20-167. Construction of double-pole magnetic watch-case-type headphone.

Because of the downward pull of the magnetic field on the diaphragm, a slight bias or curvature is caused at the center, as indicated by the dotted lines below the diaphragm.

In use, the two coils are connected in series. When an alternating current is applied to them, the diaphragm will be caused to move either toward or away from the pole pieces, depending on the instantaneous polarity of the magnetic field of the signal. When the signal field is of the same polarity as the permanent magnetic field, the pull is increased, and when it is of opposite polarity, the field strength is decreased. This causes the diaphragm to oscillate back and forth.

Movement of the diaphragm causes alternate compression and rarefraction of the air at the hole in the ear cap, resulting in the generation of sound waves which are passed to the ear of the listener.

Because of the constant downward pull of the magnetic field, the diaphragm does not move an equal amount for each half cycle of the signal voltage. Thus, distortion is introduced. Maximum sensitivity is obtained when the current flows through the coils in a given direction. This direction is generally indicated by the manufacturer by a plus sign. If the correct direction is not indicated, it may be ascertained by reversing the individual phone polarity to obtain the loudest click when a battery is applied to the coil.

When two headphones are used together, they must be phased; that is, the diaphragms must move in a given direction for a given sine.

20.168 How is a crystal headphone constructed?—In a manner similar to the crystal loudspeaker shown in Fig. 20-41A, except on a smaller scale.

20.169 How are crystal headphones coupled to an amplifier?—They should be connected by means of a capacitor and resistor, as shown for the crystal loudspeaker in Fig. 20-41B.

An output transformer may also be used; however, it is rather difficult to obtain such transformers, as the impedance of a crystal headphone is quite high.

An alternate method is to use a 600-ohm output transformer terminated in 600 ohms, with the crystal headphones bridged across the termination. In this type connection, the output level from the phones will be considerably lower, as the phones are fed as a bridging load. The loss in level will be:

$$dB = 20 \operatorname{Log}_{10} \frac{Z}{Z_1}$$

where,

Z is the impedance of the headphones,
Z_1 is the output impedance of the transformer.

20.170 What is the frequency characteristic for a crystal headphone?—As shown in Fig. 20-170.

20.171 What is the average impedance of crystal headphones?—Approximately 45,000 to 80,000 ohms per pair, with the headphones connected in parallel.

20.172 Give the frequency characteristic for a 12.5-ohm dynamic (moving-coil) headphone. — A typical frequency characteristic for a single 12.5-ohm dynamic headphone of the moving-coil type is given in Fig. 20-172A. It will be observed that there is a considerable dip at 4500 Hz and a peak at 7500 Hz. If these deviations are of importance, as they are in most monitoring circuits, an equalizer can be inserted in the circuit feeding the headphones to obtain a more uniform response (Fig. 20-172B).

20.173 How is an equalizer connected for crystal-headphone operation?—As shown in Fig. 20-173.

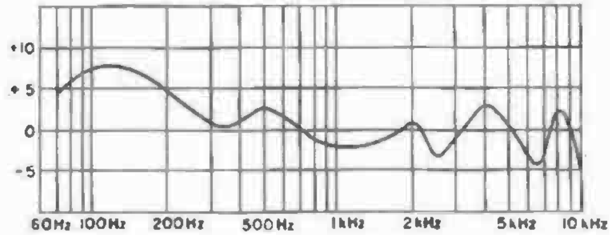


Fig. 20-170. Frequency response of a typical crystal headphone.

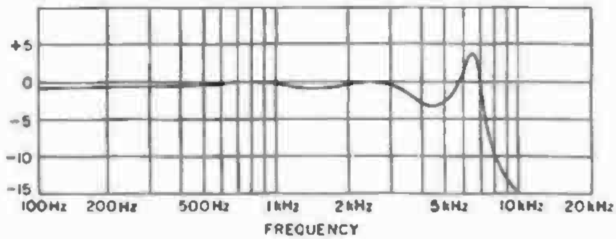


Fig. 20-172A. Frequency characteristic for a 12.5-ohm dynamic (moving-coil) headphone. Note dip at 4500 Hz and peak at 7500 Hz.

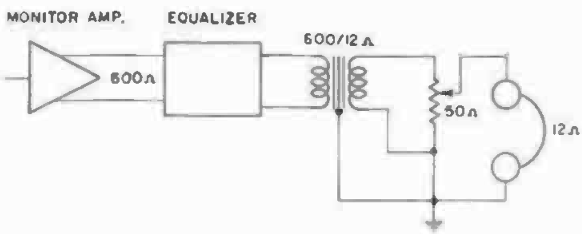


Fig. 20-172B. Headphone monitor circuit with equalizer and volume control.

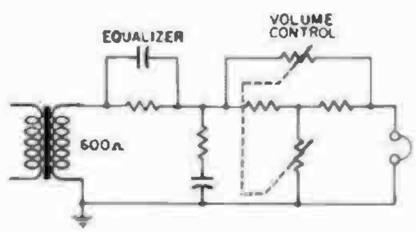


Fig. 20-173. Crystal headphones and equalizer with volume control, operating from a 600-ohm transformer.

20.174 How should magnetic headphones be connected to the plate circuit of an amplifier?—They must always be isolated from the high voltage that feeds the plate circuit. This may be accomplished by the use of a 0.5 to 1.0-μF capacitor in series with the headphones or by using a coupling transformer. Headphones must *never* be connected directly to the plate because the high voltage will damage the headphone windings. If capacitor coupling is employed, a low-leakage capacitor must be used. Suggested circuits are given in Fig. 20-174A and Fig. 20-174B.

20.175 How are headphones rated relative to their impedance?—The impedance is generally rated relative to a frequency of 400 Hz. Headphones of the magnetic type (Fig. 20-167) are rated in dc resistance for one or two headphones in series, which is no indication of the impedance. A typical pair of magnetic headphones rated at 2000 ohms dc resistance will have an impedance of 17,000 ohms at 400 Hz. Dynamic headphones designed for aircraft generally have an impedance of 125 to 600 ohms. Monitor headphones used in sound recording have an impedance of 10 to 50 ohms per pair. Dynamic headphones of

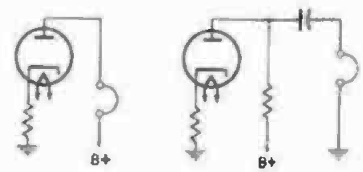


Fig. 20-174A. Capacitor-coupled headphones, isolated from plate circuit.

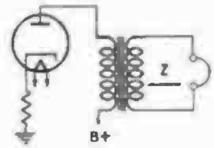


Fig. 20-174B. Transformer-coupled headphones.

the high-impedance type employ a step-up impedance-matching transformer, which is usually located in the headphone case or on the cord.

20.176 Is phasing important in headphones?—Yes, it is qnite important. The two headphones must be connected in such a manner that the diaphragms move in the same direction at any given instant.

20.177 Describe the construction of stethoscope headphones.—A pair of stethoscope headphones manufactured by Armaco of Japan are shown in Fig. 20-177. These phones consist of either a dynamic or crystal driver unit placed at the junction of the two plastic ear tubes. The driver unit generates a sound pressure, which is conducted to the ears through two plastic acoustic tubes. Although these headphones are quite satisfactory for general use, they are not suitable for high-quality monitoring purposes.

20.178 What type headphones are recommended for motion picture sound recording?—Dynamic moviug-coil types, as pictured in Figs. 20-178A and B. Both types of headphones pictured have a total impedance of 25 ohms. The units are connected in series. A typical frequency response for an individual unit is given in Fig. 20-172A.

In the last few years, recordists have turned to the use of hearing-aid-type phones because of their light weight and close contact with the ear. In Fig. 20-178B is shown a pair of such headphones. Each earphone has a plastic insert moulded to fit the recordist's ear contours. The inserts are snapped on the dynamic unit and can be removed for cleaning. In some instances, to present a more uniform frequency characteristic, equalizers are employed to remove the peaks and valleys in the frequency characteristics of the headphones used for high-quality monitoring. The two diaphragms must be phased to move inward (or outward) for a given sign at a given instant.

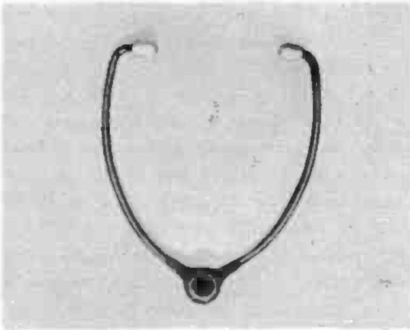


Fig. 20-177. Stethoscope type headphones by Armaco of Japan.

Crystal headphones have also been used for recording purposes but, due to their changing characteristics with temperature and hnmidity, they are not used to any great extent.

Fig. 20-178A. Permoflux Co., dynamic monitoring headphones, Model DHS-28.

20.179 Describe the characteristics of stereophonic headphones.—Stereophonic headphones are generally of the dynamic type and of the highest quality. Each headphone is terminated in a tip, ring, and sleeve-type plug (see Question 24.5) and is carefully polarized to assure that the diaphragms move in the same direction at a given instant. The frequency range is on the order of 30 to 15,000 Hz, with about 1 percent THD for sound power levels up to 140 dB. The impedance of stereo headphones is such that they may be used with amplifiers having output impedances ranging

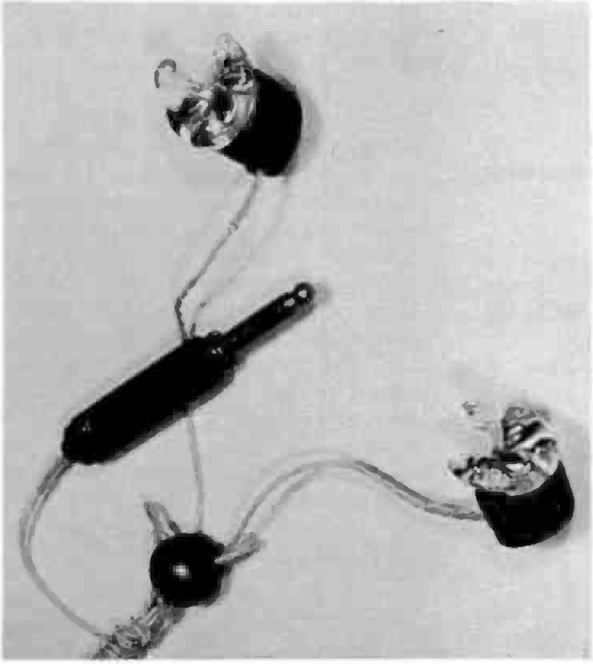


Fig. 20-178B. Hearing-aid type monitoring headphones, used for motion picture sound recording.

from 4 to 16 ohms. A pair of such headphones is shown in Fig. 20-179A.

A typical control box for using two sets of headphones is showu in Fig. 20-179B. The controls P1 and P2 are used for balancing the left and right sides of the program material. In connecting the headphones to the plug, the tip is always the right side of the channel and the ring the left channel, while the sleeve is the common or ground connection.

Fig. 20-179A. Dynamic stereophonic headphones manufactured by Koss Electronics, Rek-O-Kut Div.

20.180 What are language laboratory headphones?—Language laboratories are installations used for teaching of foreign languages. Headphones and microphones used for this work must have good frequency response, low distortion, and a high rate of intelligibility. Such a pair of headphones and a noise canceling lip microphone, manufactured by Telex Acoustic Products, are pictured in Fig. 20-180. The headset is equipped with foam-filled vinyl cushions for comfort, but is designed to attenuate the ambient noise level to a negligible amount. The headphones have a frequency response within plus or minus 2 dB, from 20 to 6000 Hz, with a somewhat greater variation up to 10,000 Hz. The microphone covers a range of 100 Hz to 8000 Hz, plus or minus 5 dB. The impedance of the headphones is 16 ohms. The impedance of the microphone is 50 ohms, with an effective rejection factor of 13 dB. (See Question 4.120.)

20.181 Show a simple resistive network for connecting stereophonic headphones to a system not involving loudspeakers.—The network in Fig. 20-181 can be used with any type reproducing equipment of stereophonic design where the system does not involve loudspeakers. The jack is designed for use with a tip, ring, and sleeve-type plug.

20.182 Describe a stereophonic headphone control center.—A stereo-

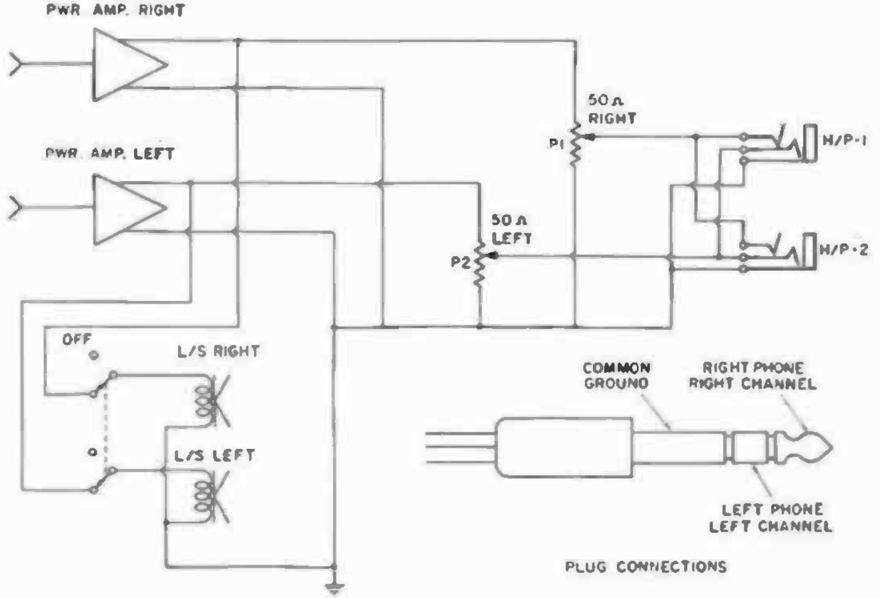


Fig. 20-179B. Schematic diagram for stereophonic headphone connection.

phonic headphone control center provides for several different modes of headphone operation, plus turning off of loudspeakers. A headphone center of interesting design Model CC-1, manufactured by the Jensen Manufacturing Div. of the Muter Co., is pictured in Fig. 20-182A. This unit incorporates a registered trademark control, termed *Space Perspective*.

To understand the operation of the Space-Perspective control and its circuitry, Fig. 20-182B shows the conventional connections for using stereo-

Fig. 20-180. Dynamic headphones and noise-canceling lip microphone for language laboratory use. Manufactured by Telex Acoustic Products.

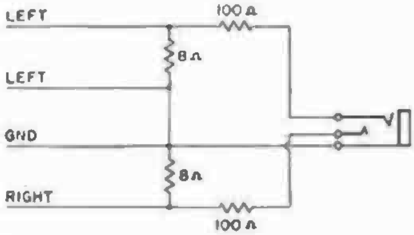


Fig. 20-181. Resistive network for feeding a pair of stereophonic headphones from a circuit not involving loudspeakers for listening.

phonic headphones. Sound from the left channel is presented exclusively to the left ear, and from the right channel to the right ear. Since there are no acoustical sound waves except those generated directly and confined to the individual ears, there is virtually no distance for the sound waves to travel. The sound appears to come from close-up left and right. The effect is as though a partition has been placed through the center of the head, with the orchestra divided on either side. The auditor feels as if he were in the center of the orchestra, rather than sitting in front of the orchestra.

Conventional stereophonic listening using two loudspeakers is shown in Fig. 20-182C. Here the sound waves from the left channel progress to the left ear and also to the right ear, reaching the right ear slightly later in time, due to the distance of travel around the head to reach the right ear. The same effect takes place for the right channel and ear. When the sound is the same on both channels, it appears to emanate from the center, and reaches both ears simultaneously.

B. B. Bauer of CBS Laboratories reasoned that an electrical network could be designed to give the time delay and frequency transmission characteristics of sound waves flowing around the head, and having the same quality as when

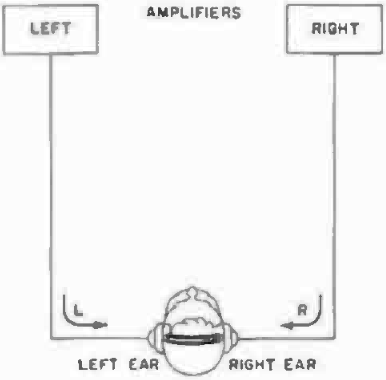


Fig. 20-182B. Listener using conventional headphones.

heard from two loudspeakers. The time delay (about 0.4 millisecond) was calculated, using the dimensions of the average human head and the velocity of sound. However, the frequency characteristics introduced in the natural process of live listening by the build-up and shadowing of the sound pressure at the ears as the sound wave flows around the head presented the greatest problem.

Exhaustive measurements of head diffraction by Weiner were used by

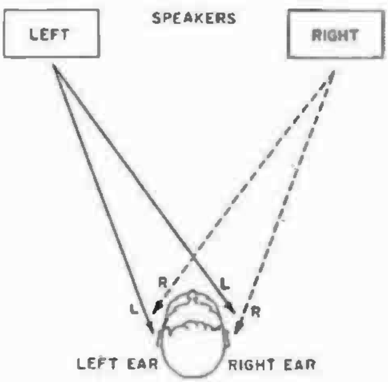


Fig. 20-182C. Listener using conventional loudspeakers.

Fig. 20-182A. Jensen Manufacturing Div. stereophonic headphone center, Model CC-1.

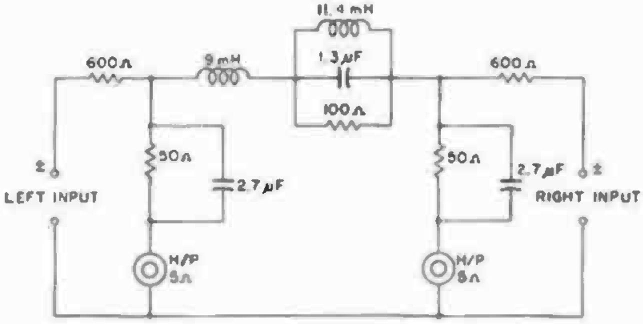


Fig. 20-182D. Original electrical network developed by Bauer for stereophonic headphone listening.

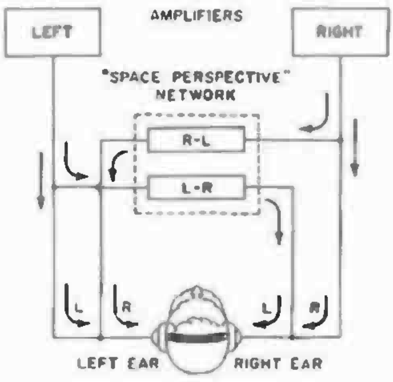


Fig. 20-182E. Current flow in the Bauer network of Fig. 20-182D.

Bauer to develop the required electrical network. In addition to the time delay and the Weiner diffraction data, an additional condition had to be imposed to make the work practical. The network must progressively be eliminated as the sound shifts from the sides to the center, if the panoramic perspective is to be accurately portrayed to the listener. The network of the original Bauer circuit is given in Fig. 20-182D. Its operation is shown in Fig. 20-182E. Observe the similarity of the current flow (indicated by the arrows) to the acoustic sound waves around the head of the auditor.

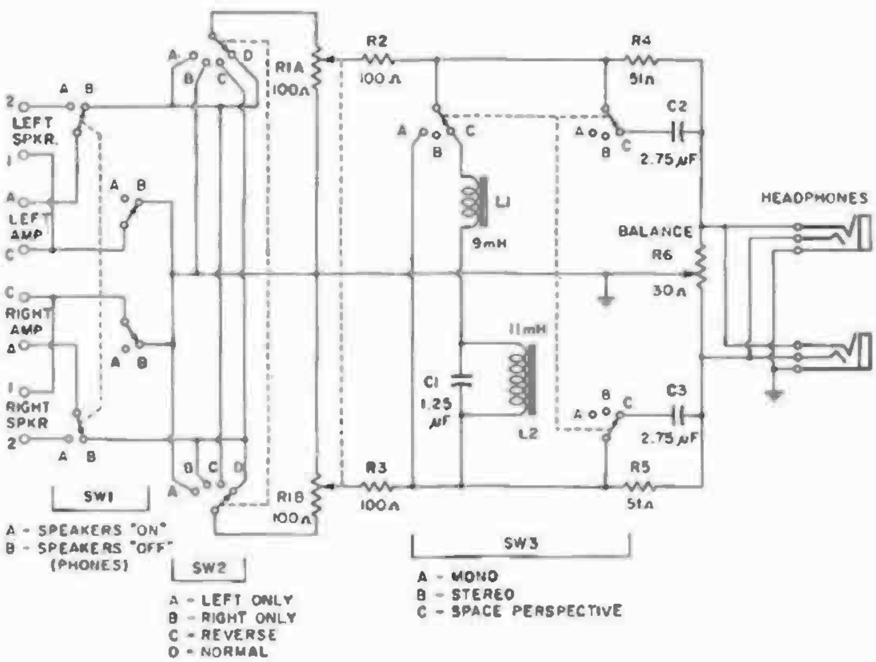


Fig. 20-182F. Schematic diagram for Jensen CC-1 stereo headphone control center.

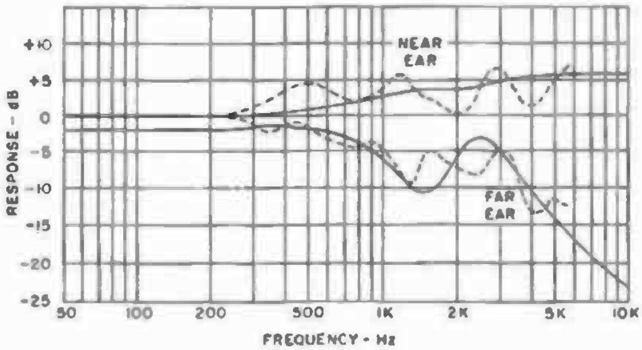


Fig. 20-182G. Comparison of frequency measurements made using the Jensen Space Perspective network, with the Weiner diffraction measurements superimposed (dotted lines).

The circuitry for the Jensen CC-1 stereo headphone control center is given in Fig. 20-182F and was developed from the previous work of Bauer. Measured frequency characteristics of the network and the Weiner diffraction measurements (dotted line) are compared in Fig. 20-182G. The listening effect of this network is that excessive liveness disappears and the sense of spaciousness is almost the same as listening to the loudspeakers as shown in Fig. 20-182C.

20.183 Describe the construction of acoustical suspension-type loudspeaker enclosures (book shelf). — The art of building loudspeaker systems small enough to go into a book shelf has progressed to a point where such loudspeaker systems rival some of the larger systems. Generally speaking, the smaller the enclosure, the less is the low-frequency response. The efficiency of a large diameter speaker unit placed in an improperly designed enclosure is much less than that of a small unit in a properly sized enclosure.

It is often assumed that the diameter of the speaker unit determines the low-frequency response. However, this is not always true because theoretically an 8-inch diameter speaker unit will reproduce low frequencies as well as a 15-inch unit. The diameter does, however, have a direct relationship to the acoustic power produced at a given frequency, since the larger diameter of the radiating surface provides greater coupling to the air and will move more air.

Fig. 20-183A shows a book-shelf system manufactured by Warfedale of England. The enclosure is of the acoustic suspension type (air tight), employing low- and high-frequency speaker units. A special treatment of the interior baffling is used to increase the low-frequency response. Also contained within the enclosure is a 500-Hz crossover network and an LC high-frequency roll-off control, which tapers off the high frequencies starting at 2000 Hz.

The low-frequency speaker is 8 inches in diameter, with high compliance and a low-resonant frequency. The high-frequency unit is isolated from the low-frequency unit by its own enclosure. Tuning slots in the interior baffling complement the low-frequency reso-

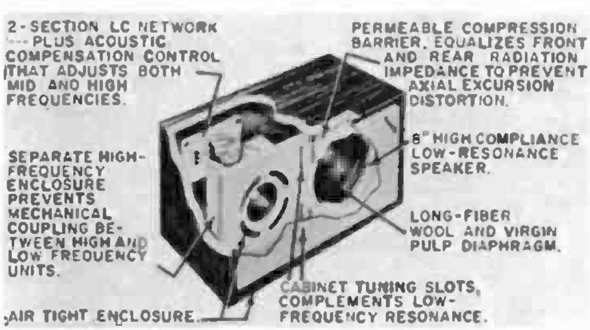


Fig. 20-183A. Interior view of book-shelf type loudspeaker system manufactured by Warfedale of England. (Courtesy, British Industries Corp.)

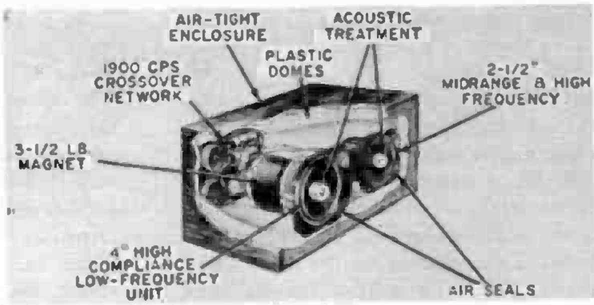


Fig. 20-183B. Interior view of a Goodmans of England Maximus I book-shelf loudspeaker system. (Courtesy, UTC Sound Division)

Fig. 20-183C. Goodmans Maximus 7 three-way book-shelf speaker system. (Courtesy, UTC Sound Division)

nance. The physical size of the enclosure is 19 × 10 × 9½ inches.

A second book-shelf system, Maximus I, manufactured by Goodmans of England, is pictured in Fig. 20-183B. It also is of the acoustic-suspension type, using an air-tight enclosure. The speaker array consists of a 4-inch low-frequency unit employing a 3½-lb magnet, with high compliance and low resonance. The high-frequency unit, approximately 2½ inches in diameter, is back-loaded and covers both the mid- and high-frequency range. Both speaker units are sealed at the front, with the high-frequency unit completely enclosed to isolate it from the low-frequency unit. The enclosure also houses a 1900-Hz LC crossover network and is completely filled with acoustic absorption material. The frequency range is within plus or minus 2.5 dB, from 110 Hz to 15,000 Hz. The low-frequency end drops off to 7 dB in the 60- to 90-Hz range. The efficiency is quite low, so the speaker should be driven with at least a 20-watt amplifier. The enclosure measures 10½ × 7¼ × 5½ inches.

A third enclosure, shown in Fig. 20-183C, is manufactured by UTC-Good-

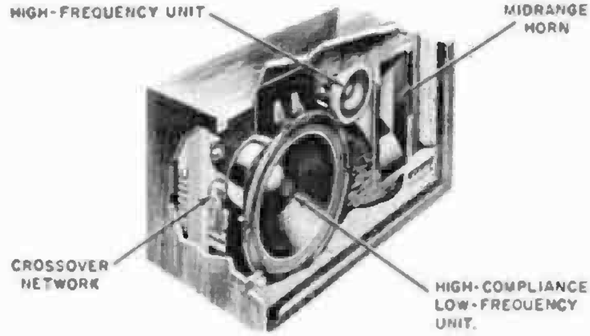


Fig. 20-183D. Electro-Voice E-V Four, book-shelf loudspeaker system.

mans and is known as the Maximus 7 three-way speaker system. The speaker unit consists of a 12-inch low-frequency unit, two midfrequency units with bipolar lenses, and a dome lens multicellular high-frequency unit. The enclosure also contains an LC 1800 and 8000-Hz crossover network. The interior of the enclosure is completely filled with an acoustic absorption material. Two controls on the front panel permit the use to adjust the mid- and high-frequency response to suit the particular acoustic conditions. The frequency response is 45 to 20,000 Hz. This is also a low-efficiency system. The overall dimensions are 24 × 14 × 12 inches.

An interior view of an Electro-Voice E-V Four speaker system is given in Fig. 20-183D. This unit, like most shelf speakers, is an acoustic-suspension type, and it employs a low-frequency dynamic unit, midrange diffraction horn driven by a compression driver, and a 5-inch high-frequency dynamic unit. Crossover frequencies of 800 and 3500 Hz are provided by an LC network, with external controls for adjustiug to meet local conditions. The low-frequency unit employs a ceramic permanent magnet. Frequency response is 30 to 20,000 Hz, with a power capability of 60-watts peak power. The enclosure measures 25 × 14 × 13½ inches.

20.184 What is a series-parallel loudspeaker array? — A series-parallel array consists of many small individual 4- or 6-inch units connected in series-parallel. If the individual impedances of the units are sufficiently high, the units may be connected in parallel. The baffling may be either a flat or a curved assembly. A curved array, designed by Brian Clarke, consisting of 32 individual 6-inch units with an impedance of 100 ohms each and mounted in an open-back baffle is pictured in Fig. 20-184A. The iudividual free-air resonant frequency of the 32 units ranges from 66 to 70 Hz. The 32 units are connected in parallel, thus presenting to the driving amplifier an impedance of 3.12 ohms. It will be noted that each unit is fitted with a high-frequency whizzer. The frequency characteristics (taken out doors) are given in Fig. 20-184B, with the harmonic distortion characteristic shown in Fig. 20-184C. The impedance variation ranges from 3 to 6 ohms.

Fig. 20-184A. Series-parallel array using 32-individual speaker units. (Courtesy, H. F. Sales, Vancouver B.C., Canada)

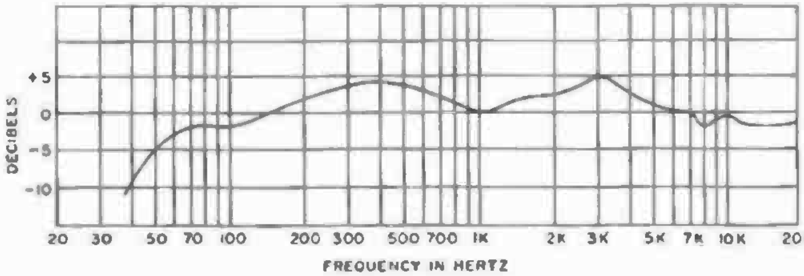


Fig. 20-184B. Frequency characteristics for a 32 unit series-parallel loudspeaker array (after Clarke).

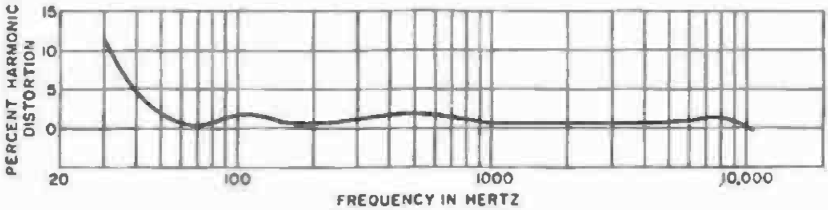


Fig. 20-184C. THD measurement for series-parallel loudspeaker array (after Clarke).

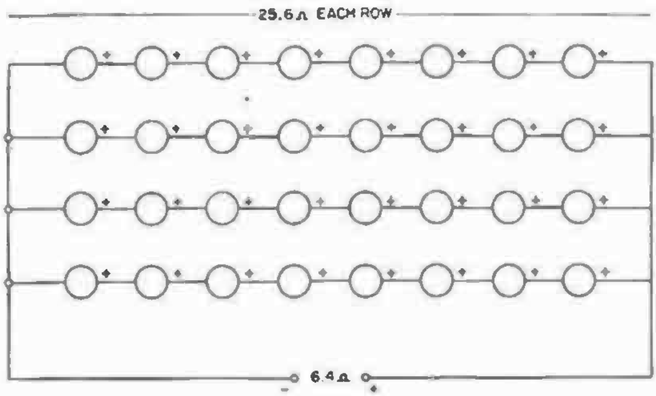


Fig. 20-184D. Series-parallel connection for thirty-two 3.2-ohm speaker units. The plus sign indicates the plus terminal for phasing purposes.

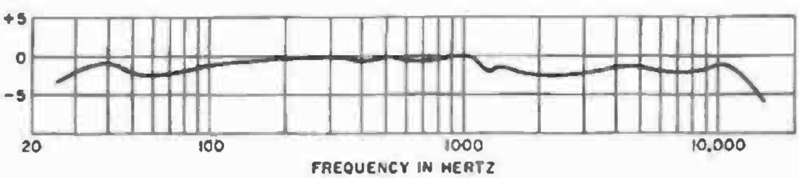


Fig. 20-184E. Frequency characteristics for a 22-unit array.

If speaker units of other impedances are used, they are series-parallel connected to reflect the desired impedance, within plus or minus 15 percent, of the driving output impedance. Fig. 20-184D shows a series-parallel mesh for the 32 3.2-ohm speaker units to provide an approximate load impedance of 6.4 ohms. Generally, an open-back flat baffle arrangement is employed with the speaker units spaced edge-to-edge, with the distance between speakers equal to one-half the radius of the speaker diaphragm. The frequency characteristics for a 22-unit array are given in Fig. 20-184E. It has been found in some instances that additional, separate high-frequency units had to be used to obtain the proper high-frequency response beyond 8000 Hz. In this instance, the crossover frequency selected was around 6000 Hz.

20.185 Describe an in-line or sound column loudspeaker array.—Sound column loudspeaker arrays are unitized speaker systems for commercial sound and public address system applications. The speaker column consists of several direct-radiator speaker units placed one above the other (Fig. 20-185A). The purpose of this design is to confine the sound distribution pattern to a fan-shaped beam, with wide horizontal and narrow vertical coverage. Although this method of mounting speaker units in a column is now used quite extensively, it is not new. It was used in the first RCA Photophone sound motion picture theater installations in 1928. In these installations, the speakers were housed in individual metal baffles, stacked in columns at each side of the screen and focused to the rear of the theater. If the theater had a balcony, a row of speakers was also placed above the screen.

Two general forms of construction have evolved from such experimental work done by individual reseachers and manufacturers of loudspeakers. They are the straight and curved types (Fig. 20-185B and Fig. 20-185C). Several individual speaker units, parallel-connected and in phase, are placed in columns. The speaker diameter may be 6 or 8 inches. The interior of the column is filled with *Fiberglas*. With a wide

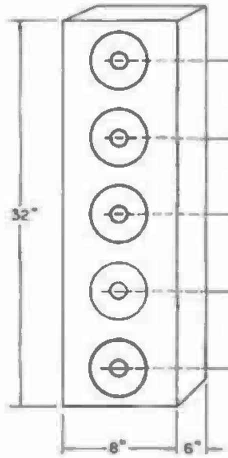


Fig. 20-185A. Simple in-line speaker column. The housing contains five identical speaker units parallel-connected and in phase.

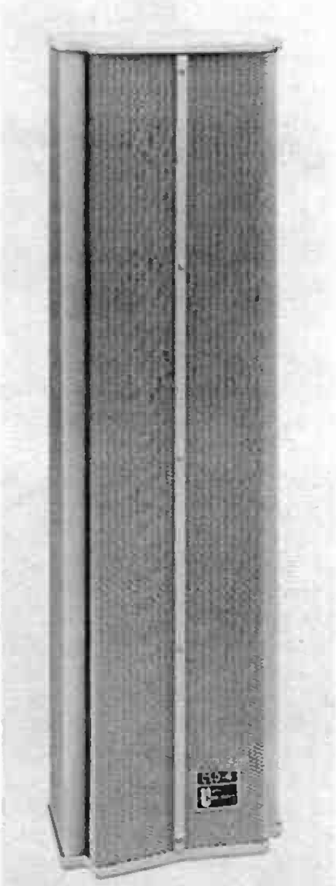


Fig. 20-185B. University Sound Model CSO-4 sound column employing acoustical tapering.

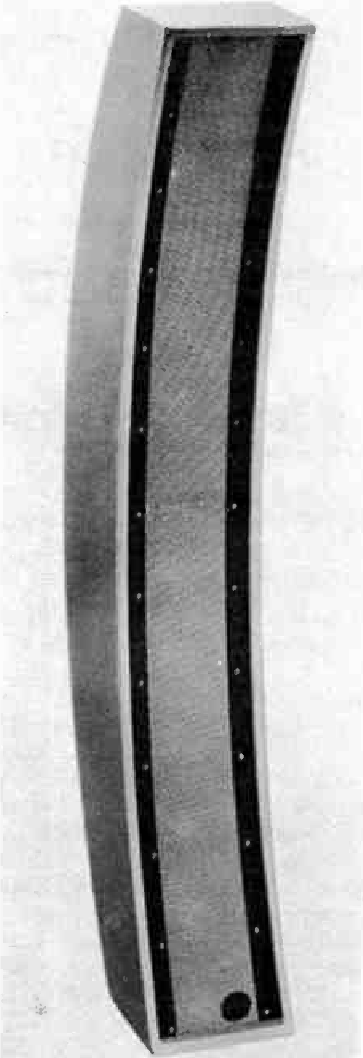


Fig. 20-185C. Electro-Voice Model LR-4A curved-line radiator.

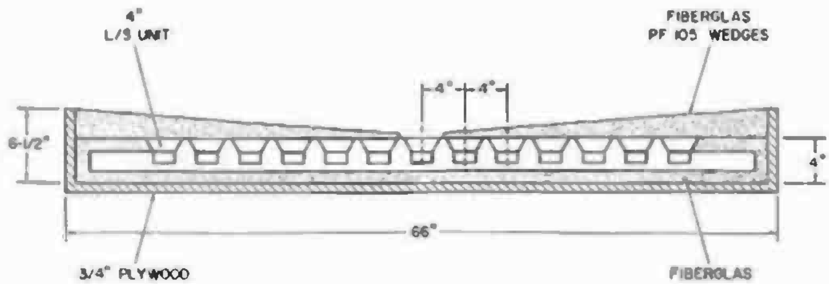


Fig. 20-185D. Cross-sectional view of a sound column employing 13 individual speaker units.

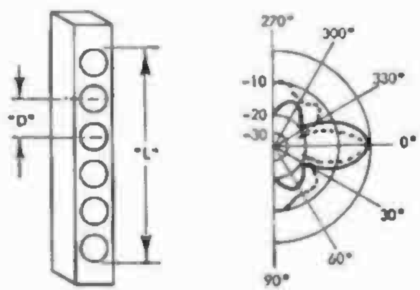


Fig. 20-185E. Vertical polar pattern for a line radiator. The heavy line is a pattern for a given separation of the speaker units. The dotted line is the result of increasing the separation between speakers. (Courtesy, Electro-Voice Inc.)

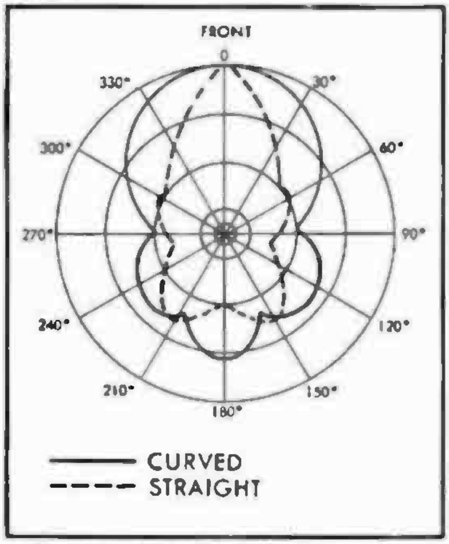


Fig. 20-185F. Vertical polar patterns showing the difference between straight and curved line radiator arrays. (Courtesy, Electro-Voice Inc.)

horizontal sound distribution and about a 30-degree spread in the vertical plane, the efficiency of such arrays approaches that of a horn. The vertical distribution varies with the column length—the

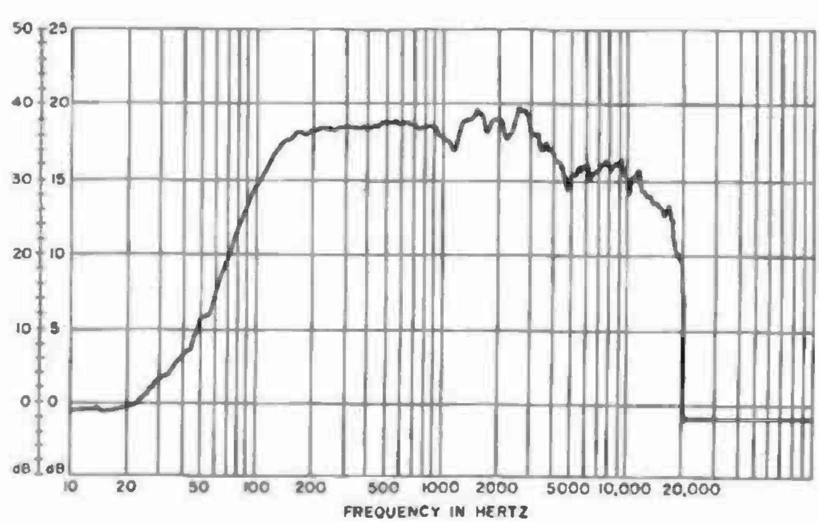


Fig. 20-185G. Frequency response for the sound column shown in Fig. 20-185B. The measurement was made with the microphone 4 feet on axis, in the center of the column. Reference frequency 1000 Hz at 1-watt input.

longer the column the narrower the vertical distribution angle. However, there is a limit to this procedure. When properly installed, the sound quality and coverage is far superior to the conventional-type installation and exhibits a reduction in reverberation effects.

A cross-sectional view of a straight in-line array is given in Fig. 20-185D. Basically, this particular enclosure consists of a heavy plywood case containing 13 individual 4-inch speaker units. To control the polar pattern, acoustical tapering is employed. This is accomplished by covering the diaphragm surfaces of all the speaker units, except the center one, with a wedge of *Fiberglas*. The inner surfaces of the case are also lined with *Fiberglas*. Although this array employs 13 speaker units, fewer speakers could be used. Tests indicate that type PF-105 *Fiberglas* is satisfactory for use with these devices.

Sound column speaker arrays are also constructed electrically by using several different methods for controlling the frequency response and the polar pattern. One method is to employ a combination of single-diaphragm speaker units in combination with dual-diaphragm speakers. In this instance, the dual-diaphragm speakers are placed in the center of the column and the single diaphragm units out to the ends. Another method uses a frequency-discriminating network to control the response, and yet another employs several large single-diaphragm speakers with several small high-frequency units placed near the center, but alongside the larger units.

A vertical polar pattern, for a typical straight-type array is given in Fig. 20-185E. It will be observed from the plot that as the physical separation between the speakers is increased, additional lobes appear and the principal lobe becomes more pointed. The polar plots in Fig. 20-185F are indicative of how the vertical polar pattern changes for a curved (solid line) and straight (dashed line) array. The frequency response for the University Sound column in Fig. 20-185B appears in Fig. 20-185G.

20.186 Describe a stereophonic speaker system using slot-loaded high-frequency speakers. — Two Model P-4000-P stereophonic loudspeaker systems, manufactured by R. T. Bozak, are illustrated in Fig. 20-186. It will be observed that a vertical column of eight high-frequency units are mounted behind the supporting baffle plate. The manufacturer claims that by mounting the high-frequency speakers in this manner, the dispersion of the high frequencies is improved. The dispersion at 5000 Hz is 150 degrees, and at 10,000 Hz it is 120 degrees, plus or minus 5 dB.

20.187 Give the plan view for a stereophonic speaker system using curved reflectors.—A plan view for a stereophonic reflector-type loudspeaker system is given in Fig. 20-187. The reflectors are made from two sheets of ¼-inch, 4 × 8 *Masonite* and strongly braced to prevent vibration. The curve is such that the speaker position can be adjusted to spread the sound completely over the listening area. The use of the center speaker is not absolutely necessary. The reflective panels are curved inward, more at the center than at the edges, to project the sound into the

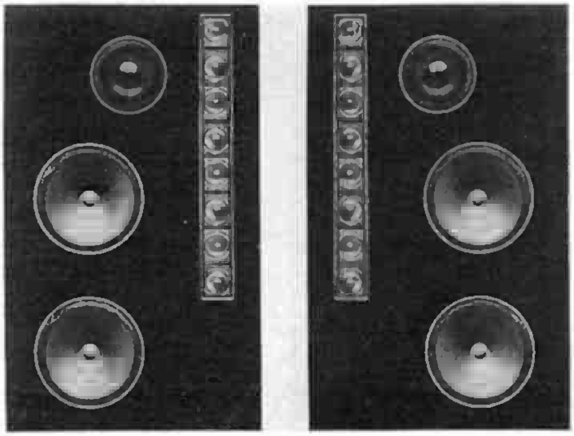


Fig. 20-186. Stereophonic loudspeaker system by R. T. Bozak Mfg. Co. Model P-4000-P.

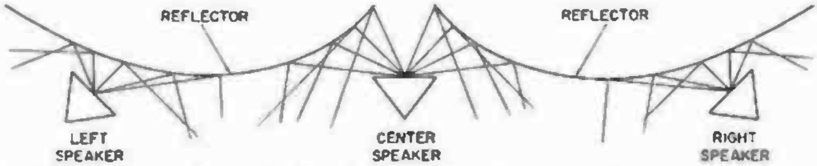


Fig. 20-187. Stereophonic loudspeaker system using two reflectors and a center-derived speaker.

center of the listening area. Phasing and positioning of the three speaker units are essential to its operation.

20.188 What is the purpose of testing loudspeaker systems in a reverberant room?—It has been the practice for many years to measure the characteristics of loudspeakers in an anechoic chamber, because the engineer may then reduce the results to simple mathematical terms. However, anechoic chambers are quite expensive to construct and generally require considerable space and special construction. On the other hand, reverberant chambers are quite easily constructed by using hard nonabsorptive surfaces. The emitted sound energy may go through numerous reflections before it is finally attenuated.

Reverberant chambers give the response characteristics for any direction of radiation. Mean energy measurements (MED) will give an immediate picture of the total energy output, and consequently a direct measurement of its efficiency. Such type measurements are quite useful in investigating the characteristics around the crossover frequency of multiple-loudspeaker systems. Such chambers are also used for many different type measurements not involving loudspeakers. Anechoic and reverberant chambers are discussed in Questions 2.83 and 2.99. The reader is referred to the references.

20.189 Describe a hearing-aid amplifier using an integrated circuit.—With the development of the integrated-circuit elements, the art of hearing aids has developed rapidly. Illustrated in Fig. 20-189A is a complete hearing aid, manufactured by Zenith Hearing-Aid Division, designed for behind-the-ear

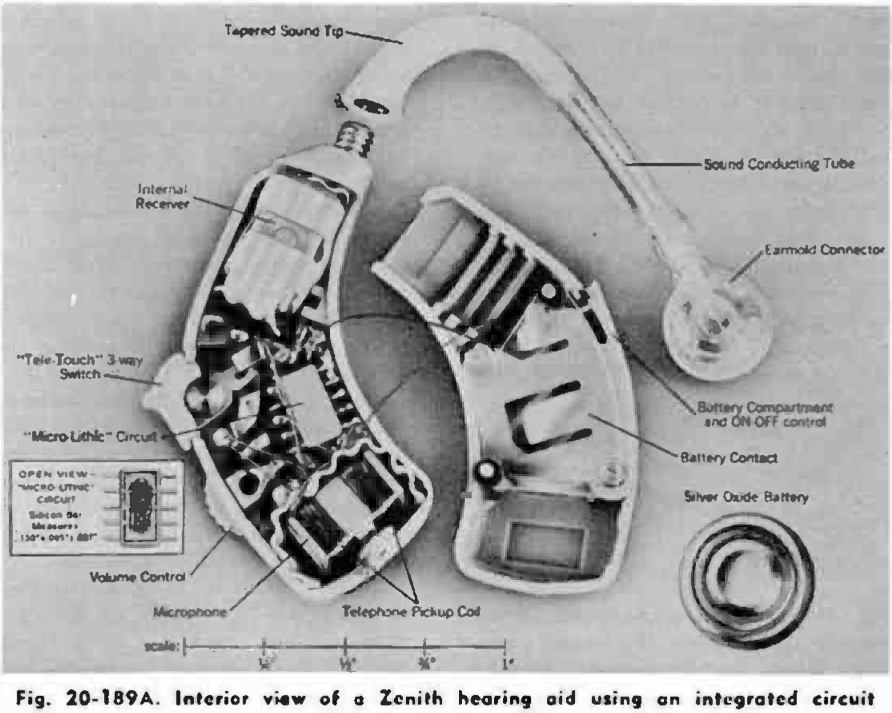


Fig. 20-189A. Interior view of a Zenith hearing aid using an integrated circuit module of very small size.

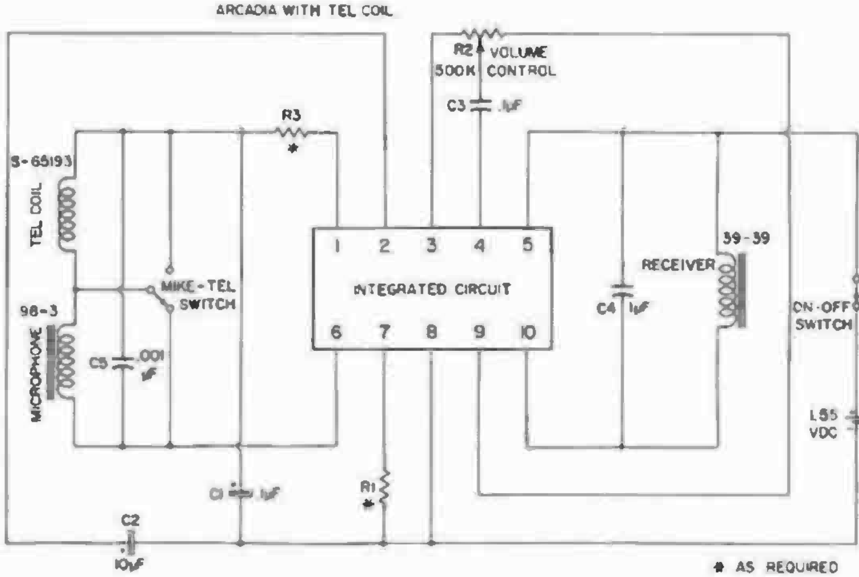


Fig. 20-189B. Internal circuitry for Zenith hearing aid using an integrated circuit module containing the equivalent of six transistors.

use. The various components are called out, with a scale in inches along the bottom for comparison of the component size.

The heart of the device is an integrated circuit which contains the equivalent of six transistors and 16 resistors. The system also includes a microphone and a telephone pick-up coil, either of which may be activated by a switch on the side of the housing. The sound, after amplificatiou, is applied to an internal receiver, and then conducted acoustically down a plastic tube to an earmould connector and into the ear cavity. The battery is a silver-oxide cell, with a life expectancy of 53 hours. The internal circuitry is given in Fig. 20-189B. The complete device may be obtained as a separate behind-the-ear unit, or as a component part of the temple on a pair of eyeglasses (Fig. 20-189C). The

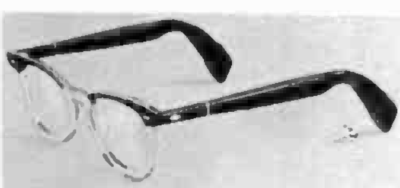


Fig. 20-189C. Eye glasses with Zenith hearing aid built into the temple piece. The volume control and the acoustic tube running to the ear cavity are seen below the temple piece.

frequency characteristic is given in Fig. 20-189D.

20.190 Describe a pyroacoustic loudspeaker.—In 1858, Leon Leconte in England observed that when a flame was subjected to sound waves, it would respond in various manners. This observation led to the development of several experimental devices in which a diaphragm driven by the application of audio frequencies modulates the flame, causing it to act as a loudspeaker with a fairly high efficiency. Sound pressures ranging from 90 to 95 dB have been achieved.

Although the high-frequency respouse is quite good, at the lower frequencies the response falls off at a rate of 6 dB per octave, starting around 2000 Hz. This is the result of poor coupliug to the surrounding air. Iucreasing the area of the flame increases the low-frequency response. Considerable development work on this device has been done by both the Sonic Department of Stanford Research Institute, and the United Technology Center of United Aircraft Corp. At the present time, no practical use has been found for the device.

20.191 Describe a compressed-air loudspeaker.—This type loudspeaker was developed sometime in the early twenties and was used for public address systems and in military airports during World War II. The device con-

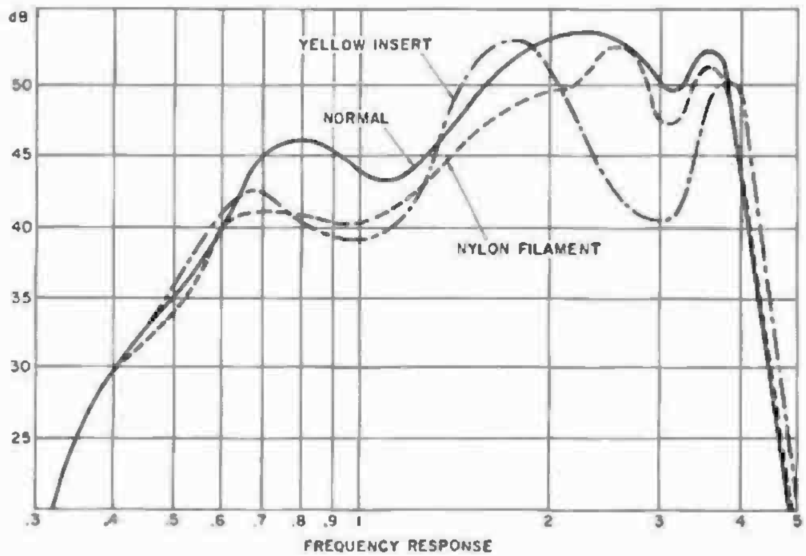


Fig. 20-189D. Frequency characteristics of Zenith hearing aid using an integrated circuit module. The solid line is the normal response.

sists of a value operated by audio frequencies which, in turn, modulates an air stream with approximately 25 lbs per square inch. Modulation of the air stream takes place at the audio frequency rate impressed at the control valve. The valve is coupled to a horn for greater efficiency. The sound power output from the horn is many times that of the power applied to the valve. Sound pressure levels of 120 dB in the mid-high-frequency range can be achieved with good intelligibility. However, the harmonic distortion is rather high compared to a conventional loudspeaker; therefore, careful adjustment of the valve mechanism is required. Generally, several valve and horn assemblies are mounted in a cluster.

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Section 21

Power Supplies

Before the advent of regulated power supplies, 10 to 20 percent in voltage variations were encountered. With regulation, this was reduced to 2 or 3 percent, but the transistor has increased the demand to variations of 1 percent or less. For highly sophisticated equipment, this may be on the order of 0.1 percent.

The regulation of power supplies has a great bearing on the quality of sound recording and reproduction. This is also of prime importance in the testing and measurement of equipment. The ac line voltage must be held within close tolerance of its specified rating (both single and three-phase), and dc ripple voltage must be negligible. This section investigates the following: regulated and unregulated power supplies; voltage doublers, triplers, and quadruplers; constant-current and constant-voltage supplies; transient recovery time; filtering; rectification; and general considerations in the design and usage of power supplies.

21.1 Describe the classification of different type power supplies.—In the early days of radio when equipment was operated from batteries, different nomenclature was assigned to the batteries to identify their positions in the circuit. The batteries were designated "A" for filament supply, "B" for plate supply, and "C" for bias voltage. In equipment using a separate screen supply, the supply was designated "D." When the conversion was made to rectified ac, the designations carried over, and the plate supply was termed "B" supply, and so on for the others.

21.2 Describe an unregulated power supply.—An unregulated power supply has no means of compensating for changes in line voltage or load conditions. The output voltage of the power supply is governed entirely by the line voltage and load conditions at any particular instant. A typical unregulated power supply is shown in Fig. 21-2. Here the high voltage from transformer T1 is rectified by a full-wave rectifier tube V1, and then filtered by means of a two-stage LC-type filter. The filter consists of chokes L1 and L2 in combination with filter capacitors C1, C2, and C3, and terminates in a bleeder resistor R_{BL}. Secondary winding S1 is used to supply heater voltage to other equipment. Two small capacitors of 0.05 μF each are

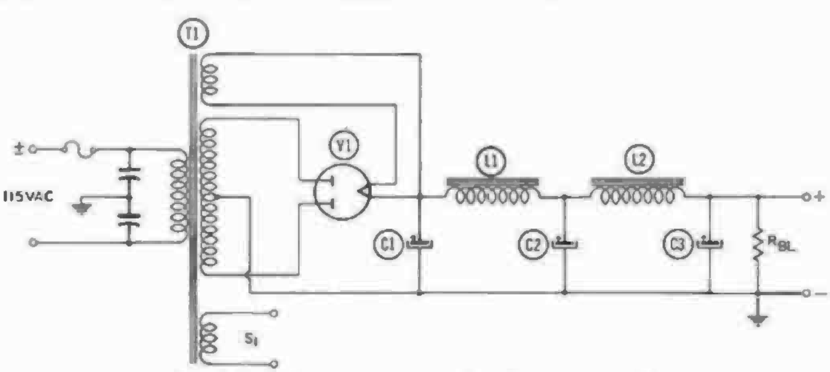


Fig. 21-2. A typical nonregulated power supply.

connected across the ac line to ground for eliminating line noise from entering the power-transformer circuitry.

21.3 Describe the operation of a regulated power supply.—It is a high- or low-voltage power supply in which the output voltage is controlled in such a manner that when the load demand varies, the output voltage of the power supply remains constant or within a small percentage of the rated ouput voltage. Power supplies may also be designed to regulate with respect to the line-voltage variations. A typical regulated power supply, similar to that used for operating recording equipment or any other devices requiring a constant source of voltage, is shown in Fig. 21-3. At the left is a conventional power transformer, with two diode rectifiers, connected for full-wave rectification, and a single-stage LC filter section using a capacitor input. Two low-voltage secondaries supply the heaters of the four pass-tubes (current regulators) V3 through V6, and the heaters of tubes V1 and V2.

The value of the output voltage is adjusted by the 50,000-ohm potentiometer P across the dc output of the supply. This potentiometer is set for a given voltage under load. When variations in the load current occur, the bias voltage applied to the control grid of V1A is changed by the change in the voltage drop across potentiometer P.

Changing the bias on the control grid of V1A controls the amount of current flowing through the common bias resistor R_k in the cathodes of V1A and V1B. As the plate current in V1A changes, the plate current of V1B is also changed, increasing or decreasing the bias at the control grid of V2. The bias voltage of V1B is fixed by the OC3 (VR105) gas regulator tube. Changing the bias voltage on the control grid of V2 causes a change in the plate current through the 520,000-ohm load resistor in the plate circuit; thus, the bias on the control grids of V3 to V6 is changed. Shifting the bias voltage on these tubes permits a greater or lesser amount of current to flow from cathode to plate, thereby satisfying the demand for more or less current at output terminals.

The ripple voltage of a regulated power supply is considerably lower than in a conventional power supply because of this regulating action. To stabilize a regulated power supply, it is customary to connect a large electrolytic capacitor across the output terminals. This helps to lower the internal impedance and prevent oscillations. To prevent oscillation in the current-regulating tubes V3 to V6, resistors of 1000 ohms are connected in each control grid and a 100-ohm resistor is connected in each plate and screen circuit.

To prevent the gas regulator tube V3 from acting as a relaxation oscillator, the value of the capacitor in parallel with this tube should not exceed 0.10 μF.

The capacitor connected at the grid of V1A should not be any larger than is required to remove the ripple voltage

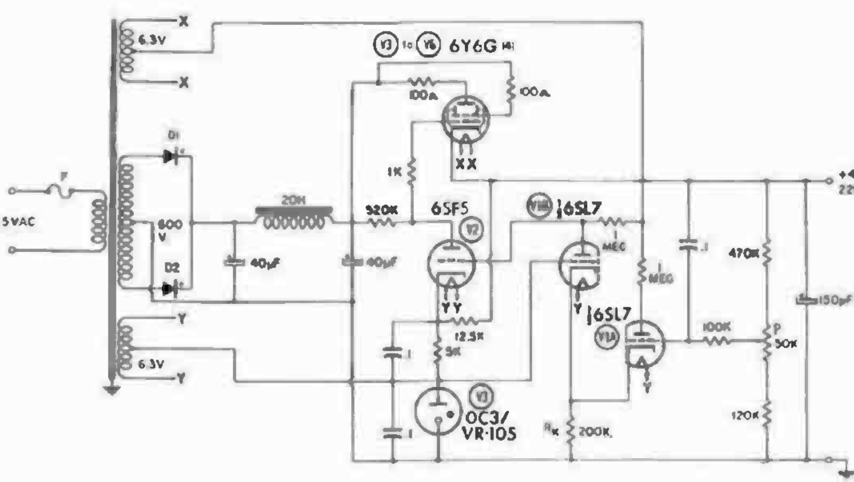


Fig. 21-3. Circuit for a 450-volt, 225-milliampere regulated power supply.

from the grid. If a value greater than 0.25 μF is used, a time lag will be introduced in the regulation time.

The gaseous regulator tube V3 establishes a reference voltage and maintains the cathode of V2 at a constant potential, regardless of the voltage at the output of the rectified system. The stability of the supply is dependent on the constancy of the voltage drop across V3.

The rectifier circuit of a regulated power supply must furnish a higher output voltage than an unregulated power supply because of the voltage drop through the current regulating tubes.

The low output impedance of a regulated power supply makes the output voltage independent of load variations and reduces common coupling of external equipment fed by the power supply.

The power supply shown in Fig. 21-3 is capable of supplying 450 volts at a current of 0.225 ampere, with a ripple voltage of 0.0015 volt, or a noise level of 109.5 dB below the maximum dc output voltage.

21.4 What is a half-wave rectifier? —A rectifier circuit in which only the positive cycles of the applied ac voltage are rectified. The circuit and waveforms for such a rectifier circuit are shown in Fig. 21-4.

21.5 What is a full-wave rectifier? —A rectifier circuit in which both the positive and negative cycles of the applied ac voltage are rectified. The circuit and waveforms for such a rectifier circuit are shown in Fig. 21-5.

21.6 Describe a voltage-doubler circuit.—A voltage-doubler circuit is one in which the rectifiers are connected in series, resulting in twice the voltage output as for a single rectifier. In Fig. 21-6A is shown such a circuit using vacuum tubes, and Fig. 21-6B shows a circuit using solid-state rectifiers. The explanation is the same for both circuits.

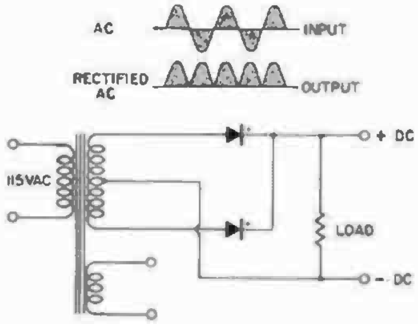


Fig 21-5. Full-wave rectifier circuit using diode rectifiers, showing waveforms at input and output.

Referring to Fig. 21-6A, on the positive peak of the input voltage E_{AC}, capacitor C1 is charged through rectifier V1B to the peak voltage of E_{AC}. The negative half of the cycle charges capacitor C2 through rectifier V1A. The polarities are such that the voltages are additive. Voltage E_{DC} at the output is approximately double the peak voltage of E_{AC}.

Voltage doublers may be used directly from the line or from a power transformer. The circuit shown is a half-wave doubler and delivers 2.82 times the rms value of the secondary voltage.

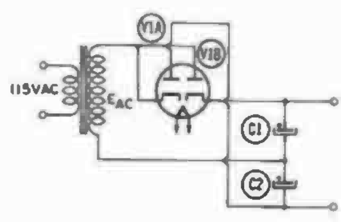


Fig. 21-6A. Voltage-doubler circuit using vacuum tubes.

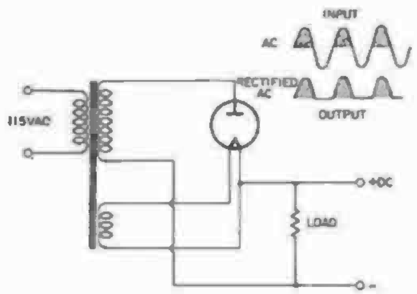


Fig. 21-4. Half-wave rectifier circuit and waveforms at input and output.

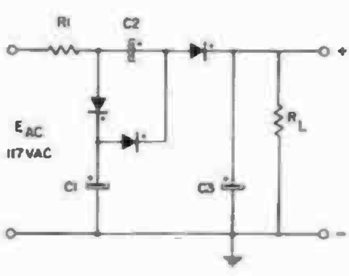


Fig. 21-6B. Half-wave voltage-doubler using solid-state rectifiers.

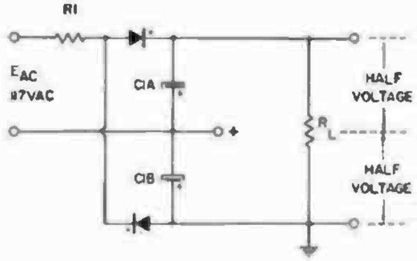


Fig. 21-6C. Full-wave symmetrical voltage-doubler circuit.

Voltage doublers may also be designed for a symmetrical configuration (Fig. 21-6C). The advantages of this circuit are that it is a full-wave circuit, with a lower ripple content, better voltage regulation, and with a ripple voltage double that of the half-wave configuration. It is quite important that the capacitors in the doubler circuit (also for triplers and quadruplers) be close to the same capacitance value, to keep the load evenly divided between the two rectifiers. A half-voltage point is also available by tapping off at the junction of the two capacitors.

Resistor R1 in the solid-state doubler is a current-limiting resistor used to reduce the in-rush current until the load current becomes normal. If a transformer is used, resistor R1 can usually be omitted, as the dc resistance of the transformer acts as a current-limiting resistor.

21.7 Explain the circuitry of a voltage tripler.—Referring to Fig. 21-7, during the first half cycle of the source voltage E_{AC} charges C1 through rectifier D1 to the peak voltage value of E_{AC}. During the other half cycle, the voltage across C1 and E_{AC} are in series-aiding and charge C3 to the same voltage through rectifier D2. The voltage across

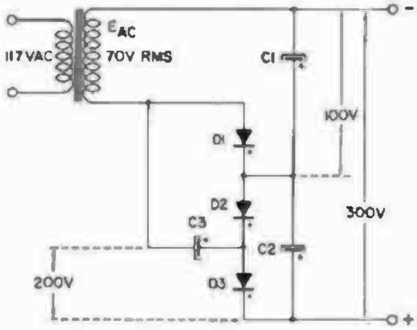


Fig. 21-7. Voltage-tripler circuit.

C3 is now the voltage across C1, plus the peak voltage of E_{AC}.

The voltage across C2 is now brought to the same value as C3, through rectifier D3. The output voltage E_{DC} is approximately three times the peak voltage E_{AC}.

21.8 Describe a voltage quadrupler.—Theoretically, it is possible by adding successive stages of rectification and capacitor combinations to raise the voltage indefinitely; however, from a practical standpoint, this becomes economically unsound. Fig. 21-8A shows a voltage quadrupler using two vacuum tubes, with a similar circuit employing solid-state rectifiers shown in Figs. 21-8B and C.

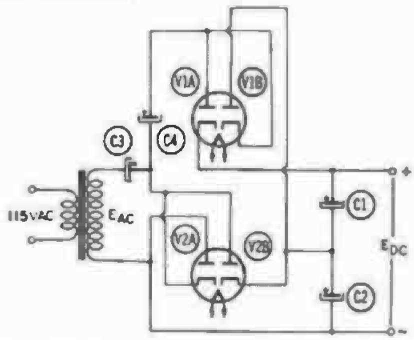


Fig. 21-8A. Voltage-quadrupler circuit. Four-times the output voltage is delivered as for a single rectifier.

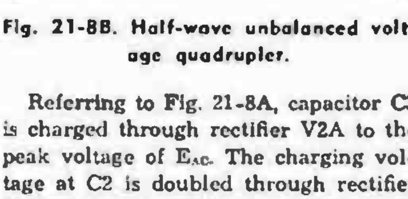


Fig. 21-8B. Half-wave unbalanced voltage quadrupler.

Referring to Fig. 21-8A, capacitor C3 is charged through rectifier V2A to the peak voltage of E_{AC}. The charging voltage at C2 is doubled through rectifier V2B. Capacitor C4 is charged through rectifiers V2B and V1B in series, to twice the voltage. Capacitor C1 is in parallel with capacitor C4 through rectifier V1A, and is charged to twice the voltage. The voltage across C1 and C2 is four times the voltage at E_{AC}. The capacitors have voltage ratings of 450 volts each.

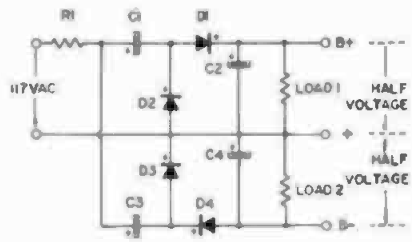


Fig. 21-8C. Half-wave symmetrical voltage quadrupler.

21.9 Describe the various type rectifiers and their uses. — High-vacuum-type rectifier tubes are used in television and radio receivers, radio transmitters, and many types of commercial control equipment. Their power rating may vary from a few milliamperes up to several amperes. These rectifiers have the ability to withstand considerable overload, and as a rule, generate no interference.

Hot-cathode, mercury-vapor-type rectifiers are most commonly used in power supplies involving high voltages and current, such as might be required in radio transmitters. This type rectifier is characterized by its low internal voltage drop and higher efficiency as compared to the high-vacuum type rectifier. However, mercury-vapor-type rectifiers must be preheated if the current requirements are high, and adequate overload protection must be provided for the protection of equipment.

Cold-cathode rectifiers have a rather limited application. When the radio industry was converting from battery-operated sets to ac operated radios, this type tube was often used in a "B" eliminator (plate voltage power supply). This tube is now more or less obsolete.

Mercury-arc rectifiers are used in applications where high voltage and current are necessary. Their size and auxiliary equipment prohibit their use except in very large installations. Such rectifiers may generate considerable interference; therefore, they must be placed in a shielded compartment, and rf chokes must be installed to prevent radiation over the power lines.

Ignition rectifiers are used in high-power installations and are more flexible and economical than the mercury-arc rectifier.

Stacked rectifiers (Fig. 21-74) are one of the most widely used forms of rectifier elements. They are divided into four groups which are: copper-sulfide, copper-oxide, selenium, and silicon. Each type has its own physical characteristics and advantages or disadvantages relative to the economy of operation.

Copper-sulfide or magnesium-copper-sulfide rectifiers are characterized by relatively poor efficiency, ability to withstand high temperature rise, good voltage regulation, favorable weight and size for a given rating, and a good life expectancy. Copper-sulfide rectifiers are used in battery charging and electroplating applications.

Copper-oxide rectifiers are characterized by their large physical size, heavy weight, high efficiency, poor voltage regulation, low temperature rise, and long life. These rectifiers find their greatest use in applications where long life and high efficiency are the most important factors. They are also used for instrument rectifiers and other very small applications.

Selenium rectifiers are characterized by long life, high efficiency, small size and weight for a given rating, and the ability to withstand temporary high current overloads and high temperature rises.

Silicon rectifiers have replaced many of the rectifiers dicussed in the preceding paragraphs. They are small in physical size and require no heater element. They are low in weight and have high-temperature reliability. Rectification efficiency up to 99 percent is possible. Germanium rectifiers are characterized by their high efficiency (98.5 percent) and the absence of ageing. Rectification is accomplished in a single crystal, which does not change by age and storage. Very often, these rectifiers are used in electroplating and anodizing plants. Small rectifiers of this type are also used for instrument rectifiers. They are generally manufactured in stacked array form.

21.10 Define maximum peak plate current.—It is the highest instantaneous plate current a tube can safely carry recurrently in the direction of the normal current flow. The safe value of this peak current in a tube using a hot cathode is a function of electron emission available and the duration of the pulsing current flow from the rectifier tube in each half cycle.

The value of the peak current is largely determined by the constants of the filter sections. With a large choke at the filter input, the peak current is not greater than the load current. However, if a large capacitor is used at the input of the filter section, the peak current may be many times the load current. The current is measured with a peak-indicating meter or oscilloscope.

21.11 Define maximum peak-inverse plate voltage, and how it is calculated.—It is the highest instantaneous plate voltage which the tnbe can withstand recurrently in the direction *opposite* to which it is designed to pass current. For tubes of the mercury-vapor or gas-filled type, it is the safe top value voltage to prevent arc-back in the tube when operating within a specified temperature range. Referring to Fig. 21-11A, when plate A of a full-wave rectifier tube is positive, current flows from A to C, but not from B to C, because B is negative. At the instant plate A is positive, the filament is positive with respect to plate B. The voltage between the positive filament and the negative plate B is in *inverse* relatiou to voltage causing the current flow. The peak value of this voltage is limited by the resistance aud nature of the path between plate B and the filament. The maximum value of voltage between these points, at which there is no danger of breakdown, is termed *maximum peak-inverse voltage.*

The relationship between peak-inverse voltage, rms value of ac input voltage, and dc output voltage depends largely on the individual characteristics of the rectifier circuit. Line surges, or any other transient or waveform distortion, may raise the actual peak voltage to a value higher than that calculated for a sine-wave voltage. Therefore, the actual inverse voltage (and not the calculated value) should be such as not to exceed the rated maximum peak-inverse voltage for a given rectifier tube. A peak-reading meter or oscilloscope is useful in determining the actual peak-inverse voltage.

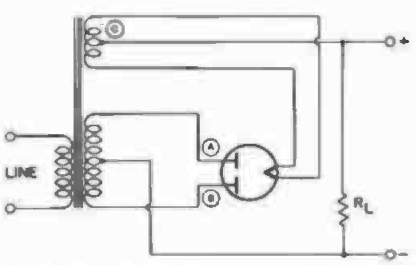


Fig. 21-11A. Full-wave rectifier circuit using a vacuum tube.

For single-phase, full-wave circuits with a sine-wave input and no capacitance at the input of the filter section, the peak-inverse voltage is approximately 1.4 times the rms value of the plate voltage. For a single half-wave circuit, with a capacitor input to the filter section, the peak inverse voltage may reach 2.8 times the rms value of the plate voltage. The same reasoning holds true for either vacuum tubes or semiconductor-type rectifiers.

When designing rectifier circuits employing either tubes or solid-state devices, several factors must be taken into consideration. They are: dc load current; dc load voltage; peak-inverse voltage; maximum ambient temperature; cooling requirements; and overload current. Assume that a full-wave rectifier using silicon diode rectifiers is to be designed (Fig. 21-11B). The dc load voltage E_{DC} under load is 250 volts at 150 milliamperes. The first step is to determine the current, per rectifier, iu terms of a half-wave recitifier.

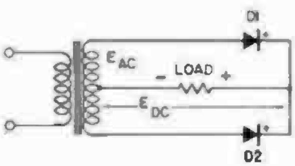


Fig. 21-11B. Full-wave rectier circuit using solid-state rectifiers.

$$I_{eq} = K_2 I_{dc}$$

where,

I_{eq} is the equivalent current of a half-wave rectifier,
K_2 is a constant,
I_{dc} is the rectified ac current.

The value for K_2 is taken from column 5 for a half-wave rectifier (Fig. 21-11C). Inserting this factor into the equation, the current is

$$I_{eq} = 0.5 \times 0.150 = 75 \text{ milliamperes}$$

This is the current the rectifier must carry as a half-wave rectifier. Next, the

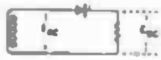
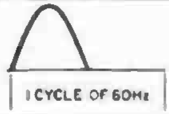
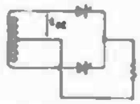
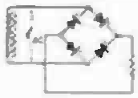
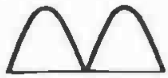
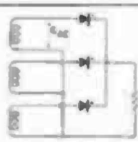
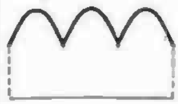
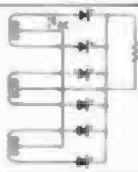
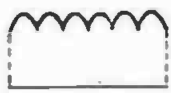
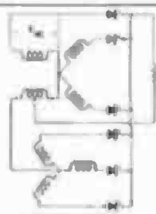
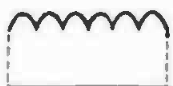
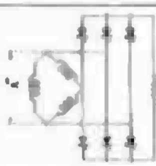
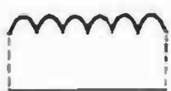
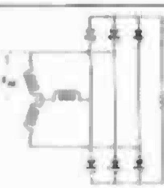
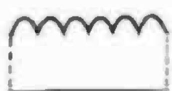
1	2	3	4	5	6
NAME	DIAGRAM	OUTPUT WAVEFORM	K1	K2	K3
1-PHASE HALF WAVE			2.22	1.0	1.414
1-PHASE CENTER-TAP			1.11	0.5	2.828
1-PHASE BRIDGE			1.11	0.5	1.414
3-PHASE HALF WAVE			0.86	0.374	2.45
3-PHASE CENTER-TAP			0.74	0.261	2.828
3-PHASE DOUBLE-WYE			0.86	0.187	2.45
3-PHASE BRIDGE (Δ SEC)			0.74	0.369	1.414
3-PHASE BRIDGE (Y SEC)			0.43	0.369	2.45

Fig. 21-11C. Basic rectifier circuits using a resistive load.

ac voltage required from the transformer is determined:

$$E_{AC} = K_1 E_{DC}$$

where E_{AC} is the transformer voltage, and K_1 is a constant from column 4 for a half-wave rectifier (Fig. 21-11C).

$$E_{AC} = 1.11 \times 250 = 277.5 \text{ volts rms}$$

This is the voltage as measured from each side of the transformer center tap; the total voltage across the secondary is 555 volts rms.

The peak inverse voltage is:

$$PIV = K_3 E_{AC}$$

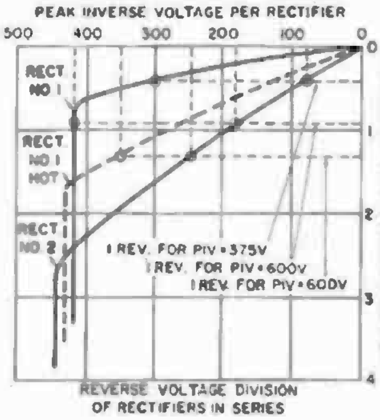


Fig. 21-11D. Reverse-voltage characteristics of two diode rectifiers connected in series.

The value for K_s is taken from column 6 and equals 2.828; therefore,

$$PIV = 2.828 \times 277.5 = 784 \text{ volts rms.}$$

A manufacturer's catalog can be consulted to find a diode rectifier with a PIV rating of 780 volts or greater, at a current of 75 milliamperes. A good selection would be a diode with a PIV rating of 900 volts and capable of carrying 100 milliamperes or more. If the rectifier diode is to carry several amperes, it must be mounted on a heat sink, otherwise it will be severely damaged within the first few seconds of operation. If a diode with the required PIV rating is not available, two or more diodes may be connected in series to obtain the desired PIV rating. An example would be two units each having a PIV rating of 400 volts at 75 milliamperes connected in series. PIV ratings of unequal values may be used, provided the lowest rating is greater than half of the total PIV rating needed.

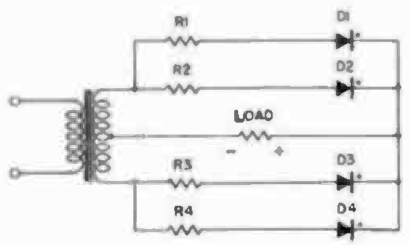


Fig. 21-11E. Parallel operation of diodes. Small resistors are connected in series with each diode to balance the current through each unit.

If the units have different reverse characteristics, the division of inverse voltage between the rectifiers will occur according to their individual reverse current-voltage curves, and the value of constant reverse current will be in relation to the magnitude of the inverse voltage. For example, if two rectifiers are connected in series, and a PIV of 375 volts is applied, the voltage will divide with a ratio of 75 volts to 300 volts, as shown in Fig. 21-11D. By increasing the PIV to 600 volts, the voltage divides to 175 volts and 425 volts, according to the reverse-current flowing and rectifier 1 in an avalanche condition. Rectifier 2 will overheat which, in turn, will increase its reverse current, as shown by the dashed line. The reverse current now shifts, and a new division of the applied inverse voltage of a ratio of 250 volts to 350 volts takes place. Because of the action described above, any number of rectifiers can be connected in series.

Parallel operation of rectifiers is also possible to obtain higher current ratings. However, because of a possible unbalance between the units due to the forward voltage drop and effective series resistance, one unit may carry more current than the other and could conceivably fail. To prevent this, small resistances are connected in series with each individual diode to balance the load currents (Fig. 21-11E). (See Question 21.91.)

21.12 ***What is hot switching current?***—The transient current that flows in a rectifier tube when the equipment is switched off and then on again before the cathode temperature has decreased by an appreciable amount. The transient current is large if the first filter capacitor is in a discharged condition when the supply voltage is reapplied.

If the supply voltage is at peak value when the power is switched on, the largest transient current will flow between the cathode and plate of the rectifier. Under these conditions, a considerable amount of active material may be removed from the cathode; and the emission and life of the tube will be reduced.

21.13 ***What is a choke-input filter system?***—A rectifier circuit which employs a choke at the input of the filter system rather than a capacitor, as shown in Fig. 21-13. Although the out-

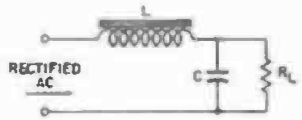


Fig. 21-13. Choke-input filter.

put voltage from this type filter is lower, the voltage regulation is better.

In an LC-type filter section, the inductive reactance of the choke tends to oppose any change in the current flowing through the winding. Therefore, it has a smoothing action on the pulsating current of the rectifier. The capacitor at the output of the choke stores and releases electrical energy, thus also smoothing out the ripple voltage. The result is a fairly smooth output current. Adding a second filter sectiou results in a steady dc current. The choke-type filter has another advantage in that its low dc resistance induces only a small voltage drop across its winding, which becomes quite important at heavy load currents.

21.14 Describe a swinging choke and how it functions.—A swinging choke is used in the first filter section of a power supply having a wide range of load current. The choke is desigued so that its inductance varies inversely with the load current. The core has little or no air gap, which permits it to *saturate* at high current, thns decreasing its inductance. The important points of its construction are the indnctance, the core gap, and the dc resistance.

Because the inductance of the choke varies with the load currents flowing through it, the inductance falls very sharply when the current becomes high enough to saturate the core. Therefore, a point of critical inductance is reached for each change of load current. As the load current changes, it is essential that the choke have a critical inductance at both the minimum and maximum load currents. If not, the reactance of the choke will be such that at some operating points, it has little effect and the rectifier will see only the capacitor at the output of the choke, which now behaves similarly to a capacitor-input filter section. This will cause a rise of voltage, and the regulation becomes quite poor.

To determine the critical inductance, the load resistance must be calculated. Assume that a power supply is to deliver 400 volts at 100 milliamperes, and the dc resistance of the swinging choke is 200 ohms. The load resistance will be:

$$\frac{400}{0.100} + 200 = 4200 \text{ ohms}$$

Assuming that the load current falls to 40 milliamperes, the load resistance then becomes

$$\frac{400}{0.40} + 200 = 10{,}200 \text{ ohms}$$

The critical inductance then becomes approximately:

$$\frac{\text{Load resistance}}{1000}$$

For the above example, the critical inductance at full-load current is 4 henries, and 10 henries at minimum-load current. The optimum Inductance is *twice* the value of the *critical* inductance. Using the above information, a swinging choke of 8 to 20 henries at 100 milliamperes is required.

21.15 Describe a capacitor-input filter section.—It is a power supply employing a capacitor at the input to the first filter section (Fig. 21-15). Such rectifier circuits have a higher output voltage than one using the choke input filter (Fig. 21-13). The higher voltage is due to the *peak* value of the rectifier output voltage appearing across the input filter. As the rectified ac pulses from the rectifier are applied across capacitor C1, the voltage across the capacitor rises nearly as fast as does the pulse. As the rectifier output drops, the voltage across the capacitor does not fall to zero, but gradually diminishes until another pulse from the rectifier is applied to it. It again charges to the peak voltage. The capacitor may be considered as a storage tank, storing up energy to the load between pulses. In a half-wave rectifier, this action occurs 60 times per second, and for a full-wave rectifier, it occurs 120 times per second.

For a single-phase circuit with a sine-wave input and no capacitor across the output, the peak-inverse voltage at the rectifier is 1.414 times the rms value

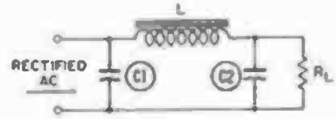


Fig. 21-15. Capacitor-input filter.

of the voltage applied to the rectifier. In single-phase, half-wave circuits with sine-wave input and with a capacitor input to the filter, the peak-inverse voltage may reach 2.8 times the rms value of the applied voltage. This data may be obtained by referring to the tables in Fig. 21-91.

Under certain conditions of usage, advantage is taken of the fact that the value of the input capacitance will control the value of the voltage at the output of the filter section. As a rule, the value of the input capacitor is on the order of 20 to 40 μF. Using a value of from 1 to 8 μF for the input capacitance will permit the output voltage to be adjusted to a given value. However, this type of design affects the regulation and is not recommended unless the load demands are small and constant. A capacitor-input filter section does not have as good a regulation as the choke input, but it does have the advantage of a higher voltage output.

21.16 Describe a resistance-capacitance (RC) filter system.—It is a filter network employing a capacitor and resistor rather than an inductance and capacitor (Fig. 21-16). The advantage of such a filter is its low cost, weight, and reduction of magnetic fields. The disadvantage of such a filter is that the series resistance induces a voltage drop which could be detrimental to the circuit operation. An RC filter system is generally used only where the current demands are low. RC filters are not as efficient as the LC type, and they may require two or more sections to provide sufficient filtering.

A rule-of-thumb design for RC filters is to first determine the value of the series resistance, based on the load current through the resistor. Capacitors are then selected that offer a low impedance at the power supply ripple frequency. Thus, a small power supply may use a 250- to 1000-ohm series resistor, and capacitors of from 40 to 100 μF. For a low-voltage power supply of 30 volts, two filter sections using 1500 μF capacitors could be used. The important point to remember in the use of RC filter sections is that the voltage at the rectifier must be increased sufficiently to compensate for the voltage drop induced by the series resistance, for a given voltage at the output of the filter section.

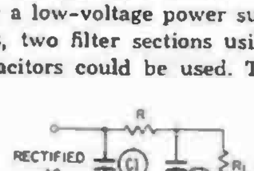


Fig. 21-16. Resistance-capacitance filter.

21.17 How is the voltage rating for an input filter capacitor calculated?—When a conventional dc voltmeter is connected across the unfiltered output of a rectifier tube, the voltage read will be the average voltage. As an example, assume a dc voltmeter is connected across the output of a half-wave rectifier as shown in Fig. 21-17. Because of the inertia of the meter pointer movement, the meter does not respond to the rapidly changing pulses of the half-wave rectified current but acts as a mechanical integrator. The pointer will be displaced an amount proportional to the time average of the applied voltage waveform. If the secondary voltage of the transformer is, say, 350 volts rms, the peak voltage will be:

$$350 \times 1.414 \text{ or } 494.9 \text{ volts.}$$

The average voltage, as read by the dc voltmeter, will be:

$$E_{AV} = \frac{1 \times E_{peak}}{\pi}$$
$$= \frac{1}{3.141} \times 494.9$$
$$= 157.37 \text{ volts.}$$

For a full-wave rectifier circuit, the average voltage at the output of the rectifier will be double that of the half-wave rectifier (assuming each half of the transformer is equal to the voltage of the half-wave transformer secondary) because there are two pulses of rectified current per cycle instead of

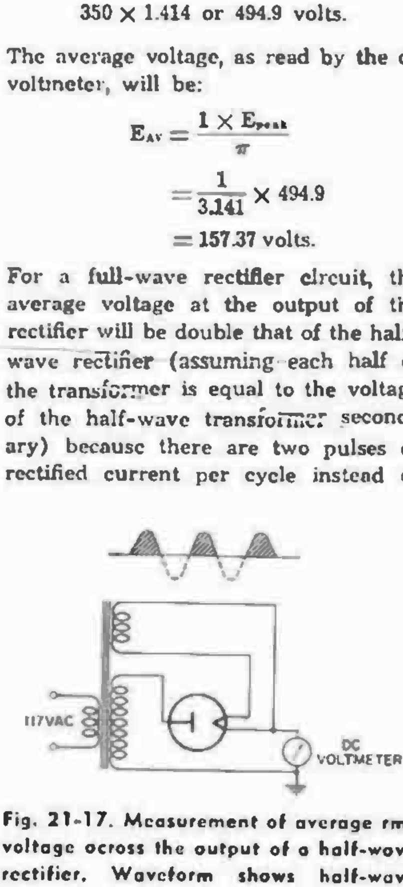


Fig. 21-17. Measurement of average rms voltage across the output of a half-wave rectifier. Waveform shows half-wave pulses of current.

one. This is approximately 90 percent of the rms voltage. Therefore, if the voltage rating of the capacitor is based on a measurement made by a dc voltmeter, it will be in error. The voltage rating must be calculated and determined by mathematically calculating the peak value of the voltage applied to the rectifier tube. The voltage rating of the capacitor must be greater than the peak voltage at the output of the rectifier tube.

In actual practice, a voltage drop takes place internally within the tube. For mercury-vapor tubes and others of similar design, this drop may be only a few volts. For a high-vacuum rectifier tube, the internal resistance is in the order of 75 to 125 ohms. This is taken into consideration when specifying the voltage rating of the capacitor.

If the voltage at the output of the rectifier tube is too high for the conventional electrolytic capacitor, two capacitors may be connected in series. (See Questions 21.46 and 21.67.)

21.18 Can solid-state rectifiers be directly substituted by a vacuum-tube rectifier?—Yes, providing that the rms voltage, current, and PIV ratings are observed. Commercial plug-in units are available to cover most situations.

21.19 What precautions must be taken in the selection of a filter choke?—Filter chokes should be selected for the lowest possible dc resistance commensurate with the value of inductance. To prevent saturating the core, the current rating should be at least 25 percent higher than the maximum current demanded through the choke. If the choke is to be placed near equipment that may be affected by the dc magnetization of the core or ripple voltage, a moderate shield should be included. It may also be necessary to orientate the core in relation to other devices, to reduce the possibility of hum pickup by other components. (See Question 21.20.)

21.20 What effect is noted when a filter choke is overloaded?—When the direct-current rating of a filter choke is exceeded, the core becomes saturated, which reduces the inductance and, in turn, reduces the filtering action of the choke. Under these conditions, the ripple voltage will rise to a value that may render the supply useless.

21.21 What is ripple voltage?—The alternating component (ac) in the dc output voltage of a rectifier-type power supply. The frequency of the ripple voltage will depend on the line frequency and the configuration of the rectifier. The effectiveness of the filter system is a function of the load current and the values of the filter components. (See Question 21.28.)

21.22 How is the ripple voltage of the first capacitor in a full-wave 60-Hz rectifier calculated?—For currents between 10 and 100 milliamperes, with an input capacitor of 1 to 32 μF the ripple voltage may be calculated:

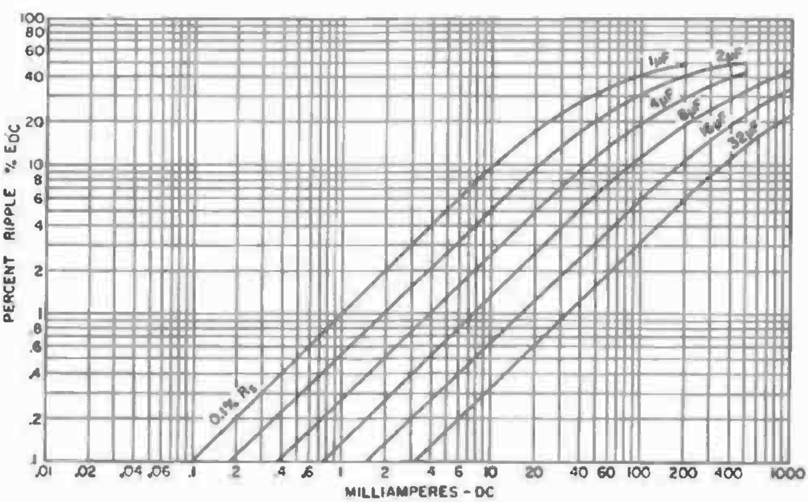


Fig. 21-22A. Percent ripple voltage across the first filter capacitor in a full-wave rectifier using 60 Hz.

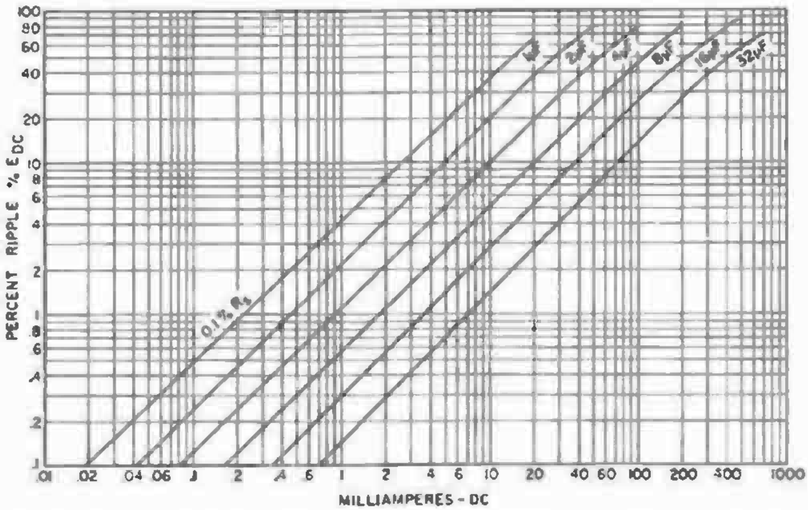


Fig. 21-22B. Percent ripple voltage across the first filter capacitor in a half-wave rectifier using 60 Hz.

$$E_R = \frac{I_{DC}}{C_1}$$

where,

E_R is the ripple voltage,

I_{DC} is the dc current through the rectifier in milliamperes,

C_1 is the value of the input capacitor in microfarads.

For a half-wave rectifier, the formula becomes:

$$E_R = \frac{4I_{DC}}{C_1}$$

with the symbols having the same meanings.

The foregoing formulas will only hold true for ripple percentages up to about 10 percent, for currents of 10 to 100 mA and input capacitors between 1 and 32 μF. Percentage ripple may be read directly, for full- and half-wave rectifiers, from the graphs in Figs. 21-22A and B.

21.23 ***How may the ripple of a dc generator commutator be eliminated?***—By the use of one or more sections of filtering similar to those used in rectifier circuits. The design of the filter sections will depend on the number of commutator bars and the rotational speed of the machine. Filter sections may be designed by the use of the formulas given in Section 7. Noise filters are described in Section 3.

21.24 ***What is the ripple percentage for a 60-Hz full-wave and half-wave rectifier?***—The ripple percentage at the output of a full-wave rectifier is 66 percent, with a ripple frequency of 120 Hz. For a half-wave rectifier, the percentage ripple is 157 percent, with a ripple frequency of 60 Hz. The percentage ripple for a given filter section may be approximated:

$$\% = \frac{100}{LC}$$

where,

L is the inductance of the filter choke in henries,

C is the capacitance in microfarads.

The measurement of power-supply ripple is discussed in Question 23.163.

21.25 ***How is ripple voltage converted to decibels below the maximum dc output voltage?***—By the following formula:

$$dB = 20 \log_{10} \frac{E_{DO}}{E_{AO}}$$

where,

E_{DO} is the maximum dc output voltage under full load,

E_{AO} is the ripple voltage at the output under full load.

A high-quality, unregulated, high-voltage power supply will have a ripple voltage about 80 to 100 dB below the dc voltage. A regulated power supply will have a ripple voltage between 100 and 120 dB below the output voltage. These measurements are made with the rated dc voltage and load current at the output.

Direct-current power supplies used for heater circuits should have a ripple

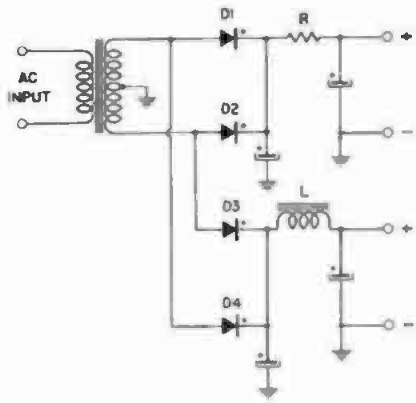


Fig. 21-26A. Combination power supply using a common power transformer for two full-wave rectifier sections.

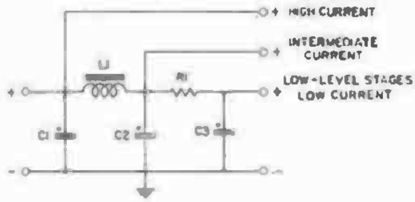


Fig. 21-27. Graded power supply filter. Circuits such as push-pull stage may be taken off at the input to the LC filter section. Lower current stages are taken off at the output of the LC section and the RC section.

voltage 50 to 80 dB below the maximum output voltage.

21.26 What is a combination power supply?—A power supply using a single power transformer serving as a voltage source for more than one rectifier section.

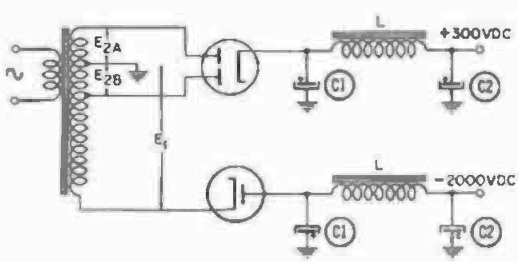
In Fig. 21-26A is shown a single power supply serving two full-wave, high-voltage sections. In Fig. 21-26B, a single power transformer serves a full-wave and a half-wave rectifier. This is a very common type of rectifier circuit and is used mainly in oscilloscopes. It will be noted that the high-voltage output is negative with respect to ground.

In Fig. 21-26C is shown a dual power supply employing a full-wave rectifier circuit using two diodes. Two completely separate filter sections, consisting of an IC network and an RC network are connected at the output of diodes D1 and D2. Filter capacitors C1 through C6 are all returned to a common ground.

21.27 Describe a graded filter system.—A graded filter system generally consists of two or more filter sections, as given in Fig. 21-27. Here, the filter sections consist of an input capacitor C1 and a low-pass filter section L1 and C2, and an RC section composed of R1 and C3. The filter sections shown are typical as found in many medium power vacuum-tube amplifiers operating in conjunction with several intermediate and low-level high-gain stages.

It will be observed that because the push-pull stage draws considerably more current than the other stages, it is connected at the output of the rectifier to eliminate the voltage drop caused by the filter sections. This permits a fairly high inductance to be used in the first filter section.

Fig. 21-26B. Combination power supply as generally found in an oscilloscope. A single transformer supplies both a full- and a half-wave rectifier tube.



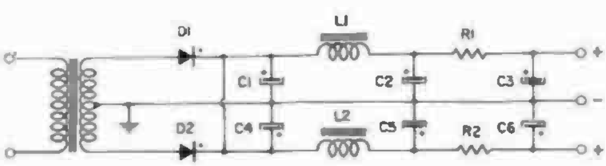


Fig. 21-26C. Dual power supply using two filter sections from one power supply.

Since a push-pull stage is balanced and the inductance of the average push-pull output transformer is on the order of 50 henries, the windings present a high inductive reactance to the power-supply ripple voltages, thus aiding in the filtering of the ripple frequencies.

21.28. *What are the ripple frequencies for single- and three-phase rectifier circuits?*—The frequency of the ripple voltage is dependent on the type and frequency of the line voltage and on whether the rectifier circuit is of half- or full-wave character. For single-phase operation, the ripple frequencies are as follows:

SINGLE-PHASE

Line Freq.	*Full-Wave*	*Half-Wave*
60 Hz	120 Hz	60 Hz
50 Hz	100 Hz	50 Hz
25 Hz	50 Hz	25 Hz

For a three-phase wye-connected circuit, the ripple frequency is 180 Hz. For a bridge circuit, a 6-phase (3-phase diametric) star-connection, and a three-phase double-wye circuit, the ripple frequency is 360 Hz. (See Questions 8.22 and 21.91.)

21.29 *What are the static characteristics of a power supply?*—The voltage characteristics when a constant load is being supplied or when steady-state conditions exist, as shown in Fig. 21-29. It will be noted that the output voltage drops steadily as the load current is increased.

21.30 *What are the dynamic characteristics of a power supply?*—The voltage characteristics that exist when a varying load is being supplied. This measurement is made by connecting an oscilloscope across the dc terminals of the power supply and noting the character of the dc voltage as the load is suddenly applied and removed. Typical

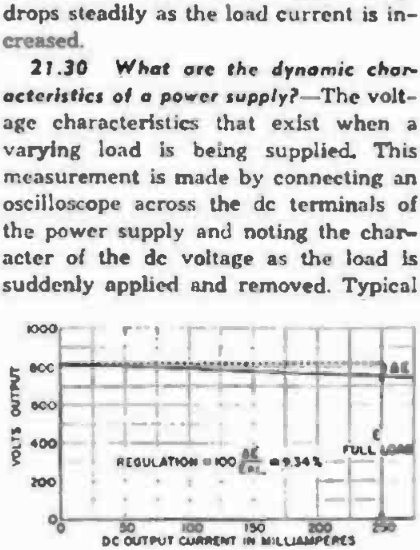


Fig. 21-29. Static characteristics of a nonregulated power supply.

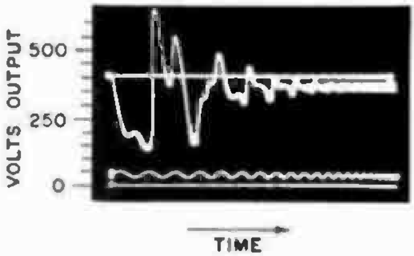


Fig. 21-30. Dynamic characteristics of a nonregulated power supply, showing oscillation caused by variation in load current from a small load to a full load.

characteristics for a power supply with poor regulation are shown in Fig. 21-30.

21.31 *Is it permissible to connect rectifier tubes in parallel?*—Yes, if they have the same characteristics. This is often done in power supplies to increase the output current.

When paralleling mercury-vapor tubes, a resistor of 10 to 50 ohms must be connected in the plate circuit of each tube to balance the current distribution. Two rectifiers connected in parallel double the output current for the same input voltage. The internal voltage drop is halved.

21.32 *Describe a feedthrough filter system.*—A feedthrough filter system generally consists of two conventional RC filter sections, and a third section containing a potentiometer used to null the ripple frequencies by a method of cancellation (Fig. 21-32A). The ripple voltage taken from the rectifier through resistor R3 is out of phase with the voltage appearing at the output of the first two filter sections. By applying the ripple voltage to a third section in the manner shown, a cancellation of the ripple is effected. The value of resistor R3 is approximately 10 times the total series resistance of R1 and R2. The ripple frequency is nulled by connecting an ac vacuum-tube voltmeter across the

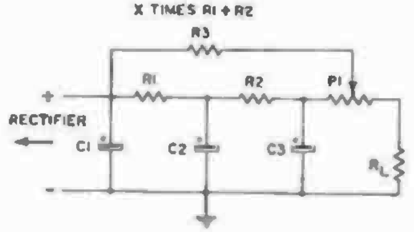


Fig. 21-32A. RC filter sections using a feedthrough resistance to null hum frequencies by cancellation method.

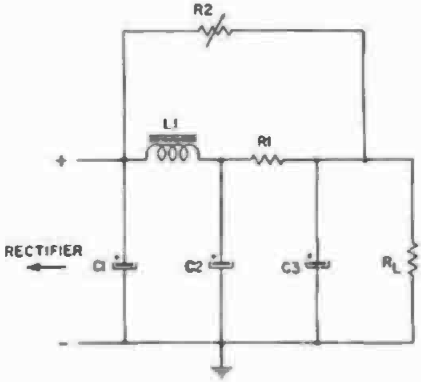


Fig. 21-32B. LC and RC filter sections using a feedthrough resistance to null hum frequencies.

output of the power supply and adjusting potentiometer P1.

This method of cancellation of ripple frequencies can also be used with a combination LC and RC filter systems, as shown in Fig. 21-32B. In this instance, the value of feedthrough resistor R2 is approximately 10 times the total series resistance of L1 and R1.

Cancellation of the higher harmonics of the fundamental ripple frequency is not quite as good in the RC-type filter as for the LC type; however, in equipment where ripple is not too important, the feedthrough method offers a convenient means of reducing the ripple frequency.

21.33 What is a two-section filter? —It is as shown in Fig. 21-33. A capacitor may be connected across the input, if desired.

21.34 What is a combination RC and L filter?—A resistance-coupled filter connected to a choke and capacitor, as shown in Fig. 21-34.

21.35 What is a tuned filter system?—As shown in Fig. 21-35. The capacitor C1 and the choke L1 form a

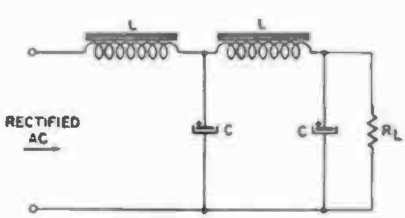


Fig. 21-33. Two-stage filter consisting of L and C with choke input. Input capacitor may be used if desired.

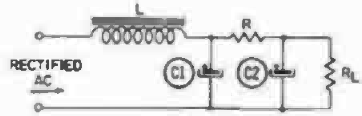


Fig. 21-34. Combination L and RC filter, using choke input.

resonant circuit which is tuned to the second harmonic frequency of the rectifier circuit. For a 60-Hz full-wave rectifier, the resonant frequency is 120 Hz. To eliminate the higher harmonics, a second section should be used as shown.

A second tuned filter is shown in Fig. 21-35B and consists of a series-resonant circuit, L1 and C2, with a parallel-tuned circuit L2-C1 in series with the high potential. The filter sections for a half-wave rectifier are tuned to a frequency of 60 Hz; a full-wave rectifier to 120 Hz.

21.36 What is a bleeder resistance? —A resistance connected across the dc output of a power supply as shown in Fig. 21-36. The bleeder functions to protect the filter capacitors when the

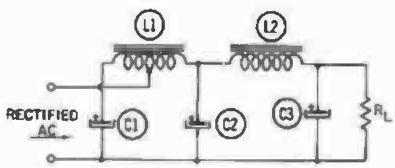


Fig. 21-35A. Tuned or hum-bucking filter L1 followed by a single-section filter L2. For a 60-Hz line supply voltage and a full-wave rectifier, L1 and C1 are resonated to 120 Hz.

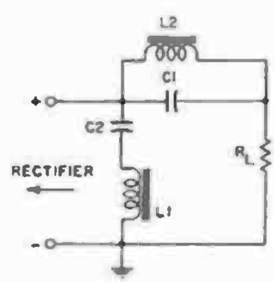


Fig. 21-35B. Tuned filter section using both series- and parallel-tuned circuits.

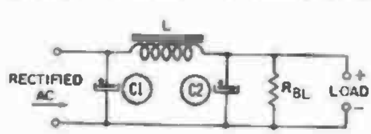


Fig. 21-36. Power supply with bleeder resistance across the output.

load is removed and also to drain off the charge from the filter capacitors when the power supply is shut off. If the bleeder resistor is of the correct value, it will tend to improve the voltage regulation.

The current taken by a bleeder resistor is generally in the order of 10 percent of the total load current. Bleeder resistors are only used with unregulated power supplies.

21.37 What is a voltage-divider resistance?—A resistor or a series of resistors connected across the output of a power supply. The voltage divider provides a means of reducing the voltage at the output of the supply to various values, as showu in Fig. 21-38.

21.38 How are the resistance values for a voltage-divider network calculated?—Two types of voltage dividers are in common use, the shunt and the series types. In Fig. 21-38 is shown a shunt type designed to supply three different voltages to external devices. The upper circuit supplies 75 milliamperes at 400 volts, the second circuit supplies 30 milliamperes at 250 volts, and the third circuit supplies 15 milliamperes at 150 volts. All circuits are common to ground.

The total current required is the total current of the three external circuits, or 120 milliamperes, plus an additional current called the bleeder current. This bleeder current flows only through the resistors and not through the external circuits and is generally 10 percent of the total current. For this illustration, the bleeder current is 12 milliamperes, making a grand total of 132 milliamperes.

Resistor R3 is calculated first. Because only bleeder current flows through this resistor, it may be calculated:

$$\begin{aligned} R3 &= \frac{E}{I} \\ &= \frac{150}{0.012} \\ &= 12{,}500 \text{ ohms.} \end{aligned}$$

where,

E is the voltage across R3,
I is the bleeder current.

The voltage at the top of R2 is 250 volts to ground. Subtracting the voltage drop across R3 results in a voltage across R2 of 100 volts. The current through R2 is the current of load 3 plus the bleeder current, or a total of 27 milliamperes. Therefore:

$$\begin{aligned} R2 &= \frac{E}{I} \\ &= \frac{100}{0.027} \\ &= 3710 \text{ ohms.} \end{aligned}$$

Resistor R1 has the current of loads 2 and 3 plus the bleeder current flowing through it, making a total current of 57 milliamperes. Therefore:

$$\begin{aligned} R1 &= \frac{E}{I} \\ &= \frac{150}{0.057} \\ &= 2630 \text{ ohms.} \end{aligned}$$

The current of load 1 does not flow through any part of the voltage divider system; therefore, it requires no further considerations.

21.39 How is the wattage rating for the voltage-divider resistors in Fig. 21-38 calculated?

$$\text{Watts} = I^2R \text{ or } \frac{E^2}{R}$$

where,

I is the total current through the resistor,

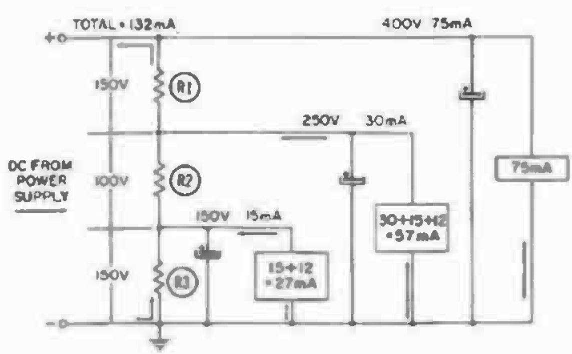


Fig. 21-38. Shunt type voltage-divider system showing the current flow in the various branches.

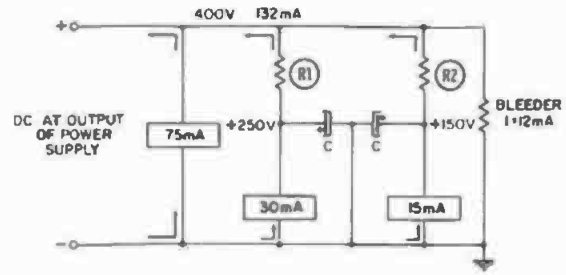


Fig. 21-40. Series type voltage-divider system showing the current flow in the various branches.

R is the value of the resistance in ohms,
E is the voltage drop across the resistor.

As a safety factor and if practicable, the wattage rating of the resistor is doubled over that actually calculated.

21.40 What is a series voltage-divider system?—It is as shown in Fig. 21-40. In this system of voltage division, the resistors are connected in series with the particular load they feed. The resistors are calculated by means of the simple Ohm's law ($R = E/I$). The wattage is computed as described in Question 21.39.

Generally, when a series-resistance voltage divider is used, a separate bleeder resistor is also used to secure better regulation.

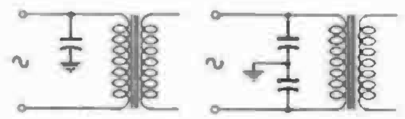
21.41 Should the various sections of a voltage-divider system be bypassed?—Yes, each section should have a separate bypass capacitor of 10 μF or more to ground. The bypass capacitors stabilize and improve the filtering. This is particularly true for the series type voltage divider.

21.42 What type power supply is the easiest to filter, a 60 Hz or a 400 Hz?—The 400 Hz, because the higher the frequency the easier it is to filter.

21.43 What is a transformerless power supply?—This term refers to a power supply that does not use a power transformer. The 115-volt ac power line is connected directly to the rectifier system. This type power supply is dangerous to both operating personnel and to grounded equipment. Also, power supplies of this type will cause hum problems that can only be solved by the use of an isolating transformer between the line and power supply.

21.44 What is the purpose of a static shield in a power transformer?—See Question 8.12.

21.45 What is the purpose of connecting capacitors to ground across the primary winding of a power transformer?—To provide a low-impedance path to ground for line noises and radio signals, thus preventing their entry into the power supply and subsequent transmission to external equipment.



(a) Single capacitor connected from hot side of line to ground. *(b) Capacitors are connected from both sides of the line to ground.*

Fig. 21-45. Noise-filter capacitors connected across the primary of a power transformer to prevent line noises and radio signals from entering the power supply.

As a rule, the capacitors need not be larger than 0.05 μF. Fig. 21-45 shows their proper connection.

21.46 May electrolytic capacitors be connected in series in the filter system of a power supply?—Yes; however, it is good practice to shunt each capacitor with a resistor to equalize the voltage drop across each capacitor. This is necessary, because as electrolytic capacitors age, their internal resistance increases. The resistors must be of equal value if the capacitors are of equal voltage rating.

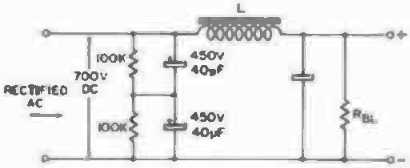


Fig. 21-46. Capacitors connected in series across the output of a rectifier to increase the voltage rating. 100,000-ohm resistors are connected in parallel with the capacitors to equalize the voltage across the capacitors.

In Fig. 21-46, two 40-μF capacitors are shown connected in series across the input of a 700-volt power supply, with each capacitor shunted by a 100,000-ohm resistor. The voltage rating of each capacitor is 450 volts; thus, the two in series have a voltage rating of 900 volts. The voltage drop across each capacitor is 350 volts. The total capacity in the circuit is 20 μF.

21.47 Define the internal impedance of a power supply and its measurements. —The internal output impedance of a power supply is the impedance presented to the equipment receiving the power-supply voltage. In operation of many devices, it is necessary that the internal power-supply impedance be as near to zero as possible. Since most load devices consist of both passive and active elements, the current drawn from the supply consists of an ac component superimposed on the dc output of the supply. This ac component is generally of a nonsinusoidal nature. For the purpose of explanation of how constant the output voltage of a power supply can remain in spite of load variations, it becomes useful to specify the output impedance in ohms over a wide range of frequencies. Power-supply output impedance may be defined:

$$Z_o = \frac{E_{ac}}{I_{ac}}$$

where,

E_{ac} is the sinusoidal voltage across the power-supply terminals as a result of the sinusoidal current I_{ac} flowing through a series loop consisting of the power supply and load equipment.

To measure the output impedance of a power supply at any frequency, it is necessary to draw a sinusoidal current from the power supply and measure the ac component of the voltage which results across the output terminals. Dividing this ac voltage by the ac component of the load current yields the output impedance at a frequency of the sine-wave load. A circuit for the measurement of the ac output impedance is given in Fig. 21-47. A signal current I_{ac} is caused to flow through the output terminals of the power supply and a current-monitoring resistor R1 in series. The output impedance is then:

$$Z_{out} = R_1 \frac{E_{out}}{E_{R1}}$$

where,

E_{out} is the superimposed sinusoidal voltage,
I_{ac} is the sinusoidal current,
R1 is the series resistor.

Several precautions must be taken to assure the accuracy of the measurements. Because the internal impedance of the power supply at the lower frequencies is quite low, an oscillator cannot generally be connected directly to the output terminals of the power supply, as a considerable magnitude of signal level is required. Therefore, an amplifier with a low-impedance output of 2 to 4 ohms and driven by an oscillator is used to supply the signal voltage It is also necessary to insert a blocking capacitor between the output of the amplifier and the power supply to prevent the flow of dc back through the amplifier output transformer. Loading resistor R_L is provided to supply proper dc loading. Since R_L is of greater value than Z_o, it has no effect on the measurement of Z_o.

Since the ac impedance of the power supply is quite low, the output of the amplifier may be mismatched, causing considerable distortion of the sine-wave

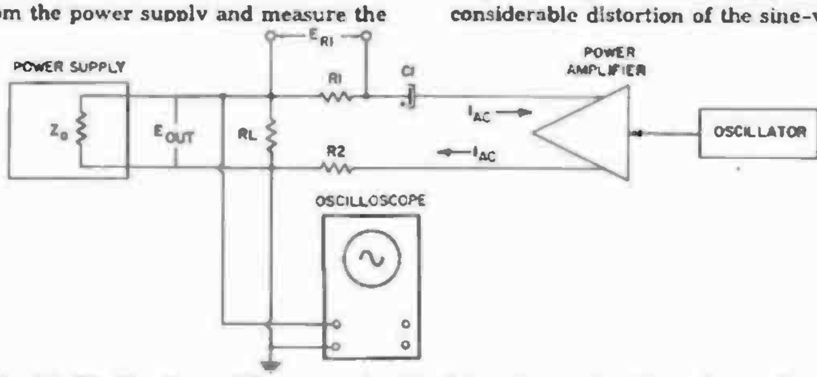


Fig. 21-47. Circuit used for measuring the internal ac output impedance of a dc power supply.

signal waveform. Harmonics of the distorted waveform can cause a larger voltage drop across the power supply, due to the higher impedance at the higher frequencies. Therefore, a matching resistor R2 may be required. At the low frequencies, the output impedance of the power supply is low and the current required to obtain a readable voltage across the output termination may be very large (assuming a power supply impedance of 1 milliohm). To obtain a 1 millivolt signal for the oscilloscope, 1 ampere I_{ac} is required at the output. Therefore, a power amplifier with a low-output impedance is required to supply the sine-wave signal.

Since the ac current required to generate a sufficient signal across the output terminals is rather large, the power supply, if not properly dc loaded, is called upon to accept rather high values of current from the amplifier. However, a power supply is a unilateral device and conducts current in only one direction. Any attempt to force current in the reverse direction results in the output capacitance of the power supply being charged to a voltage higher than the normal output voltage. Also, if too great a peak current is drawn from the supply, the current-limiting circuits may be activated. Therefore, both I_{ac} and I_{dc} must be correctly adjusted before proceeding with the measurements.

The steps for the measurement are as follows: The value of load resistor R_L is so chosen that the current through it at the dc output voltage is equal to or greater than the peak value of the signal current I_{ac}. Also, the maximum instantaneous sum of I_{ac} and I_{dc} should be less than the rating of current-limiting device of the supply. An optimum value is about one-half the rated load current.

R1 should be noninductive over the frequencies of interest, and its hot resistance should be accurately known. Resistor R2 and the impedance of the electrolytic coupling capacitor C1 is the total impedance seen by the output of the amplifier and is equal to the nominal impedance of the amplifier output. The value of C1 must be large enough so that at the lowest frequency of interest, its impedance will be small compared to the output impedance of the amplifier.

It is essential that the voltage drop across resistor R1 be measured at its terminals, so as not to include any voltage drop across its connecting leads. This also holds true for the voltage measured at the output terminals of the power supply. The waveform shapes should be monitored for the presence of undue distortion across resistor R1 and the output of the supply unit. If a large 60-Hz component appears across the output terminals, it is indicative of a ground loop, which must be corrected before completing the measurement. The I_{ac} should be held to a peak-to-peak value of less than the current rating of the supply or to twice the dc current through resistor R1, whichever is smaller. Having satisfied the foregoing conditions, the output impedance may then be stated:

$$Z_{out} = R1 \frac{E_{out}}{E_{R1}}$$

where,

E_{out} is the voltage across resistor R_L,
E_{R1} is the voltage drop across resistor R1. (See Question 21.112.)

21.48 How do the internal impedances of a constant-voltage and constant-current power supply compare?— An ideal constant-voltage power supply would have zero impedance. A constant-current supply would have infinite impedance at all frequencies. However, these ideals are not achieved in a practical power-supply manufacture. The constant-voltage supply has a very low impedance at the lower frequencies, and the impedance rises with frequency. The constant-current supply has a rather high impedance at the lower frequencies and decreases at the higher frequencies. (See Question 21.49.)

21.49 What is the average internal output impedance for a regulated power supply? — For well-designed regulated power supplies, the internal output impedance will range from 0.001 ohm to 32 ohms for frequencies of dc to 1 megahertz. The actual impedance is a function of the load and the type equipment being fed by the supply. However, the impedance of most high-quality supplies ranges from 0.001 ohm to 3 ohms. Such low internal impedauce holds true for both high and low voltage, of either constant-current or constant-voltage design.

21.50 How is the percent of regulation calculated for a power supply?

$$\text{Percent regulation} = \frac{E_L}{E_{NL}} \times 100$$

where,
E_{NL} is the no-load voltage,
E_L is the voltage under full load.

A typical regulation curve for an 80-milliampere regulated power supply is shown in Fig. 21-50.

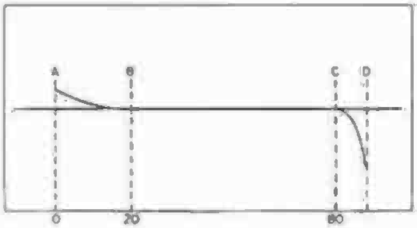


Fig. 21-50. Regulation curve for an 80-milliampere regulated power supply. Point A—no load. Point B—20-milliampere load. Point C—80-milliampere load (maximum). Point D—overload.

21.51 What percent of ripple voltage is permissible for a given type of service?

Radio receivers	0.05%
High-quality amplifiers	0.01% or less
Public-address amplifiers	0.1 to 1.0%
Cathode-ray power supls.	0.5 to 1.0%
Recording equipment	0.001 to 0.01%

21.52 How does a bleeder resistor affect the regulation of an unregulated power supply?—Lowering the value of the bleeder resistance will improve the regulation at the expense of consuming more current. If the supply can assume a rather high bleeder current, better regulation is obtained, up to a point.

21.53 What trouble is indicated if the plates of a rectifier tube glow red?—It is generally an indication of a shorted filter capacitor at the input of the filter system or an excessively high current drain.

21.54 What does a blue glow between the plate and cathode of a rectifier tube indicate?—If the tube is not of the mercury-rectifier family, it indicates an excessive load is being drawn, or the tube is gassy.

21.55 What is a cold-cathode rectifier power supply?—A power supply using a cold-cathode-type rectifier tube. Such tubes require no heater elements. The cathode is caused to emit electrons by bombardment from ions from within the tube, when the high voltage is applied to the plate element. These tubes were generally used in automobile vibrator supplies to reduce battery drain. Tubes of this nature require a starting voltage of about 300 volts per plate and must be operated at a minimum current of 30 milliamperes. A typical tube of this type is the OZ4A.

21.56 What are the advantages of using a mercury-vapor-type rectifier tube?—The mercury tube has a lower internal resistance and therefore, a lower internal voltage drop which is important in high-current power-supply design. Fig. 21-56 shows how the internal voltage drop differs for a vacuum-type rectifier and a mercury rectifier.

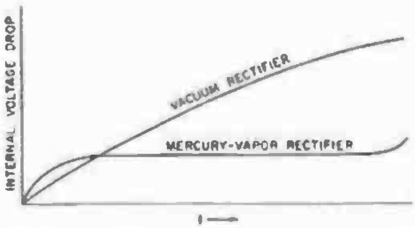


Fig. 21-56. The internal voltage drop of a mercury-vapor tube compared to a vacuum tube.

The internal voltage drop for the vacuum-tube rectifier varies almost in direct proportion to the load, while the internal voltage drop for a mercury tube is practically independent of the load current, up to the point of overload.

21.57 How can interference from mercury-vapor tubes be eliminated?—Mercury-vapor and gas-filled rectifiers occasionally produce a form of local interference in radio equipment through direct radiation. This interference appears as a broadly tunable 120-Hz buzz

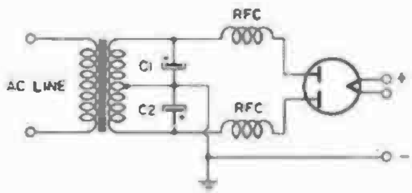


Fig. 21-57. Installation of capacitors across the power transformer secondary, and radio-frequency chokes in the plate circuits to eliminate interference caused by mercury-vapor rectifier tubes.

(100 Hz for a 50-Hz line). This interference is caused by the formation of a steep wavefront at the positive half of each cycle of the ac supply voltage.

Such interference can generally be eliminated by shielding the tube or by connecting a 1-millihenry rf choke in each plate line. The rf chokes must be placed inside the tube shielding. The capacitors must have a voltage rating of approximately 1.414 times the voltage appearing across the entire secondary winding. A typical circuit is shown in Fig. 21-57. (See Question 7.101.)

21.58 Describe a Tungar rectifier power supply.—It is a power supply which uses Tungar rectifier tubes, manufactured by General Electric. The Tungar bulb is basically a diode rectifier, consisting of two elements—a spiral tungsten wire serving as a filament and an anode made in the form of a carbon button. The glass envelope is filled at low pressure with argon gas, which is ionized by electrons from the filameut and becomes the current carrier during rectification. As a result, the internal voltage drop is on the order of 6 to 8 volts. For the 6-ampere bulb, the filament requires 2.2 volts at 18 amperes and an anode potential of 60 volts. The bulb is mounted on a mogul screw-type lamp base. Such rectifiers are generally used for battery charging and power supplies where the current drain is small. One-half-ampere bulbs are also available. Such supplies have now been replaced by solid-state devices.

21.59 What is the effect of an electrolytic filter capacitor having a high-power factor?—Filtering efficiency is reduced and the internal leakage is increased. Electrolytic capacitors should be removed when their power factor reaches an excessive value. In an ideal capacitor, the current would lead the voltage by 90 degrees. However, capacitors are never ideal, as a small amount of leakage current always exists around the dielectric. Also, a certain amount of power is dissipated by the dielectric, the leads, and their connections. All this adds up to power loss. This power loss is termed *phase difference* and is expressed in terms of power factor. The smaller the power factor value, the more effective is the capacitor. Since most service capacitor analyzers indicate these losses directly in terms of power factor, capacitors with large power factors may be readily identified. Generally speaking, when an electrolytic capacitor reaches a power factor of 15 percent, it should be replaced. However, the manufacturer's data sheet should be consulted before replacing certain types of electrolytic capacitors, as they may be designed to operate with a relatively high power factor. Filtering efficiency for different values of power factor can be read directly from the table given below.

Power Factor	*Filtering Efficiency*
5 percent	0.999
10 percent	0.995
15 percent	0.989
20 percent	0.980
25 percent	0.968
30 percent	0.955
35 percent	0.935
40 percent	0.915
45 percent	0.895
50 percent	0.857
60 percent	0.800
70 percent	0.715
80 percent	0.600
90 percent	0.436
100 percent	0.000

21.60 What is a current-limiting resistor and its use?—A resistor connected in series with a rectifier element. It is used to prevent damage by limiting the in-rush current when the rectifier is first energized before the load current becomes normal.

21.61 What is an interlock switch in a power supply?—A switch connected in the power-line leads of a power supply and actuated by the door or cover of the enclosure for the purpose of removing the input power when the enclosure is opened. It is required by Underwriters' Laboratories, Inc., for the safety of operating personnel.

21.62 Describe a high-voltage dynamic loudspeaker field-coil power supply.—A typical high-voltage dynamic loudspeaker field-coil power supply is shown in Fig. 21-62. In this type power supply, no filter system is used, as the field coil and filter capacitor constitnte the filter. Resistor R1 is for adjusting the current through the coil, which is generally on the order of 100 to 150 milliamperes. Low-voltage field-coil supplies are constructed along the same lines, except for the output voltages.

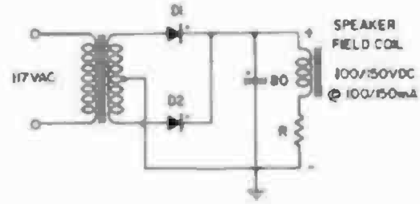


Fig. 21-62. Field-coil power supply for electrodynamic loudspeaker. Power supplies for low-voltage coils are similar except for the output voltage and current demand.

21.63 What are decoupling filter circuits?—They are generally RC circuits connected in the plate supply leads of an amplifier or oscillator circuit, to prevent interstage coupling through the impedance of the power-supply section. Decoupling circuits are discussed in Question 12.136.

21.64 Describe a low-voltage vacuum-tube heater power supply.—In Fig. 21-64 is shown a low-voltage power supply suitable for vacuum-tube heater or filament operation and employing a full-wave bridge circuit, using diode rectifiers.

The filter system consists of a two-section RC filter network using three 1000-μF capacitors. The first resistance, R1, serves as both a voltage-adjusting resistor and as the first section of the RC filter network. The second resistor, R2, also serves as a filter section. In many instances only a single resistor, R1, and a single 1000-μF capacitor, C2, will supply adeqnate filtering.

The heaters or filaments are connected in series across the output, with the lowest-level stages connected at the negative end or ground. This power supply is typical of those used in magnetic recorders to supply the heater circuits of the record and playback amplifiers.

Applying direct current to the heaters in the low-level stages of an amplifier will result in a reduction of hum and noise of at least 20 dB compared to the use of ac on the heaters.

Because of the series connection of the heater circuits, certain precautions must be taken to prevent damage to the heaters, if they are not all of the same current rating. If certain heaters draw less current than others in the string, equalizing resistors must be connected in shunt with these heaters to limit the current through them to its correct value. If this provision is not made, the heater or filament drawing the lowest current will be burned out because of the greater current drawn by the other heaters. (See Question 21.65.)

21.65 How are the shunt resistors for a series heater circuit calculated?—Referring to the circuit in Fig. 21-65, it

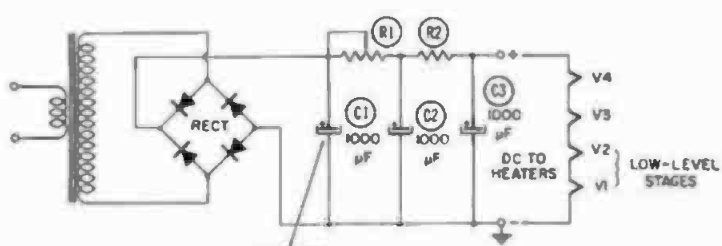


Fig. 21-64. Low-voltage vacuum-tube filament or heater supply circuit, using a full-wave bridge circuit with diode rectifier.

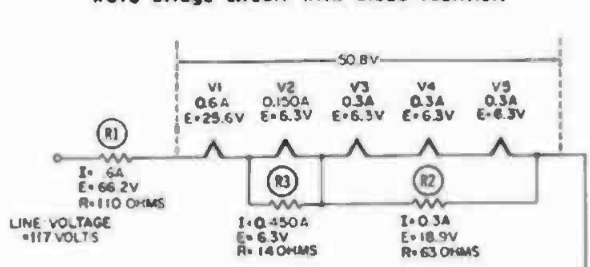


Fig. 21-65. Series-heater operation using shunt resistors across low-current heaters to equalize the current.

will be noted that the five tubes are connected in series. Three tubes draw 300 milliamperes, one draws 600 milliamperes, and the other draws 150 milliamperes of current.

Because the current is not the same for all heaters and they are in series, shunt resistors are required around certain heaters to carry the excessive current. The maximum current for this illustration is 600 milliamperes, the current required for the heater of V1.

The first step is to calculate the line voltage dropping resistor R1. Assuming the line voltage to be 117 volts, the voltage to be dropped across the resistor can be calculated as follows:

$$\begin{aligned} ER &= 117 - [(4 \times 6.3) + 25.6] \\ &= 117 - 50.8 \\ &= 66.2 \text{ volts} \end{aligned}$$

where,

6.3 is the heater voltage of V2, V3, V4, and V5,

25.6 is the heater voltage of V1.

The resistance of R1 is then calculated:

$$\begin{aligned} R1 &= \frac{66.2}{0.6} \\ &= 110 \text{ ohms} \end{aligned}$$

where,

0.6 is the maximum current required.

The wattage dissipated by R1 may be calculated:

$$\begin{aligned} \text{Watts} &= E \times I \\ &= 66.2 \times 0.6 \\ &= 39.7 \text{ watts.} \end{aligned}$$

For maximum safety, a resistor capable of dissipating at least 75 watts should be used. A good rule to follow is to double the wattage rating of the resistor after calculating its wattage dissipation.

The voltage drop across resistor R2 is the same as the drop across the tubes it shunts, or 18.9 volts (3×6.3). The current through R2 is the difference between the maximum current of the circuit (600 milliamperes for V1) and the current required by the tubes to be shunted. The excess 300 milliamperes is carried by resistor R2 around V3, V4, and V5, while R3 shunts 450 milliamperes around V2, which requires only 150 milliamperes for its operation.

Resistor R2 may be calculated:

$$\begin{aligned} R2 &= \frac{18.9}{0.3} \\ &= 63 \text{ ohms.} \end{aligned}$$

The dissipation of resistor R2 is 5.6 watts. Resistor R3 is calculated in a similar manner. To simplify the problem, all voltages and currents are indicated in the diagram of Fig. 21-65. The calculations are the same for either ac or dc operation.

21.66 Describe an electronic ripple filter.—Power supply ripple voltage can also be reduced by electronic means. It is not implied that the electronic method can replace the conventional LC or RC type filter. It is used to supplement conventional filter systems, in order to achieve exceptionally low values of ripple voltage. In using an electronic ripple filter, the ripple voltage applied to the electronic section cannot be more than 3 volts peak-to-peak. If this condition is met, the reduction of ripple voltage obtained with an electronic filter can be 250:1. If the ripple voltage is initially 250 mV peak-to-peak, it can be reduced electronically to a value of 1 mV. Two circuits, developments of the Delco Radio Corp., which are shown in Fig. 21-66A and Fig. 21-66B, have a re-

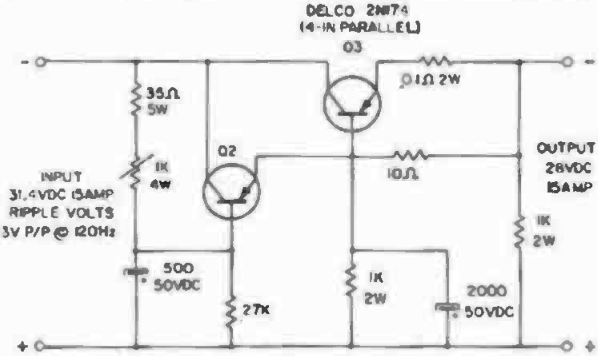


Fig. 21-66A. Electronic ripple filter. Ripple reduction ratio 250:1. (Courtesy, Delco Radio Corp.)

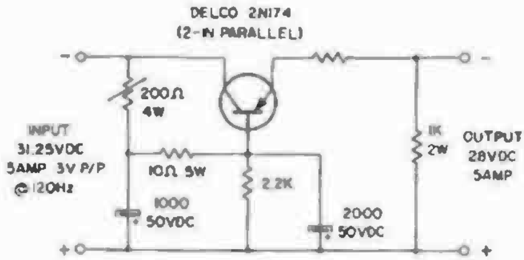


Fig. 21-66B. Electronic ripple filter. Ripple reduction ratio 60:1. (Courtesy, Delco Radio Corp.)

duction ratio of 250:1 and 60:1, respectively.

The basic operation of an electronic ripple filter will now be explained. Fig. 21-66C is the basic circuit, with Fig. 21-66D showing a modified common-collector amplifier which will be compared to the circuit of Fig. 21-66C. The common-collector circuit is inherently degenerative. Increasing its voltage at the input will cause a decrease of output voltage across the load, and vice versa. The ripple voltage is reduced by degeneration. This circuit has at times been erroneously referred to as a capacitance multiplier.

Assuming that the dc level of the input voltage in Fig. 21-66D remains unchanged and that the battery voltage biases the transistor to a V_{ce} of 3 volts, the following action takes place. As the ripple voltage adds to the 31-volt dc input, the voltage across load resistor R_L tries to increase. However, the polarity of the increase is such that it causes the base of Q1 to become less negative with respect to the emitter, thus tending to turn off the transistor. When the ripple voltage subtracts from the 31-volt input voltage, the opposite action takes place.

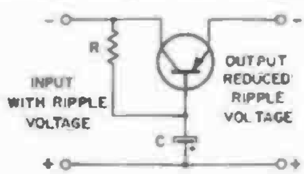


Fig. 21-66C. Basic circuit for electronic filter.

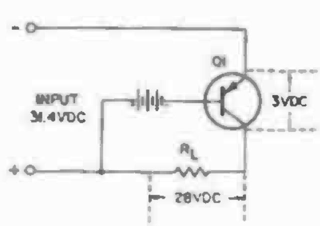


Fig. 21-66D. Modified common-collector amplifier.

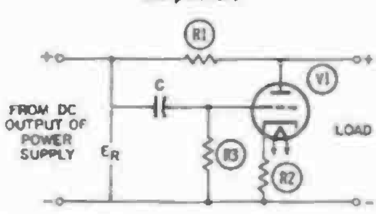


Fig. 21-66E. Vacuum-tube electronic filter.

Referring to Fig. 21-66C, capacitor C tries to maintain a nearly pure dc voltage as the battery would do, except it does have the ability to change the dc level if the input level changes. Capacitor C should be as large as is practical. Resistor R serves to forward-bias the transistor. It may be desirable to insert an inductance in series with resistor R to supply a purer source of dc to the base of the transistor. The circuits in Figs. 21-66A and B are practical circuits and employ two or more pass transistors Q3 connected in parallel to avoid high collector currents. All transistors must be mounted on heat sinks. Resistors are connected in series with the emitters to aid in sharing the load current between transistors. The input voltage may be decreased, if necessary. The temperature range is about minus 65° to 50°C.

In designing such a circuit, pass transistor Q3 should be biased so that the collector to emitter voltage V_{ce} is equal to or greater than,

$$V_{ce} \geq \frac{\text{input pk-pk ripple}}{2} + 1.5 \text{ Vdc}$$

This is done to prevent the pass transistors from going into saturation. With the transistors at saturation, ripple reduction is greatly reduced because a small change in V_{ce} no longer has any control of the collector current. Excess V_{ce} will increase the transistor dissipation, which means additional transistors must be used in the pass section of the circuit. The transistor V_{ceo} rating should be equal to about 1.4 times the supply voltage. This is necessary in order to allow the peak supply voltage to initially appear across the transistors while the filter capacitors charge through the series resistor. Capacitor C in Fig. 21-66C may be replaced with a zener diode, provided the conditions for V_{ce} are met.

A similar circuit, using a vacuum tube is shown in Fig. 21-66E. The total current drawn from the power supply flows through resistor R1. Tube V1 is biased to draw a steady value of plate current. Ripple voltages appearing in the dc output of the power supply will take the path of least resistance to ground, which is through capacitor C and resistor R3. Since the dc component cannot flow through capacitor C, slow changes in the dc output voltage do not activate the tube. However, the ripple component and transient voltages are readily passed through the capacitor to the control grid and cause V1 to function.

If the instantaneous polarity of the ripple voltage is *positive*, this will add to the output voltage and increase its value. The positive increase at the control grid of V1 causes the plate current to increase which, in turn, increases the voltage drop across resistor R1. This action cancels the rise in voltage caused by the ripple voltage. When the polarity of the ripple voltage changes to negative, the action is reversed; in this manner, the output voltage is held constant.

21.67 ***What is the procedure for designing a high-voltage nonregulated power supply?***—The design of a high-voltage power supply requires the same fundamental design considerations, regardless if it employs vacuum tubes or solid-state rectifiers. Before the filtering system can be designed, it is necessary to know the type rectifier circuit to be employed, the load current, and the permissible ripple current in the

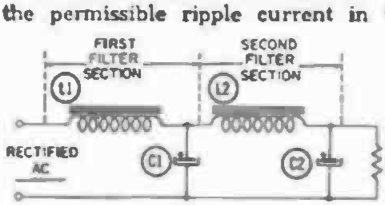


Fig. 21-67. Two-section filter for example in Question 21-67.

output. The number of filter sections required will depend upon this latter information.

Knowing the load voltage and current, it is possible to determine the ac voltage and current that must be delivered by the power transformer for a given type rectifier.

Proper allowance for the internal voltage drop of the rectifier must be made when calculating the amount of ac voltage to be applied to the input in order to obtain the necessary dc voltage at the output of the rectifier for a given load current. If good regulation is essential, the input inductance of the filter system must exceed the critical inductance. The critical inductance for a 60-Hz, full-wave rectifier can be calculated by the following equation:

$$L_c = \frac{R_L}{1000} = \frac{\frac{E_{DC}}{I_{DC}}}{1000}$$

where,

L_c is the critical inductance in henries,
R_L is the dc load resistance,
E_{DC} is the dc output voltage,
I_{DC} is the dc output current.

To prevent the peak current from being excessive, the inductance of the first filter choke must be not less than twice the critical inductance. Thus, the formula is reduced to:

$$L_c = \frac{R_L}{500}$$

The first filter choke must be equal to or greater than this value.

Having determined the minimum value for the first choke L1, the value of the second capacitor C2 can be calculated using conventional low-pass-filter formulas as given in Section 7. The cutoff frequency for a full-wave, 60-Hz rectifier is 120 Hz. Frequencies for other type rectifiers are given in Question 21.28.

The first filter capacitor C1 adds to the filtering but raises the input voltage to the input choke by a factor of 1.414. The second filter section, consisting of L2 and C3, is calculated the same as for the first section. It is important that the filter chokes maintain their inductance values at the required value of direct current that must be passed through them. The first filter capacitor, C1, may vary in value from 1 μF to 10 μF, however, increasing this capacitor above 8 μF increases the output voltage by a negligible amount.

The ripple-voltage reduction for a single section, for each frequency component may be calculated:

$$\frac{E_1}{E_2} = \frac{1}{(2\pi f)^2 LC - 1}$$

where,

E_1 is the ac ripple voltage at the output of the section,
E_2 is the ac ripple voltage at the input of the section,
L is the inductance in henries,
C is the capacitance in microfarads,
f is the frequency of the ripple voltage in Hz.

The total reduction in ripple voltage is approximately the product of the voltage-reduction factors of each section. The percentage of ripple voltage across the first capacitor C1 for a 60-Hz full-wave rectifier, may be approximated:

$$\% E_R = \frac{I_{DC}}{C1}$$

where,

E_R is the ripple voltage,
I_{DC} is the load current,
C1 is the capacity of the first capacitor in microfarads.

An approximation of the percent of ripple may be arrived at for one or two sections of filtering by the equation:

$$\text{One Section } \% E_R = \frac{1}{2\pi f \, L1 \, C1}$$

$$\text{Two Sections } \% E_R = \frac{1}{2\pi f^2 \, C1 \, C2 \, L1 \, L2}$$

where,

L is in henries,
C is in microfarads,
f is the ripple frequency in Hz.

If a swinging choke is to be used, it is selected on the basis of the information given in Question 21.14.

The design of voltage-divider networks is discussed in Question 21.38.

21.68 ***Is it permissible to connect regulated power supplies in parallel?***—Power supplies may be connected in parallel, but the supplies must all have the same maximum compliance voltage ratings. If they do not, and if the load circuit is opened, the terminal voltage will rise to the maximum voltage of the highest-rated supply. If this voltage is greater than the rating of the other supplies, damage may result. To prevent this possibility, diodes are connected in the positive lead of each power supply (Fig. 21-68). Since the diode is in the normal conducting mode, it is capable of withstanding the short-circuit current of its regulator. The PIV rating of the diode must be equal to or greater than the maximum open-circuit potential of the highest-rated power supply.

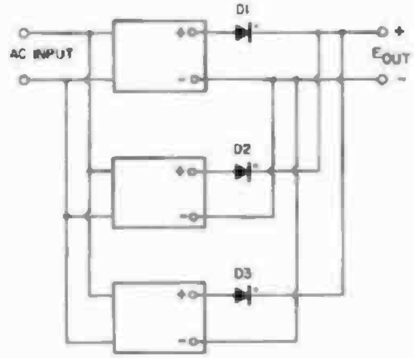


Fig. 21-68. Three regulated power supplies connected in parallel, with diodes connected in series with the plus side to prevent damage from a voltage reversal.

21.69 ***Can regulated power supplies be connected in series?***—Yes, if certain precautions are observed. The isolation voltage rating of the individual power supplies must not be exceeded, and the power supplies must be protected against reverse potential. Diodes are connected in the nonconducting direction, across the output of each supply unit (Fig. 21-69). These diodes will start to conduct the instant a reverse potential appears, and they provide a path around the supplies for short-circuit current. If possible, the regulating circuit for one supply should be connected as a master, and the other as

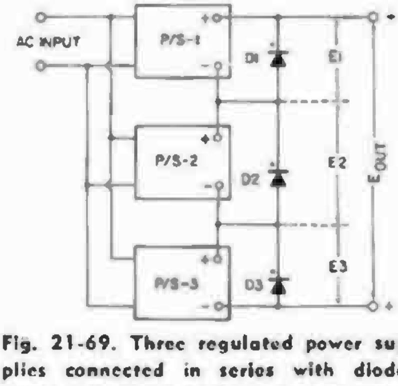


Fig. 21-69. Three regulated power supplies connected in series with diodes connected across each supply to prevent damage because of voltage reversal.

slaves. The voltages of the supplies need not all be the same.

21.70 Can unregulated power supplies be connected in parallel?—Yes, provided they are of similar design and each is adjusted to carry its rated load.

21.71 What are the characteristics of a selenium rectifier?—Basically, a selenium rectifier cell consists of a nickel-plated aluminum base plate coated with selenium, over which a low temperature alloy is sprayed. The aluminum base serves as a negative electrode, and the alloy as the positive. Current flows from the base plate to the alloy, but encounters high resistance in the opposite direction. The efficiency of conversion depends to some extent on the ratio of the resistance in the conducting direction to that of the blocking direction. Conventional rectifiers generally have ratios of 100:1, and 1000:1 for special applications.

Selenium rectifiers may be operated over temperatures of minus 55 degrees to plus 150 degrees centigrade. Rectification efficiency is high, generally on the order of 90 percent for three-phase bridge circuits and 70 percent for single-phase bridge circuits. As a selenium cell ages, the forward and reverse resistance increases for approximately one year, then it stabilizes. Aging decreases the output voltage by approximately 15 percent. Generally, different taps are provided on the power transformer secondary to feed the rectifier and compensate for the loss of the output voltage. The internal impedance of a selenium rectifier is extremely low and exhibits a nonlinear characteristic with respect to the applied voltage, thus maintaining a good voltage regulation.

By nature of their construction, selenium rectifiers have considerable internal capacitance, which limits their operating range to audio frequencies. Approximate capacitance ranges are 0.10 to 0.15 μF per square inch of rectifying surface. A minimum voltage required for conduction in the forward direction is termed the *threshold voltage*. Therefore, selenium rectifiers cannot be used successfully at voltages below 1 volt. A cross-sectional view of a typical selenium disc rectifier is shown in Fig. 21-71.

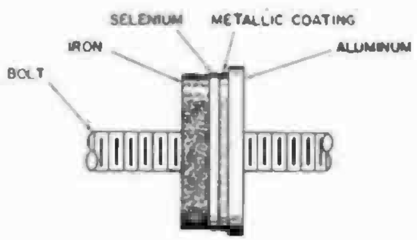


Fig. 21-71. Selenium disc rectifier unit.

21.72 What is a copper-oxide rectifier?—Another form of a disc rectifier, similar in construction to the selenium rectifier. The construction for both a copper-oxide and copper-sulfide disc rectifier is shown in Figs. 21-72A and B.

21.73 Describe silicon rectifiers.—Silicon rectifiers are discussed in Section 11.

21.74 What is a stacked rectifier?—A group of rectifier cells stacked as shown in Fig. 21-74. The unit pictured is a full-wave rectifier.

21.75 What is the forward direction of a rectifier?—It is the direction of

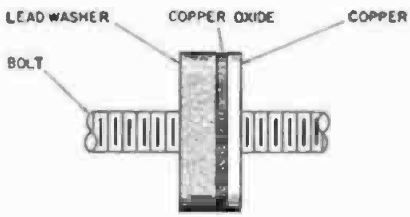


Fig. 21-72A. Construction of a copper-oxide disc rectifier unit.

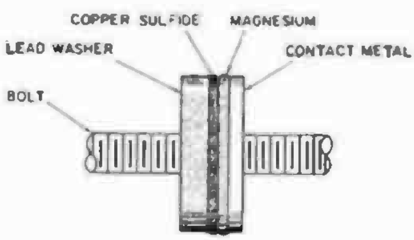


Fig. 21-72B. Copper-sulfide disc rectifier unit.

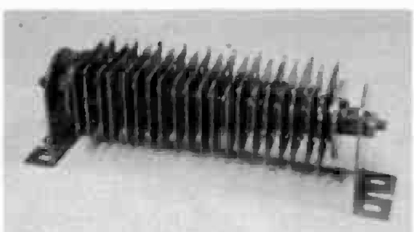


Fig. 21-74. Group of rectifier elements assembled into a complete rectifier unit.

lowest resistance to the current flow through the cell, from the negative to the positive pole of the cell.

21.76 What is the forward resistance of a rectifier?—The resistance of an individual cell measured at a specified forward voltage drop or current.

21.77 What is the reverse resistance of a rectifier?—It is the resistance of an individual cell measured at a specified reverse voltage or current.

21.78 What is the forward voltage drop in a rectifier?—Internal voltage drop in an individual cell, resulting from the flow of current through the cell in the forward direction.

21.79 What is the reverse current of a rectifier?—The current which flows through an individual cell in the reverse direction.

21.80 What are the minimum number of cells that can be employed for rectification?

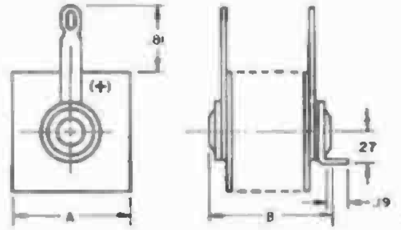
Single-phase, half-wave	1
Three-phase, half-wave	3
Single-phase, full-wave bridge	4
Two-phase, full-wave bridge	8
Three-phase, full-wave bridge	6
Single-phase, full-wave, center tap	2
Three-phase, full-wave, center tap	6
Voltage doubler	2
Voltage tripler	3
Voltage quadrupler	4

21.81 How can the configuration of a rectifier stack be identified?—Cross-sectional views of rectifier stacks are given in Fig. 21-81. With the aid of the circuit configurations given in Fig. 21-91, the connections can easily be identified. For voltage-doubler circuits, the data given in Questions 21.6 to 21.8 will be helpful.

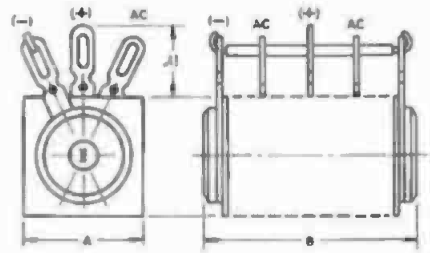
21.82 How are selenium rectifiers rated for motor operations?—They must be rated at least 20 percent higher than for normal operation.

21.83 How are rectifier stages cooled?—Either by forced air or by immersing the stack in oil. Rectifiers to be operated in oil require a special treatment. A suitable oil for rectifier cooling is Transil 10-C.

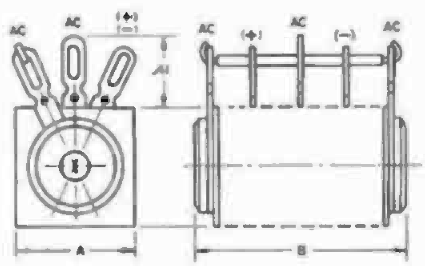
If the rectifiers are to be operated in open air, they are equipped with fins, or heat sinks. Heat sinks and their design are discussed in Questions 21.125 and 21.126.

21.84 How should stacked rectifiers 

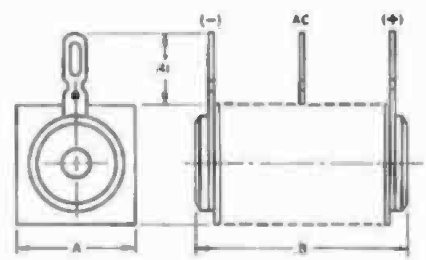
(a) A half-wave stack.



(b) Full-wave center-tap.



(c) Full-wave bridge.



(d) Voltage doubler.

Fig. 21-81. Rectifier stack connections.

be installed?—They should never be installed one above another, but placed alongside each other in the horizontal plane. Mounting in this manner will prevent overheating. For power supplies dissipating considerable heat, a small fan can be installed to provide forced-air cooling.

21.85 Describe a pass element in a power supply.—It is a group of transistors or vacuum tubes connected in par-

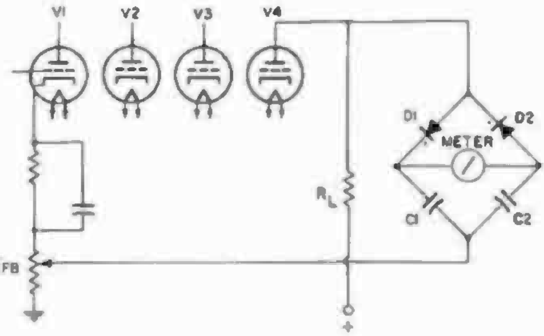


Fig. 21-86. Series half-bridge with capacitors in the lower legs. This circuit is often used in vacuum-tube voltmeters to control the low- frequency response.

allel and placed in series with the output of a regulated power supply to control the flow of the output current. Pass elements are shown in Fig. 21-3 (tubes V3 to V6) and in Fig. 21-121 (transistors Q3, Q5, and Q7).

21.86 ***What is the purpose of connecting capacitors in the lower legs of a series half-bridge circuit?*** — This type circuit is often used in the negative-feedback circuit of a vacuum-tube voltmeter to isolate the metering circuit from the plate voltage of the meter driving tube and to provide equalization for the metering circuit. Fig. 21-86 shows a typical circuit employing capacitors in the lower legs of the bridge, which are returned to the cathode circuit of an early voltage-amplifier stage. With a value of 2-μF capacitance, the frequency response is uniform down to about 30 Hz. Reducing the value of the capacitors decreases the negative feedback (this subject is discussed in Question 12.218), thereby increasing the frequency response at the lower frequencies. A circuit similar to that shown is employed in the vacuum-tube voltmeter discussed in Question 22.100.

21.87 ***Are input capacitors permissible with solid-state rectifiers?***—Yes. Any of the standard rectifier circuits are adaptable to conventional filter-section circuits.

21.88 ***What is a crowbar voltage protector?***—A circuit which monitors the output voltage of a power supply and instantaneously throws a short circuit across the output terminals to operate a preset voltage-limiting device. This is generally accomplished by the use of a silicon controlled rectifier (SCR) connected across the output terminals of the supply unit. (See Question 11-150.)

21.89 ***Over what frequency range can solid-state rectifiers be operated?***—Rectifiers designed for low-frequency use generally have considerable internal capacitance. The amount depends on the current-carrying capacity and the physical design of the rectifier. However, they can be used for power rectification up to several thousand hertz.

Rectifiers used for high-frequency use, such as in radio receivers and metering circuits, are especially designed for such usage. The operating frequency range can be obtained from the manufacturer's data sheets.

21.90 ***What factors determine the choice of a rectifier circuit?***—For selenium rectifiers (when factors like regulation and ripple voltages are not too important), a single-phase half-wave rectifier, using a voltage doubler or tripler for the higher voltages, with the current limited to 120 milliamperes, is quite satisfactory. For 120 milliamperes to 100 amperes, a choice of a full-wave circuit is desirable, using a bridge or center-tap configuration. Below 15 volts, the center-tap circuit offers economy, because only two arms are necessary. However, when the voltage limitations of a single cell have been reached, the choice of a bridge circuit is indicated. For currents between 100 and 1000 amperes, a three-phase wye, double-wye, or six-phase star configuration is recommended for voltages up to 15 volts, and a bridge circuit at higher voltages. The graphs in Figs. 21-90A and B will be helpful in determining these factors.

The charts in Fig. 21-90C and D may be used as an aid in the selection of the proper circuitry for selenium and silicon rectifiers. Referring to Fig. 21-90C, half-wave, single-phase circuits are practical. Doublers and triplers at currents up to and including 1 ampere are also practical. The exception is for cur-

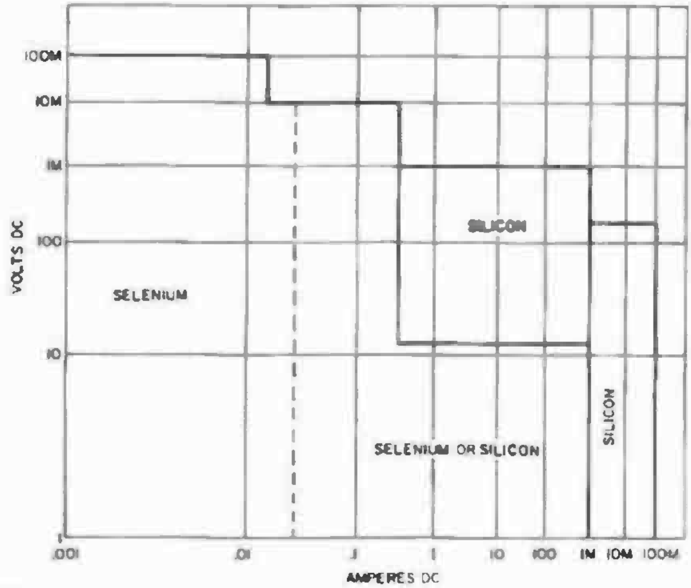


Fig. 21-90A. Voltage and current chart for the selection of a rectifier type. (Courtesy, Sarkes Tarzian Inc.)

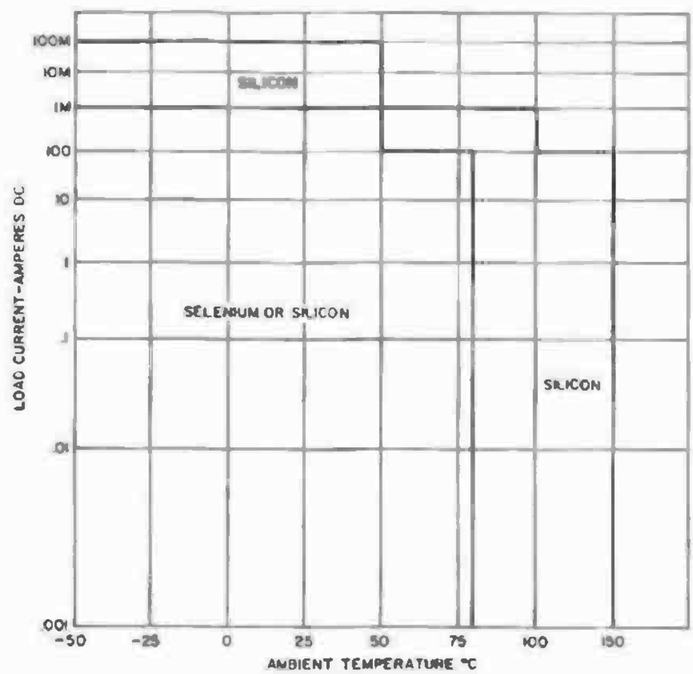


Fig. 21-90B. Current and temperature chart for the selection of a rectifier type. (Courtesy, Sarkes Tarzian Inc.)

rents over 500 milliamperes. A bridge circuit might be preferable when the voltage exceeds 250 volts dc. Availability of high-voltage, single-junction rectifiers extends the practical range of the single-phase, center-tap configuration to 250 volts, with a similar pattern for the selenium rectifier with wye or star connections preferred up to 250 volts, and bridge connections above 250 volts. It should be understood that the above recommendations are only general and are subject to modification depending on the application.

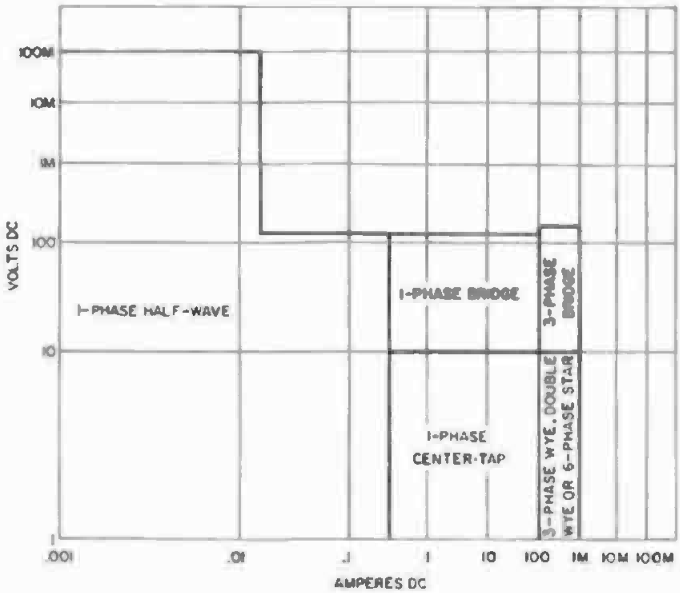


Fig. 21-90C. Chart for circuit selection, selenium rectifiers. (Courtesy, Sarkes Tarzian Inc.)

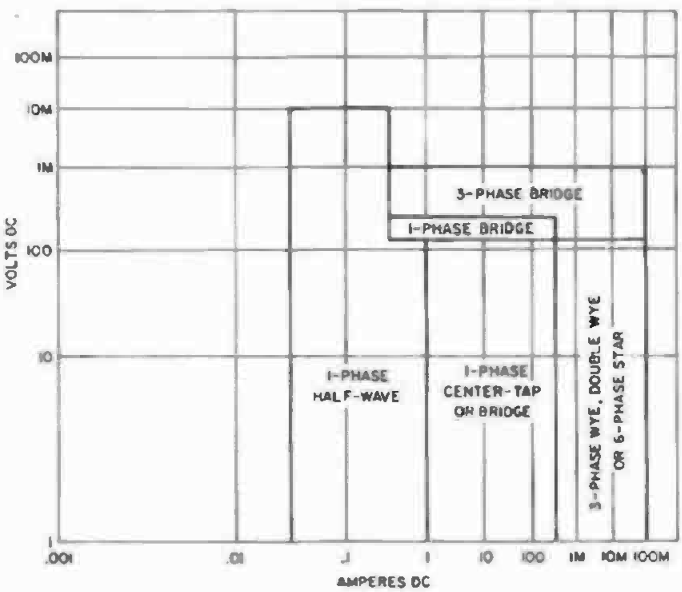
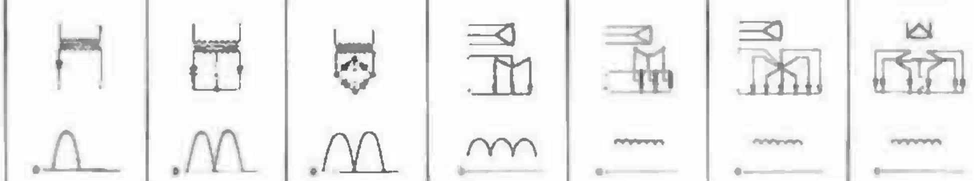


Fig. 21-90D. Chart for circuit selection, silicon rectifiers. (Courtesy, Sarkes Tarzian Inc.)

21.91 Give the formulas and configurations for single- and three-phase rectifier circuits.—The formulas and circuitry for designing the majority of rectifier circuits, both single and three phase, are given in Fig. 21-91. By referring to the desired configuration, factors involved in the design can be read directly from the chart. The percent ripple quoted is the ripple frequency at the output rectifier circuit without the benefit of filtering. The information given can be used for solid-state or vacuum-tube rectifiers. (See Question 21.11.)

21.92 What are the peak-inverse voltages (PIV) for single- and three-phase rectifier circuits?—Full-wave con-



Type of Circuit	Single Phase Half Wave	Single Phase Center-Tap	Single Phase Bridge	Three Phase Star (Wye)	Three Phase Bridge	Six Phase Star (Three Phase Diametric)	Three Phase Double Wye with Interphase Transformer	* Assumes zero forward drop and zero reverse current in rectifiers and no AC line or source resistance. To determine actual value of parameter in first column, multiply factor shown by value of:
Number of rectifier elements in circuit	1	2	4	3	6	6	6	
Average D.C. volts output	1.00	1.00	1.00	1.00	1.00	1.00	1.00	Average D.C. voltage output
RMS D.C. volts output	1.57	1.11	1.11	1.02	1.00	1.00	1.00	× Average D.C. voltage output
Peak D.C. volts output	3.14	1.57	1.57	1.21	1.05	1.05	1.05	× Average D.C. voltage output
Peak reverse volts per rectifier element	3.14	3.14	1.57	2.09	1.05	2.09	2.42	× Average D.C. voltage output
	1.41	2.82	1.41	2.45	2.45	2.83	2.83	× RMS secondary volts per transformer leg
	1.41	1.41	1.41	1.41	1.41	1.41	1.41	× RMS secondary volts line-to-line
[illegible]	1.00	1.00	1.00	1.00	1.00	1.00	1.00	× Average D.C. output current
[illegible] per rectifier element	1.00	0.500	0.500	0.333	0.333	1.67	0.167	× Average D.C. output current
RMS current per rectifier element, resistive load	1.57	0.785	0.785	0.587	0.579	0.409	0.293	× Average D.C. output current
RMS current per rectifier element, inductive load	----	0.707	0.707	0.578	0.578	0.408	0.289	× Average D.C. output current
Peak current per rectifier element, resistive load	3.14	1.57	1.57	1.21	1.05	1.05	0.525	× Average D.C. output current
Peak current per rectifier element, inductive load	----	1.00	1.00	1.00	1.00	1.00	0.500	× Average D.C. output current
Ratio peak to average current per element, resistive load	3.14	3.14	3.14	3.63	3.15	6.30	3.15	
Ratio peak to average current per element, inductive load	----	2.00	2.00	3.00	3.00	6.00	3.00	
% Ripple (RMS of ripple / average output voltage)	121%	48%	48%	18.3%	4.2%	4.2%	4.2%	
	Resistive Load	Inductive Load or Large Choke Input Filter						
Transformer secondary RMS volts per leg	2.22	1.11	1.11	0.855	0.428	0.740	0.855	× Average D.C. voltage output
Transformer secondary RMS volts line-to-line	2.22	2.22	1.11	1.48	0.740	1.48	1.71	× Average D.C. voltage output
Secondary line current	1.57	0.707	1.00	0.578	0.816	0.408	0.289	× Average D.C. output current
Transformer secondary volt-amperes per leg	3.49	1.57	1.11	1.48	1.05	1.81	1.48	× D.C. watts output
Transformer primary RMS amperes per leg	1.57	1.00	1.00	0.471	0.816	0.577	0.408	× Average D.C. output current
Transformer primary volt-amperes per leg	2.46	1.11	1.11	1.21	1.05	1.28	1.05	× D.C. watts output
Average of primary and secondary volt-amperes	3.49	1.34	1.11	1.35	1.05	1.55	1.26	× D.C. watts output
Primary line current	1.57	1.00	1.00	0.817	1.41	0.817	0.707	× [illegible]
Line power factor	----	0.900	0.900	0.826	0.955	0.955	0.955	

Fig. 21-91. Rectifier design data. The data given are applicable to both solid-state and vacuum tubes.

ventional rectifier, 2.82 times one-half the secondary voltage.

Full-wave bridge circuit, 1.5 times the rms voltage of the secondary. Half-wave, 2.82 times the rms secondary voltage.

Three-phase star and bridge, 2.45 times the rms voltage per transformer leg.

Six-phase star (three-phase diametric) and three-phase double-wye, 2.83 rms per transformer leg. (See Question 21.91.)

21.93 Are selenium rectifiers suitable for battery chargers?—Yes. Their characteristics provide a reduction of the charging current as the voltage of the battery rises. This characteristic prevents overcharging and raising the temperature of the battery.

21.94 Describe the terminal symbols for a semiconductor rectifier.—For the small tubular and top-hat-type rectifier, the polarity symbols are indicated on the rectifier unit, as given in Fig. 21-94. The theory of operation for the semiconductor-type rectifier is discussed in Question 11.109. It will be noted in the data given in Fig. 11-109 that the bar representing the cathode (negative) of the rectifier always points away from the source of voltage to develop a positive potential at its output.

21.95 What is a full-wave bridge rectifier and what are its advantages?—The chief advantage of the full-wave bridge rectifier is its ability to supply full-wave rectification without a center tap on the transformer. However, the bridge rectifier is not a true single-ended circuit, since it has no terminal common to both the input and output circuits.

In the vacuum-tube-type rectifier, the bridge rectifier circuit is more easily adapted to indirectly heated cathode-type tubes than to those using a filament-type heater. The reason for this is that a separate heater winding is required for each tube in the circuit. This means that for a full-wave bridge circuit, four separate heater windings are required. For tubes using a cathode element, only a single heater winding is required. However, this is not true for high-powered rectifier tubes because most of them employ heater-type construction. This has, in some instances, deterred the use of full-wave bridge rectifier circuits.

The semiconductor-type rectifier, requiring no heating, may be adapted to the full-wave bridge circuit without inconvenience. Such rectifiers include the copper oxide, selenium, copper-sulphide, silicon, germanium, and titanium types.

Several types of bridge circuits are available for full-wave rectification and are classified according to the number of diodes or rectifier elements employed.

A full-wave bridge rectifier consists of four rectifier elements, as shown in Fig. 21-95. This is the most familiar circuit and is the one most commonly employed in the electronics industry. The circuits function in the following manner:

When the ac input Terminal 1 is positive, current flows in the direction of the solid arrows from Terminal 1, through the low forward resistance of diode D2, through the load resistance R_L, diode D3, and, finally, to Terminal 2 which is negative at this instant. Conduction through D1 and D4 is negligible because these diodes present their high back resistance when Terminal 1 is positive and Terminal 2 is negative.

On the next half-cycle when Terminal 2 swings positive, current flows in the direction of the dotted arrows from Terminal 2, through the low forward resistance of diode D4, through the load resistance R_L, diode D1, and, finally, to Terminal 1 which is negative at this instant. Couduction through diodes D2 and D3 is negligible because these diodes present their high back

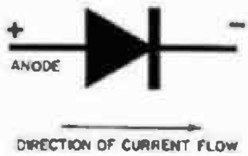


Fig. 21-94. Symbols on a solid-state diode rectifier, and the direction of current flow.

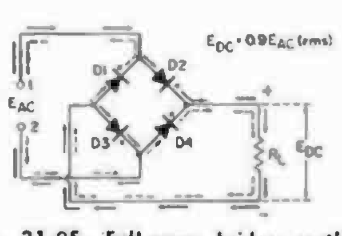


Fig. 21-95. Full-wave bridge rectifier.

resistance when Terminal 2 is positive and Terminal 1 is negative. It will be noted that the rectified current indicated by the solid arrows for one half-cycle and the dotted arrows for the other half-cycle always passes through the load resistance R_L in the same direction. The top of R_L is always positive and the bottom negative. Thus, full-wave rectification is obtained.

With the full-wave bridge circuit, the dc output voltage is equal to 0.9 of the rms value of the ac input voltage.

21.96 What is a three-quarter bridge?—A bridge connection in which one diode rectifier has been replaced with a resistor R_s as shown in Fig. 21-96. When the ac input voltage at Terminal 1 is positive, the current flows in the direction of the solid arrows from Terminal 1 through the low forward resistance of diode D2, through load resistance R_L, resistor R_s, and, finally, to Terminal 2 which is negative at this instant. Any conduction through diodes D1 and D3 is negligible during this half-cycle because these diodes now present their high back resistance and provide effective blocking action.

On the next half-cycle, when Terminal 2 is positive, current flows in the direction of the dotted arrows from Terminal 2, through diode D3, load resistance R_L, diode D1, and, finally, to Terminal 1 which is negative at this instant. Any conduction through diode D2 is negligible, since this diode now presents its high back resistance.

Rectified current flowing through load resistance R_L is shown by the solid and dotted arrows to be in the same direction during each half-cycle of ac input voltage. The top of R_L, therefore, is always positive and the bottom is always negative. This satisfies the conditions for full-wave rectification. With the three-quarter bridge circuit, the dc output voltage is equal to 0.84 of the rms value of the ac input voltage. The rectification efficiency of the three-quarter bridge is 93.3 percent of that of the full-wave bridge circuit of Fig. 21-95.

21.97 What is an opposed half-bridge?—Two diodes have been replaced with resistors in the circuit, as shown in Fig. 21-97. Using half the number of diodes employed in the full-bridge, this circuit is termed a half-bridge. Specifically, it is of the opposed half-bridge type in which the equal resistors R1 and R2 are connected in series across the ac input.

In this circuit, when ac input Terminal 1 is positive, current flows in the direction of the solid arrows from Terminal 1 through diode D1, load resistance R_L, resistor R2, and, finally, to Terminal 2 which is negative at this instant. Conduction through diode D2 is negligible during this half-cycle, since the diode now presents its high back resistance.

On the next half-cycle of the ac input voltage, Terminal 2 swings positive and current flows in the direction of the dotted arrows from Terminal 2, through diode D2, load resistance R_L, resistor R1, and, finally, to Terminal 1 which is negative at this instant. Conduction through diode D1 is negligible, since this diode presents its high back resistance during this half-cycle.

The rectified current flowing through the load resistance R_L is shown by the solid and dotted arrows to be in the same direction for each half-cycle of ac input voltage. This satisfies the conditions for full-wave rectification.

With the opposed half-bridge circuit, the dc output voltage is equal to 0.72 of the rms value of the ac input voltage. The rectification efficiency of the opposed bridge is 80 percent of that of the full-bridge circuit in Fig. 21-95.

21.98 What is a series half-bridge?—It is a second type of half-bridge circuit and is illustrated in Fig. 21-98. In

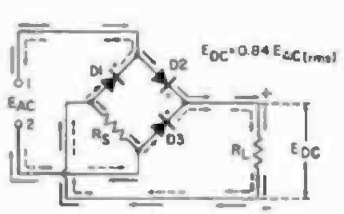


Fig. 21-96. Three-quarter bridge rectifier circuit.

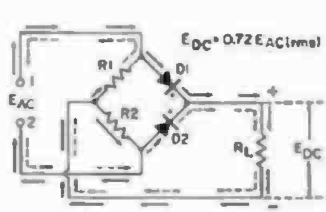


Fig. 21-97. Opposed half-bridge rectifier circuit.

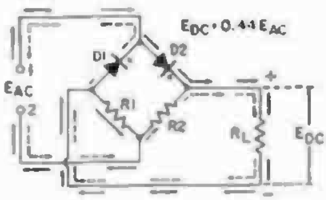


Fig. 21-98. Series half-bridge rectifier.

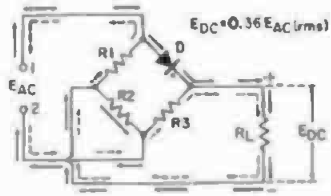


Fig. 21-99. Quarter-bridge rectifier.

this arrangement, two series diode and resistor legs are connected across the ac supply, one diode D2 being polarized for high conduction when ac Terminal 1 is positive and the other diode D1 polarized for high conduction when ac Terminal 2 is positive.

When ac terminal 1 is positive, current flows in the direction of the solid arrows from Terminal 1 through diode D2, load resistor R_L, resistor R1, and, finally, to Terminal 2 which is negative at this instant. Conduction through diode D1 is negligible, since this diode presents a high back resistance during this half-cycle.

On the next half-cycle of ac input voltage, Terminal 2 swings positive and current flows from Terminal 2 in the direction of the dotted arrows through resistor R2, load resistance R_L, diode D1, and, finally, to Terminal 1 which is negative at this instant. Conduction through diode D2 is negligible, since this diode presents its high back resistance during this half-cycle.

The rectified current flowing through load resistance R_L is shown by the solid and dotted arrows to be in the same direction during each half-cycle of the ac input voltage. The top of R_L is always positive and the bottom negative. This satisfies the conditions for full-wave rectification.

With the series half-bridge circuit, the dc output voltage is equal to 0.44 of the rms value of the ac input voltage. The rectification efficiency of the series half-bridge is 48.9 percent of that of the full-wave bridge circuit in Fig. 21-95.

21.99 What is a quarter-bridge?—This unique bridge circuit is shown in Fig. 21-99. It employs only one diode and three resistors. Since this arrangement employs only one-fourth the number of diodes required for a full-wave bridge rectifier, it is termed a quarter-bridge.

The quarter-bridge consists of the left (series resistance) half of the opposed half-bridge of Fig. 21-97 and the right (diode resistance) half of the series half-bridge of Fig. 21-98.

In the quarter-bridge circuit, when ac input Terminal 1 is positive, current flows in the direction of the solid arrows from Terminal 1 through diode D, load resistance R_L, resistor R2, and, finally, to Terminal 2 which is negative at this instant.

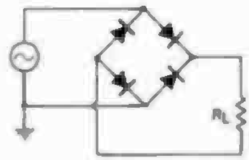
On the next half-cycle of the ac input voltage, Terminal 2 swings positive and current flows in the direction of the dotted arrows from Terminal 2, through resistor R3, load resistance R_L, resistor R1, and, finally, to Terminal 1 which is negative at this instant.

The rectified current flowing through the load resistor is shown by the solid and dotted arrows to be in the same direction for each half of the ac input voltage. The top of R_L is always positive and the bottom negative. Therefore, the condition for full-wave rectification is satisfied. The dc output voltage is equal to 0.36 of the rms value of the ac input voltage. Rectification efficiency of this circuit is 40 percent of that of the full-bridge circuit in Fig. 21-95.

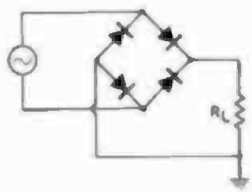
21.100 Where are series half-bridge and quarter-bridge rectifiers employed?—For light-duty applications such as instrumentation and control devices to supply a ripple frequency twice the frequency of the applied voltage.

21.101 What precautions must be taken when using bridge rectifier circuits?—In order that both even- and odd-numbered positive peaks of the dc output voltage have the same amplitude thus insuring maximum average current, the values of the bridge resistors must be correctly proportioned with respect to the forward resistance of the diode or diodes at the applied voltage level.

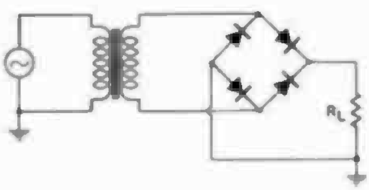
21.102 How should a bridge rectifier circuit be grounded?—Three methods of grounding are shown in (a) to



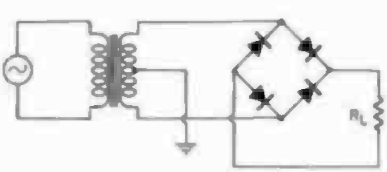
(a) *Ac grounded.*



(b) *Dc grounded.*



(c) *Ac and dc grounded using an isolation transformer.*



(d) *Isolation transformer with center tap to ground.*

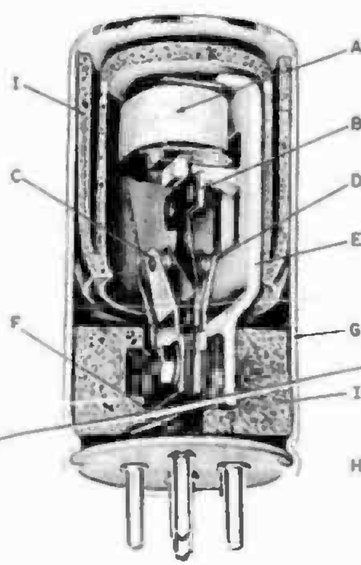
Fig. 21-102. Methods of grounding a bridge-rectifier circuit.

(c) of Fig. 21-102. Either the input (ac source) or output (dc load) may be grounded, but not both simultaneously. However, if an isolation transformer is used between the ac source and the input to the rectifier, as shown in (c) of Fig. 21-102, both ac and dc sides may be grounded permanently. An alternate method of grounding is shown in (d) of Fig. 21-102, where the center tap of an isloation transformer is grounded.

21.103 What precautions should be taken in the selection of a power transformer for bridge rectifier circuits?—An advantage of the bridge rectifier is its ability to utilize the full winding and total voltage of a conventional center-tapped transformer, providing full-wave rectification, while the conventional full-wave, center-tapped circuit utilizes only one-half the total voltage. The power capabilities of the transformer remain the same.

Although twice the dc output voltage is available with the bridge rectifier under these circumstances, the permissible power drain will be only one-half that allowed with full-wave, center-tapped operation.

21.104 What is a vibrator power supply?—A power supply employing a mechanical vibrator operated from a source of direct current such as a battery. The vibrator breaks the current in the primary of a step-up transformer. The secondary of the transformer is connected to a rectifier for changing the alternating current generated by the vibrator in the primary to a high-voltage direct current. Such power supplies are used in auto radios, transmitters, and portable test and sound equipment. Vibrator power supplies are designed to operate both synchronous and nonsynchronous. The construction of a typical nonsynchronous vibrator manufactured by P. R. Mallory and Co. is shown in Fig. 21-104. The operation and circuit are described in Question 21.105.



A—ACTUATING COIL
B—VIBRATOR REED
C—CONTACTS
D—CONTACTS
E—FRAME
F—CONNECTING LEADS
G—CASE
H—CONNECTING PINS
I—SPONGE RUBBER

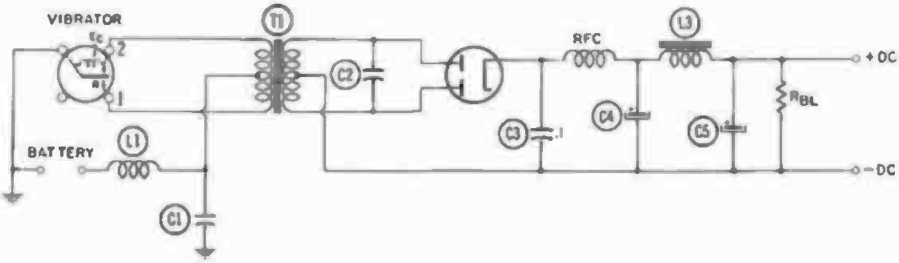


Fig. 21-105. Nonsynchronous vibrator supply.

21.105 How does a nonsynchronous vibrator supply function?—The circuit of a nonsynchronous vibrator is shown in Fig. 21-105. The device consists of a mechanical vibrator, step-up transformer, rectifier, and filter system. The vibrator is caused to operate by the application of direct current to an energizing coil E_c.

The coil surrounds a soft-iron core placed close to the end of the vibrator reed R. The core of the coil attracts the reed, causing it to close with contact 2. This action short-circuits the energizing coil and the reed swings in the opposite direction to close with contact 1. Mechanical inertia makes the reed swing back toward contact 2. This process will be repeated as long as a source of voltage is applied to the energizing coil E_c.

Each time the reed closes with one contact or the other, a surge of current passes through the primary of transformer T1 in a given direction. In contacting first one contact and then the other, surges of current of opposite polarity are produced. These pulses are repeated many times per second and are then rectified on the secondary side by the use of a conventional rectifier and filter system. Two radio-frequency filter sections are used to remove hash or radio-frequency interference. These sections contain rf choke L1 and capacitor C1 in the primary circuit and L2 and C2 on the rectifier side. The make-and-break action of the vibrator causes high instantaneous voltages to be generated because of the high rate of change of current as the circuit is broken. To reduce the effect of these surges every time the primary circuit is broken, a buffer capacitor C2 is connected across the secondary of the step-up transformer T1. The use of this capacitor across the secondary prevents the vibrator contacts from being burned and the rectifier tube from being damaged.

As a rule, the value of the buffer capacitor is about 0.01 μF, with a voltage rating of at least 1000 volts. Since the turns ratio of the transformer is about 100 to 1, the reflected capacitance to the primary will be on the order of 1 μF. (See Question 8.34.) Vibrator frequencies vary between 60 and 180 Hz.

21.106 What is a synchronous vibrator power supply?—It is similar in construction and operation to the nonsynchronous vibrator described in Question 21.105, except that two additional contacts, 3 and 4, are added to the vibrator reed. (See Fig. 21-106.)

When the reed is in contact with contact 1, contact 3 connects the top end of the secondary of transformer T1 through the reed to ground. When the

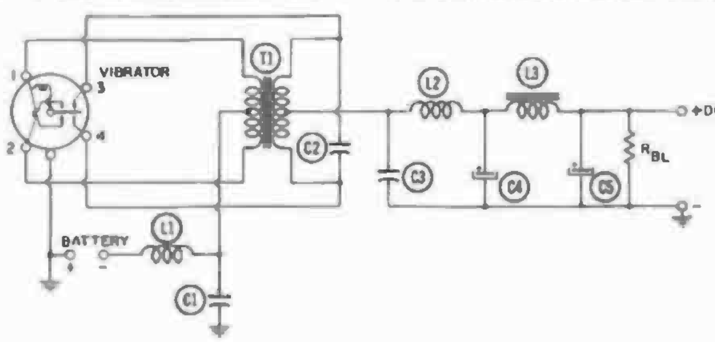


Fig. 21-106. Synchronous vibrator power supply.

reed swings back to the opposite side, the bottom of the transformer secondary is grounded. In this manner, the center tap on the secondary side is always positive with respect to ground. Thus, the output of the secondary is rectified without the benefit of a rectifier tube. Only a filter is required. With this type of operation, current flows in opposite directions through each half of the secondary winding during alternate halves of the cycle. The battery polarity is established in the original design of the power supply and must be observed for proper generation and rectification of the secondary voltage.

21.107 What is an electronic inverter?—A circuit which converts direct current to alternating current. Inverters are used in applications where the primary source of power is direct current. Because direct current cannot be transformed, it is convenient to convert direct current to alternating current so that ac output from the inverter may be applied to a transformer to supply the desired voltage. In Fig. 21-107 is shown the basic circuit of such an inverter.

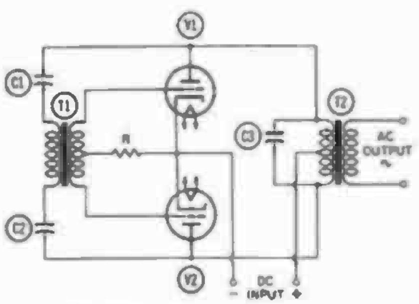


Fig. 21-107. Electronic inverter for converting dc to ac.

When the dc voltage is first applied, the electron flow is approximately equal through both branches of a parallel circuit formed by tubes V1 and V2 in series with corresponding halves of the primary of transformer T2. The voltage across each half of the transformer will then be equal but opposite in polarity and no voltage will appear across C3. However, any slight transient creates a slight voltage across C3 as well as across the series circuit comprising C1, the primary of T1, and C2. This small voltage induces a voltage across the secondary of T1 which is applied to the control grid of V1 and V2. Tubes V1 and V2 amplify the small voltage and apply the amplified voltage to C3 in the proper phase to add to the original unbalance.

This process repeats itself until one of the tubes reaches saturation and the other is cut off. At this instant, C3 and the primary of T2 form a parallel resonant circuit and the discharge of C3 through the primary of T2 produces oscillations in the resonant circuit at a frequency which is determined by the value of C3 and the effective inductance of the primary of T2. The oscillation induces an alternating voltage of the same frequency in the secondary of T2, the output of the inverter. The energy coupled out of the circuit by the secondary of T2 must be supplied through V1 and V2 from the dc source.

Energy is lost in the resistance of the transformer windings and must be supplied from the dc source. This is accomplished by coupling some of the resonant circuit voltage to the grids of V1 and V2 through C1, T1, and C2.

The voltage is fed back in the proper phase to cause V1 and V2 to conduct every half-cycle. The direct-current input is fed to the resonant circuit at just the right moment to add to the instantaneous oscillating energy.

For example, suppose the energy in the resonant circuit is charging the top of C3 positive. The charging voltage is impressed across the primary of T1 and its secondary is connected so that the voltage is a negative-going voltage on the grid of V1 and a positive-going voltage on the grid of V2. V1 will conduct less and V2 more. Therefore, the plate of V1 is made more positive and the plate of V2 more negative. The dc source is thus connected to C3 to aid the charging of C3 by supplying energy in the resonant circuit. When this energy is next discharged into the primary of T2, it is available as output energy at the secondary of T2.

The energy supplied by the dc source will be largely determined by the energy drawn from the alternating-current output. The output of this type inverter is very nearly a sine wave, as a result of the parallel resonant properties of the circuit.

21.108 Describe the basic operating principles of a dc-to-dc converter, using switching transistors.—Basically, a dc-to-dc converter consists of a dc source of potential (generally a battery) applied to a pair of switching transistors. The transistors convert the applied dc voltage to a high-frequency ac voltage. The ac voltage is then transformed to a higher voltage which is rectified to dc again and filtered in the conventional manner. Power supplies of this nature are often used for a source of high voltage, where the usual ac line voltage is not available.

The circuitry for such a device appears in Fig. 21-108. The power transformer is rather special and consists of a "Deltamax" toroidal core, with a saturation flux density of 14,000 kilogauss. Three windings are required: a 52-turn primary, a 68-turn feedback winding, and a 550-turn, high-voltage secondary winding. The primary and feedback windings are bifilar wound to obtain a tight coupling between the two halves of the winding on each side of the center tap. The switching frequency is generally on the order of 1000 to 2000 Hz to reduce the size and weight of the transformer and filter chokes (if used).

Designing a device of this nature for a frequency of 60 Hz would increase the weight and size considerably over that of one using a higher frequency. The action of the switching transistors is now explained. Applying a voltage from the positive input terminal of the battery causes a current to flow to the emitter of Q1 and induces a voltage in the feedback winding F_w in such a manner that the base of Q2 is made positive with respect to the base of Q1. These polarities at the bases of Q1 and Q2 cause an increase in the current through Q1. At the same time, they cause the current through Q2 to decrease. This amounts to positive feedback. Therefore, transistor Q1 is turned on and Q2 is turned off.

The current continues to increase from the input terminal until the core becomes saturated. The induced voltage in the feedback winding then drops to zero, turning off transistor Q1. The collapsing field of the core induces a voltage of opposite polarity in the feedback winding, which further turns off Q1 and turns on Q2. Current then flows from the plus voltage to the emitter of Q2 through the feedback winding. Saturation again takes place—this time in the negative direction. This action provides suitable conditions for oscillation repeated at a frequency dependent on the circuit constants.

The voltage at the secondary is of a square-wave nature and is rectified by a fnll-wave bridge-rectifier circuit and filtered iu the usnal manner. The freqnency of the described circuit is about 1600 Hz and is capable of supplying 250 Vdc at a current of 100 milliamperes. The discussed circuit may be, in some manners, compared to that of the nonsynchronous vibrator iu Question 21.105.

21.109 *Describe high-frequency power supplies.*—These are power supplies employing a high-frequency oscillator, which operates at 30 to 80 kHz. They are used in 16-mm motion picture projectors, television receivers, electron microscopes, and other devices. Sharp turns in the wiring must be avoided to prevent corona discharge at the point of the bend. Corona discharge can cause serious interfereuce in adjacent circuits and other equipment in

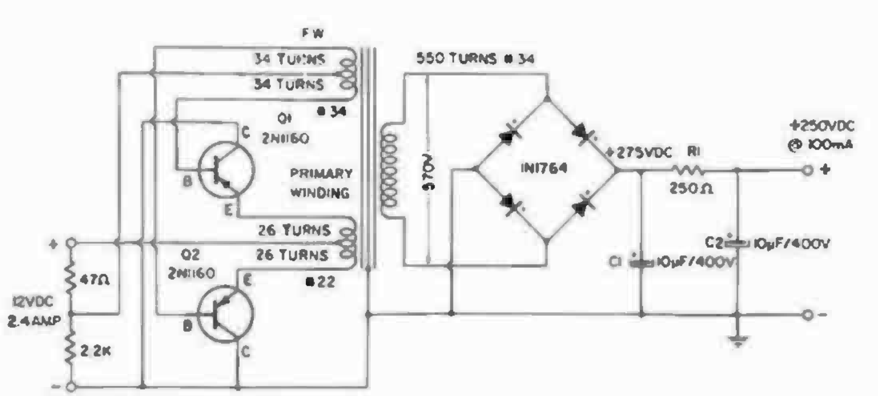


Fig. 21-108. A dc-to-dc converter using switching transistors. Switching frequency is approximately 1600 Hz, using the above circuit components.

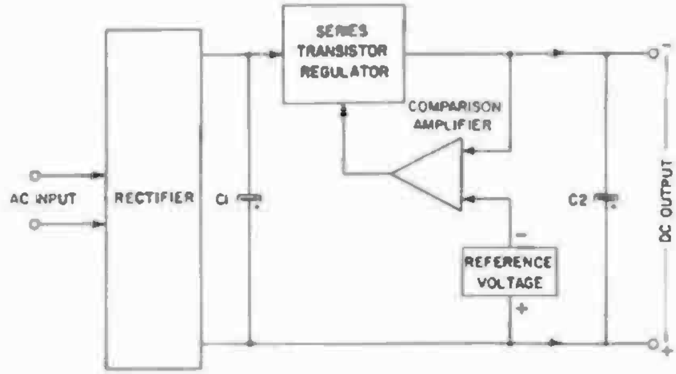


Fig. 21-111A. Block diagram for a constant-voltage regulated power supply.

the general area. These supplies are discussed in Question 19.165.

21.110 *What is a tuning fork power supply?*—A power supply using a precision tuning-fork oscillator to control the output frequency. It is used in applications where constant speed of electric motors is a prime requisite, such as in sound-recording equipment. The device consists of a two-stage RC amplifier with a 60-Hz tuning fork in a feedback circuit acting as the frequency-determining element. Following the oscillator section is a push-pull power amplifier capable of developing about 70 to 100 watts of power.

21.111 *Describe a constant-voltage power supply.*—It is a regulated power supply designed to keep its output voltage constant, regardless of the changes in load current, line voltage, or temperature. For a change in the load resistance, the output voltage remains constant to a first approximation, while the output current changes by whatever amount is necessary to accomplish this. A block diagram of this type supply appears in Fig. 21-111A. Its impedance characteristics are given in Fig. 21-111B. (See Question 21.47.)

21.112 *Describe a constant-current power supply.*—It is a regulated power supply designed to keep its output current constant, regardless of the changes in load current line voltage, or temperature. For a change in the load resistance, the output current remains constant to a first approximation, while the output voltage changes by whatever amount is necessary to accomplish this. A block diagram of this type supply appears in Fig. 21-112A. Its impedance characteristics are given in Fig. 21-112B. (See Question 21.47.)

21.113 *Describe a constant-voltage, constant-current power supply.*—It is a power supply that acts as a constant-voltage source for comparatively large values of load resistance and as a constant-current source for comparatively

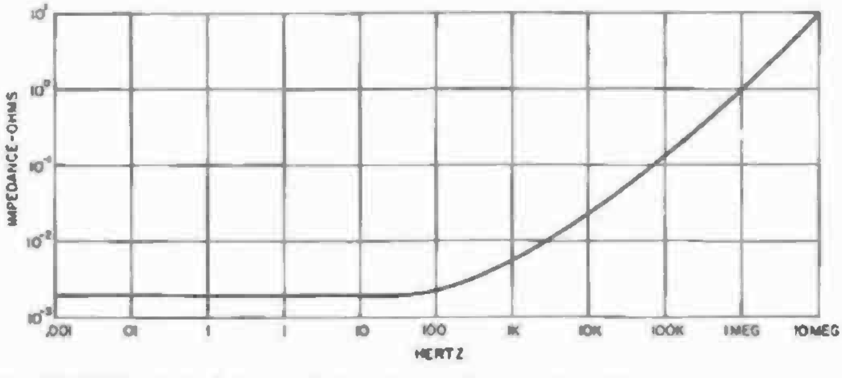


Fig. 21-111B. Typical internal impedance characteristic for a constant-voltage power supply.

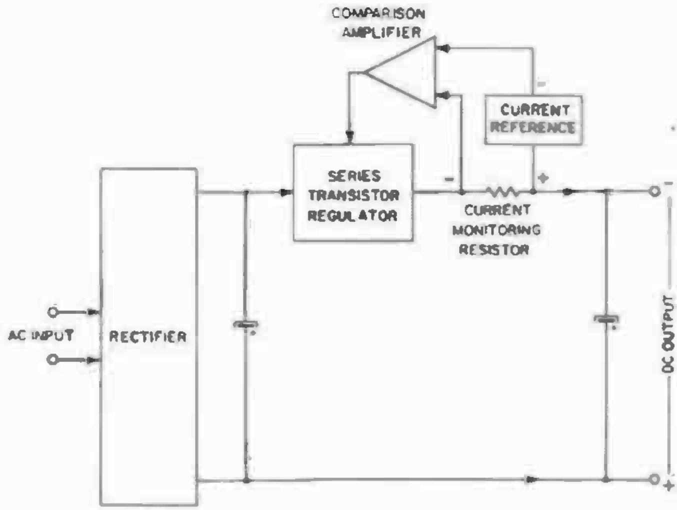


Fig. 21-112A. Block diagram for a constant-current regulated power supply.

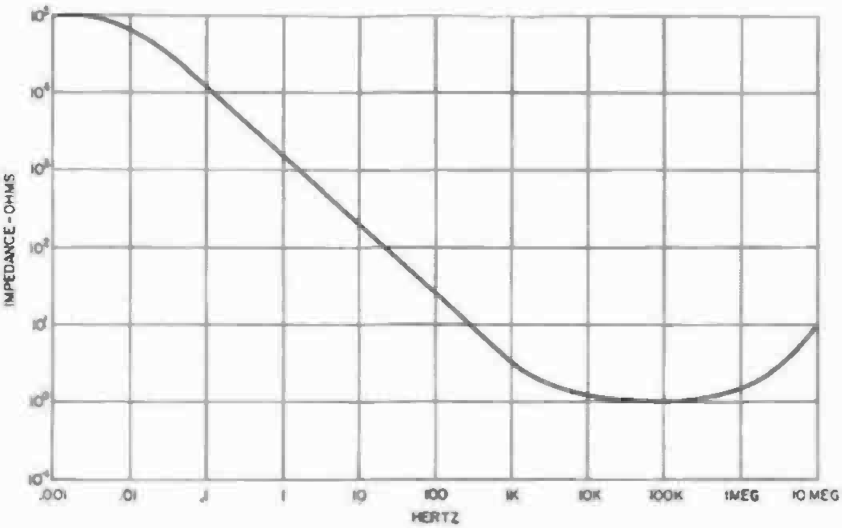


Fig. 21-112B. Typical internal output impedance characteristic for a constant-current power supply.

small values of load resistance. An automatic crossover (or transition) between these two modes of operation occurs at a critical or crossover value of load resistance where R_c equals E_s/I_s, where,

- E_s is the voltage-control setting,
- I_s is the current-control setting generally indicated by meters on the front panel.

A block diagram of this type supply appears in Fig. 21-113.

21.114 How is automatic crossover accomplished in a constant-voltage, constant-current power supply?—Referring to Fig. 21-113, the disconnect diodes are connected so that when the supply is in the constant-voltage mode, the upper diode is forward-biased (shorted), and the lower diode is reverse-biased (open). Conversely, when the supply is in the constant-current mode, the upper diode is reversed-biased, and the lower diode is forward-biased. Thus, the series transistor regulator is only called upon to respond to either the constant-voltage or the constant-current comparison amplifier, and the effectiveness of one amplifier is not affected by the shunt presence of the other.

21.115 What is transient recovery time in a power supply?—Transient re-

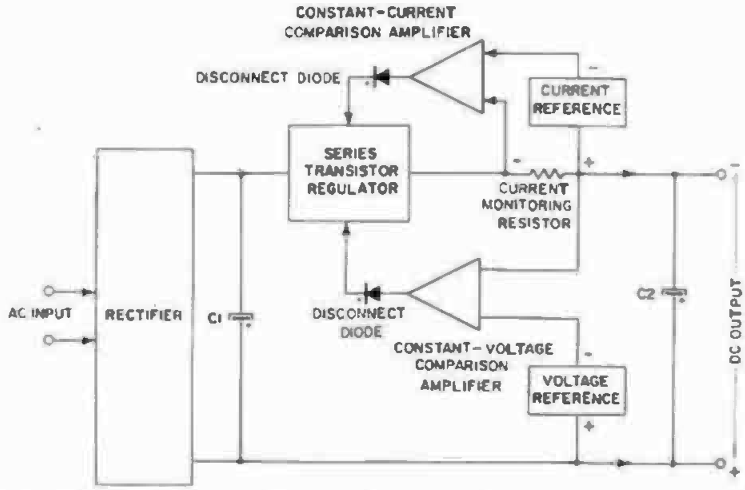


Fig. 21-113. Block diagram for constant-voltage/constant-current power supply with automatic crossover.

covery time is defined as the time required for the output voltage to recover within X millivolts of the normal output voltage, where the nominal output voltage is the mean between no-load and full-load voltage.

When the output load current of a power supply is changed at a rapid rate, a voltage spike occurs across the output terminals. This voltage spike may lie outside the normal static regulation specifications for a relatively short period of time. In order to describe the performance of power supplies for pulse-loading applications, it is necessary to define and measure the duration and amplitude of the unwanted output voltage transient. The term used for this operation is transient recovery time, or recovery time.

Fig. 21-115A shows the general nature of the output voltage transient resulting from an imposed external square-wave load current (Fig. 21-115B). The feedback amplifier within the power supply reduces the effective output impedance at all frequencies within its band of amplification. At high frequencies however, the gain of the amplifier disappears and the output impedance of the supply increases. Under these conditions, the power supply becomes essentially a passive device which has an output impedance dependent mainly on the internal effective inductance of the electrolytic capacitors and the leads connecting to the power-supply output terminals. Special electrolytic capacitors are used for this purpose and are located physically close to the output terminals to reduce the effect of lead inductance. An expanded view of transient recovery time, showing its characteristics, is given in Fig. 21-115C.

21.116 What is a remote sensing circuit.—It is a circuit incorporated in a regulated power supply. By means of

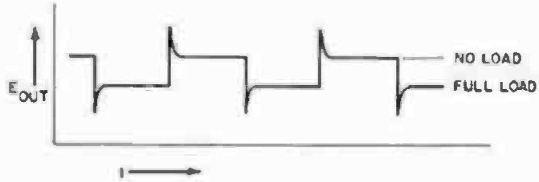


Fig. 21-115A. Characteristic of square wave during the recovery period.

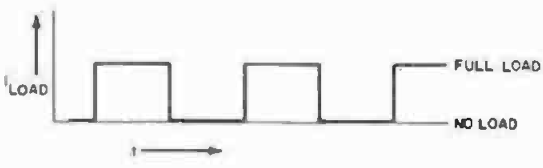


Fig. 21-115B. Square wave applied across the output terminals to measure the recovery time.

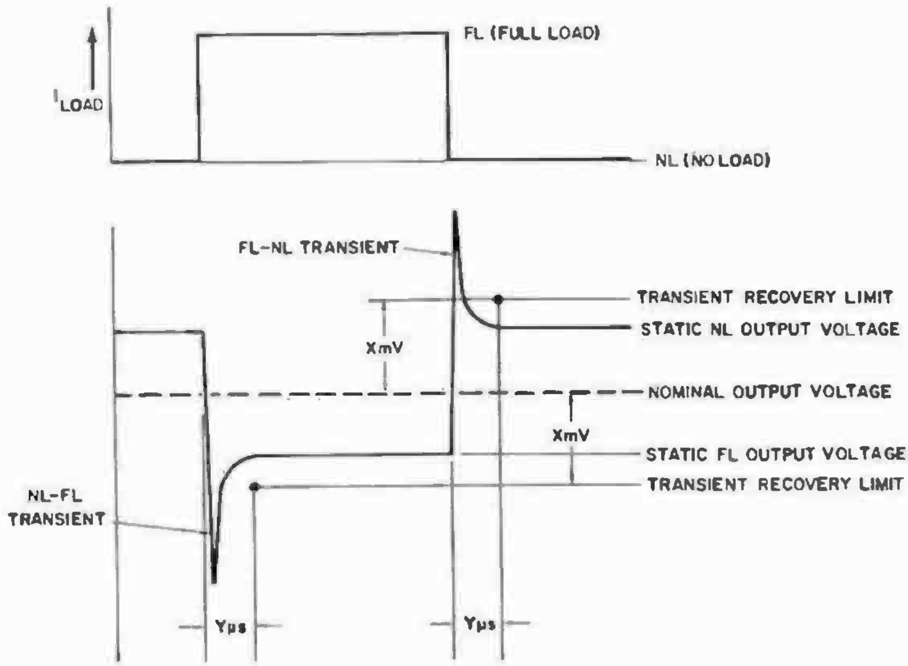


Fig. 21-115C. Expanded view of transient recovery time for a single square wave pulse. Suggested limits are indicated.

two extra wires between the supply and the load, this circuit permits the supply to achieve its *optimum regulation* at the load terminals, rather than at the power-supply output terminals. In this manner, the circuit compensates for the IR drop in the line from the power supply to the equipment receiving its voltage. The current through the sensing lines is quite small; therefore, the voltage drop is negligible.

21.117 Describe the design procedure for gaseous voltage-regulator tubes.—The theories of gaseous voltage-regulator tubes and of the zener diode are discussed in Questions 11.30 and 11.148, respectively. Therefore, only the circuitry generally employed with these devices will be discussed.

In Fig. 21-117A is shown the basic circuit for a gaseous voltage-regulator tube (VRT) connected across the output of a power-supply filter section. The input voltage to the regulator tube is designated E_1, the nominal or regulated output voltage, E_2. Variable quantities are I_1 the tube current, I_2 the load current, and the load R_L. As there are several variables to be considered in the design, several separate calculations are required. These calculations are based on the following conditions: The current through the tube must fall within the minimum and maximum current limits, and the supply voltage must be equal to or be greater than the maximum breakdown voltage of the VRT. The current through the tube may be expressed:

$$I_1 = \frac{E_1 - E_2}{R1} - I_2$$

The current through the tube varies directly with the input voltage and inversely with the load current. The current I_1 will therefore be maximum under the following conditions: when E_1

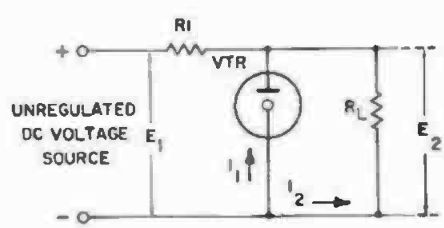


Fig. 21-117A. Basic circuit for designing a regulated voltage circuit employing a gaseous voltage-regulator tube (VRT).

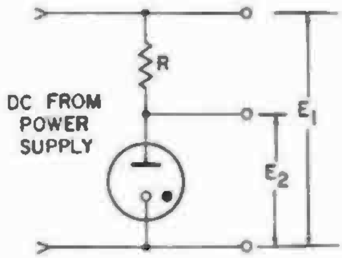


Fig. 21-117B. Voltage-regulator tube connected for use as a regulator and voltage divider.

is maximum, when the load current I_2 is minimum, and when the voltage E_2 is the minimum for this current. Thus, the lower limit of R1 is established as:

$$R1 = \frac{E_{1\,max} - E_{2\,min}}{I_{1\,max} + I_{2\,min}}$$

The minimum value of the current I_1 will occur under the following conditions: when input voltage E_1 is minimum, when the load current I_2 is maximum, and the tube voltage E_2 is maximum for this current. The upper limit for R1 is determined:

$$R1 = \frac{E_{1\,min} - E_{2\,max}}{I_{1\,min} + I_{2\,max}}$$

To assure that the tube will fire, the following conditions must also be satisfied:

$$\frac{E_{1\,min} \times R_L}{R_L + R1} > E_{breakdown}$$

In Fig. 21-117B is shown a VRT supplying two sources of voltage. Here, the tube functions as both a regulator and a voltage divider. Voltage E_1 is unregulated, while voltage E_2 is regulated. Two VRT's connected in series are shown in Fig. 21-117C. Both voltages E_1 and E_2 are in this instance regulated.

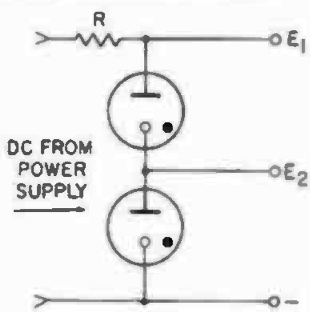


Fig. 21-117C. Two voltage-regulator tubes connected in series to provide two sources of regulated voltages, E_1 and E_2.

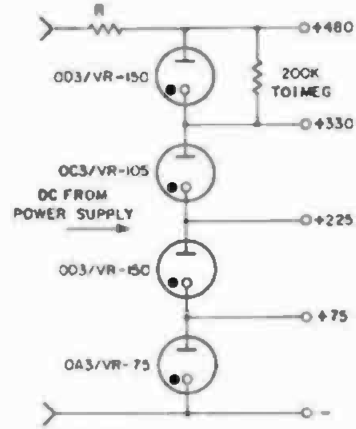


Fig. 21-117D. Four voltage-regulator tubes connected in series to provide four sources of regulated voltage.

A series string of four VRT's is shown in Fig. 21-117D. Each tube is of a different voltage rating, and voltage taps are taken at each tube. This connection is satisfactory only if each tube has the minimum and maximum current ratings. The starting voltage applied to each tube in the string will be determined by the individual leakage resistances. If a resistor of 200,000 ohms to 1 megohm is connected across one of the tubes, the remaining tubes will fire first, thus assuring the instant firing of all tubes when the voltage is applied. A slight reduction in the regulation effectiveness may be noted for the shnnted tube.

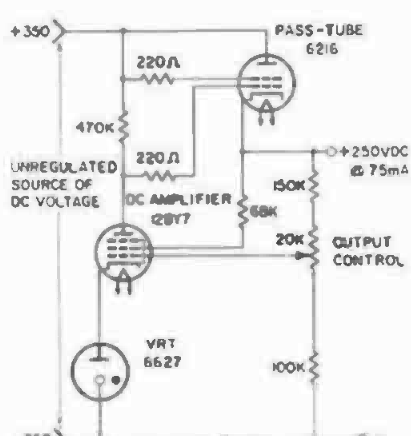


Fig. 21-117E. Voltage-reference tube connected in the cathode circuit of a dc amplifier for controlling the current through the pass tubes of a regulated voltage power supply.

Voltage-regulator tubes may also be connected in parallel to provide greater current capabilities, provided a small resistor of 10 to 100 ohms is connected in series with each tube. However, in doing so, the regulation is impaired; therefore, parallel operation is not recommended.

An advantage of using VRT's and zener diodes is their ability to decrease the ripple voltage or small pulsations in the output of an unregulated voltage source. Variations in the applied voltage cause current variations through the tube, and since the current through the tube is maintained at a *constant* value, the tube functions as a filter. A typical circuit, using a 6627 voltage reference tube, is given in Fig. 21-117E. Here, the tube is used to provide a fixed-bias voltage at the cathode of a 12BY7 tube controlling the current through pass tube 6216. Typical VRT's are the OD3/VR150, OC3/VR105, OA3/VR75, 6627/OB2, 6831, and the 6830.

An interesting aspect of the VRT is the effect of light on the tube, since its operation depends on ionization of gas (helium or argon). An ion is an atom which has lost or gained an electron due to some physical or chemical reaction. In the VRT, the gases are under heavy electrical stress; yet this alone is not always sufficient to cause the gas atom to ionize. Enclosing the VRT in a light-tight box requires a higher value of voltage to fire it, as compared to daylight. If shielded with an ⅛ inch of lead, it fires at even higher voltage, and in some instances, not at all. To overcome this characteristic, a trigger is required. This trigger may be in the form of a low-energy photon from an ordinary incandescent lamp, or from a high-energy cosmic ray. Without this help, the operation becomes erratic and the tube may not fire at all. Some types of VRT's contain a small amount of radioactive nickel placed on one element; therefore, they fire regardless if they are shielded or not. Voltage-reference tubes are stable and are operated in the low-current region. Tubes not designed for reference use are operated near their maximum current region. This stabilizes the tube, and prevents jumping of the interior glow with its attendant voltage shifts. The design procedure is the same for both type tubes.

Typical regulation characteristics are:

OA2	1.3 percent
OA3	6.7 percent
OB2	0.94 per cent (voltage reference)
OD3	2.6 percent
6627	0.10 percent

21.118 Describe the design procedure for a zener diode voltage-regulator circuit.—Referring to the basic design in Fig. 21-118A, the zener diode is connected in series with the limiting resistor R1 and in parallel with the source of voltage to be regulated. Assume that a 10-volt source of regulated dc voltage E_{out} is desired, at a maximum load current of 100 milliamperes. The unregulated voltage source E_s is to be 15 volts. As a rule, the zener diode current I_z is chosen for a value of 10 percent of the load current I_L, or for this example, 10 milliamperes. The value of the series resistance R1 can now be calculated:

$$R1 = \frac{E_s - E_{out}}{I_L + I_z} = \frac{15 - 10}{.100 + .010} = 45.5 \text{ ohms}$$

The power dissipated in R1 is I^2R, therefore:

$$P = (I_L + I_z)^2 \times 45.5 = 0.55 \text{ watt.}$$

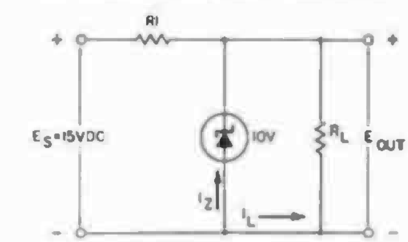


Fig. 21-118A. Basic circuit for designing a zener voltage regulator circuit.

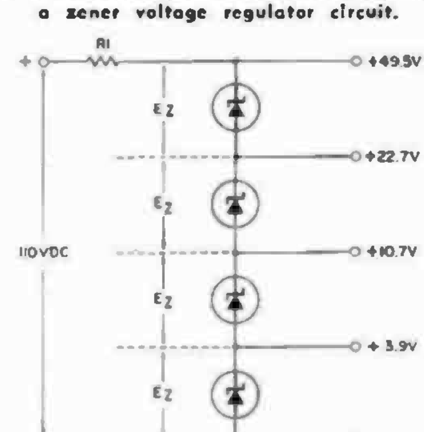


Fig. 21-118B. Zener diodes connected in series for regulation and voltage division circuits.

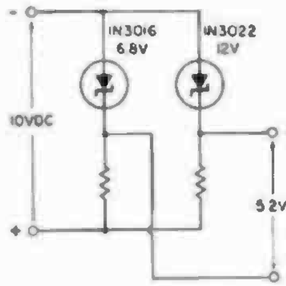


Fig. 21-118C. Zener voltage-regulator circuit (see text).

For practical purposes, a 47-ohm, 5-percent, 1-watt resistor is used. The power dissipated by the diode is equal to:

$$E_{out} \times I_z = 10 \times .01 = 0.10 \text{ watt.}$$

The above dissipation is only for a condition where the load current remains constant at 100 milliamperes. If the load current is completely removed, the current through the diode increases to 110 milliamperes and the zener wattage dissipation rises to 1.10 watts. In this instance, a diode capable of dissipating 2 watts would be in order. (See Question 11.148.)

Several voltage-regulating circuits developed by International Rectifier Corp., are shown in Figs. 21-118B to H. Fig. 21-118B shows a string of four diodes connected in series across the output of the regulated voltage. Diodes like the gaseous regulator tube can also be connected in series where the voltage drops are different, provided the power-handling capabilities and the current-operating ranges are similar. The circuit shown might be used for meter calibration for checking the scale linearity. Various voltage combinations are available by connection between the several voltage points.

The circuit in Fig. 21-118C is a low-voltage difference supply. This circuit can be used where the delivered voltage is lower than that normally available with zener diodes. Two diodes are used with the regulated potential difference being utilized. Because the temperature drift is the same for both diodes, the regulation is exceptionally good. Any combination of diodes can be used to achieve the desired output voltage.

Another source of well-regulated output voltage is obtained by the circuit shown in Fig. 21-118D. Here, the first diode acts as a preregulator. Any combination of diodes can be employed. A

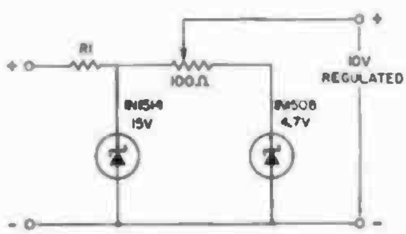


Fig. 21-118D. Zener voltage-regulator circuit (see text).

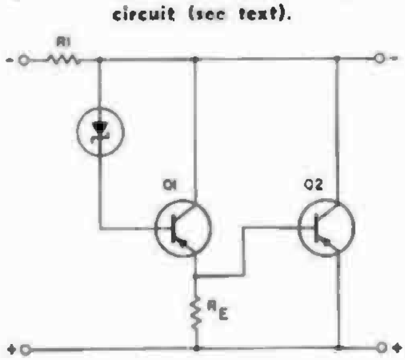


Fig. 21-118E. Cascade shunt zener diode voltage-regulator circuit. Transistor Q2 may consist of several units in parallel for higher current ratings.

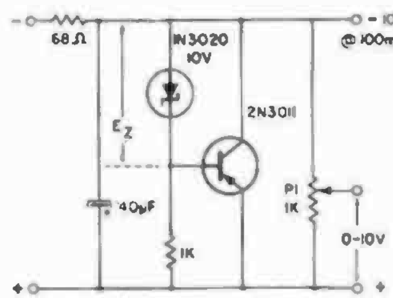


Fig. 21-118F. Zener voltage-regulator circuit where voltages lower than zener voltage are desired.

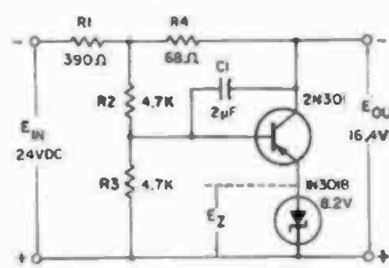


Fig. 21-118G. Zener voltage-regulator circuit for voltages greater than zener diode voltage.

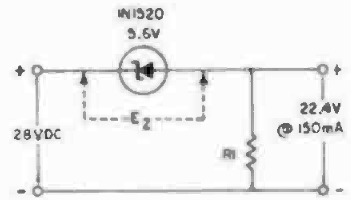


Fig. 21-118H. Series connection for a zener diode when only a small voltage drop is required.

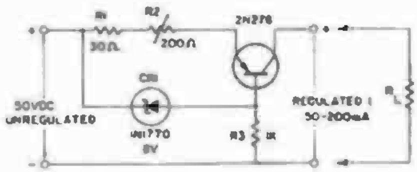


Fig. 21-118I. Current-regulator circuit using a transistor and zener diode.

cascade shunt regulator is given in Fig. 21-118E. The zener diode controls the base potential of transistor Q1, which functions as an emitter-follower and circuit amplifier. The voltage V_{eb} of Q1 determines the bias of Q2, the shunt regulator. This circuit is used where large current variations are encountered. Transistor Q2 may consist of several parallel-connected transistors.

Referring to Fig. 21-118G, shunt-type regulators can be designed to supply voltages lower or higher than the zener diode voltage E_z. Disregarding R1 and R2, the output voltage is determined by the ratio of R2/R3, therefore:

$$E_{out} = \frac{R2 + R3}{R3}$$

Thus, if R2 and R3 are of the same value, the output voltage will be twice the value of E_z. Resistor R4 compensates for variations in the supply to the regulator. The exact value is found by substituting a variable resistor for R4. The input voltage is varied, while R4 is adjusted for a minimum change in the output voltage E_{out}. Current amplification causes capacitor C1 to appear as a large value of capacitance across the output terminals. The ripple voltage is less than 10 millivolts when the input dc voltage E_{in} is supplied from a full-wave rectifier, using 20-μF capacitance across the output of the rectifier circuit.

When voltages lower than E_z are required, the circuit shown in Fig. 21-118F can be used. The transistor collector-emitter voltage is regulated at E_z. Potentiometer P1 should be as low as possible, yet compatible with the load requirements to minimize voltage variations, because of load current changes. If only a small voltage drop is required, say, 28 to 22.4 volts, the configuration in Fig. 21-118H might be employed. In this instance, the entire load current plus the current through R1 must flow through the diode and it could be easily damaged.

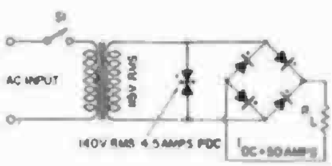
A current-regulator circuit is shown in Fig. 21-118I. Basically, the circuit consists of a grounded-base 2N278 transistor. As it will be observed, resistor R2 is variable. By changing the value of R2, the emitter current flowing through R3 is changed. Resistor R3 serves as a keep-alive current for the zener diode. With the load R_L, the transistor supplies a small portion of the current drawn by R3; therefore, less current will flow through the reference diode. Resistor R3, however, must draw enough current through the reference diode so the voltage drop across the diode remains at 8 volts as the current regulator is loaded. The load current remains essentially constant until R_L increases to where the average voltage drop across R_L is as large as the voltage drop across R3. The transistor must be mounted on a heat sink of approximately 165 square inches.

21.119 What are the voltage-operating ranges for zener diodes?—Zener diodes, ranging from 2 volts to around 200 volts and covering a large range of current operation, can be obtained for voltage regulation and reference use. In using higher-current types, the power dissipated by the diode must be given consideration. This latter information is taken from the manufacturer's data sheet.

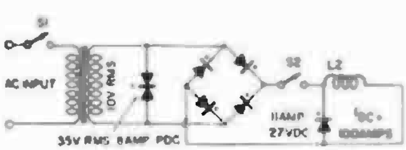
21.120 Describe transient suppressors and their functions.—Transient suppressors are solid-state devices used to protect silicon rectifiers, which are sensitive to voltage and overload conditions, for short time periods. Magnetic relays, transformers, and reactors generate transient voltage peaks far in excess of normal voltage, and can cause rectifier failure. The use of transient suppressors (or clippers) can increase the life and reliability of a power supply without affecting the circuit operation. Transient suppressors are manufactured in both polarized and nonpolarized units. Their characteristics are discussed in Question 11.158.

Polarized suppressors are selected by their rated dc-blocking voltage. For a 100-volt dc circuit, the proper suppressor would be rated 108 volts dc, or 180 volts peak. After the voltage rating of the suppressor has been determined, it is necessary to select a unit capable of dissipating the watt-seconds of the energy content of the inductive field. Suppressor-current ratings are therefore expressed in terms of peak discharge amperes. For a properly rated suppressor, the steady current through the suppressor is very low and does not add appreciably to the losses or temperature rise. Typical circuits for the application of suppressors are given in Fig. 21-120. To approximate the discharge current, a factor of 0.08 is used where the dc currents do not exceed 100 amperes, and a factor of 0.04 is used above 100 amperes. For a full-wave circuit of 125 volts at 30 amperes, the suppressor would be rated 140 volts rms, 200 volts peak, 2.5 amperes peak discharge current (PDC). Typical devices of this type are manufactured by Sarkes Tarzian, under the trade name of *Klipvolt*, and by International Rectifier Corp., as *Klip-Sels*.

21.121 Describe a series-type transistor voltage regulator.—In the series-type regulator (Fig. 21-121), regulation is accomplished by varying the current through three 2N3055 parallel-connected transistors in series with the load (these transistors are often referred to as pass transistors). Reverse-



(*a*) *Resistive or capacitive loads.*



(*b*) *Inductive loads.*

Fig. 21-120. Surge or transient suppressors connected for different type load conditions. The suppressors are of the nonpolarized type.

biased zener diode CR1 provides a source of reference voltage. The voltage drop across the diode remains at 12 volts, over a wide range of current.

If the output voltage V_{out} tends to rise, the total increase in voltage is distributed across resistors R8, R9, and R10. By setting potentiometer R9 to its midpoint, one-half the increase in output voltage is applied to the base of Q6. This increased voltage is coupled to the base of Q4 by resistor R5, the common-emitter resistor for Q4 and Q6. Reference diode CR1 and its series resistor R3 are connected in parallel with the bleeder resistors R8, R10, and potentiometer R9. The increase in output voltage is reflected across the diode resistor network. Since the voltage drop across the diode remains constant, the full increase in voltage is developed across R2 and is thus applied directly to the base of Q4. Because the increase in voltage at the base of Q4 is higher than that of the emitter, the collector current through Q4 increases.

As the collector current through Q4 increases, the base voltage of Q1 decreases by the amount of the increased drop across R1. The resultant decrease in current through Q1 causes a decrease in the emitter voltage of this transistor, and in the base of voltage of Q2. Similar action by Q2 results in a negative-going voltage at the base of the three pass transistors, Q3, Q5, and Q7.

As a result of this action, the current through these transistors and the load impedance in series with them decreases. The decrease in load current tends to reduce the voltage developed across the load circuit and cancels the original tendency for an increase in output voltage. Similarly, if the output voltage tends to decrease, the current through the pass transistors and through the load circuit increases; therefore, the output voltage remains constant.

The circuit shown is capable of regulating the load voltage within 0.5 percent for currents of zero to 10 amperes, with a voltage output of 22 to 30 volts.

21.122 Describe a shunt-type voltage regulator.—A shunt-type voltage-regulator circuit is not as efficient as the series-regulator circuit of Fig. 21-121; however, it does have advantages because of its simplicity. In the shunt regulator (Fig. 21-122), the current

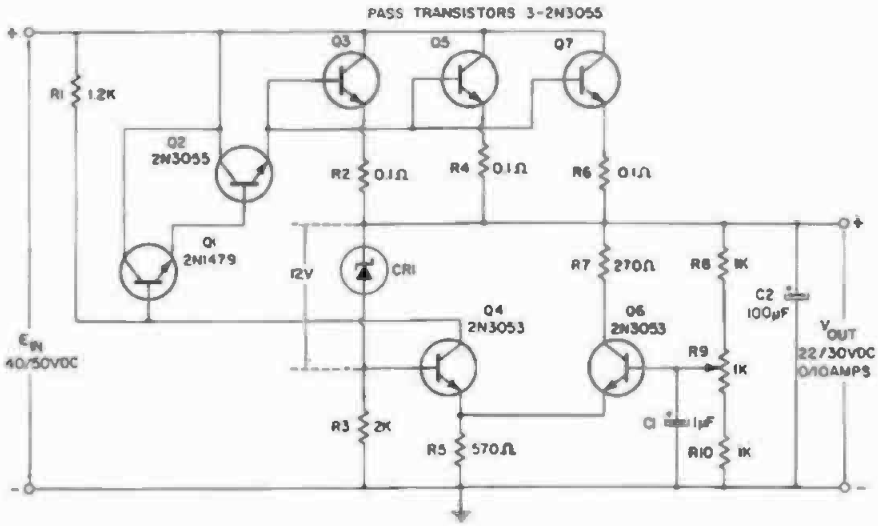


Fig. 21-121. Series type transistor voltage regulator. Load regulation 0.5 percent, input 1 percent. The output voltage is adjustable within 22 to 30 volts. (Courtesy, Radio Corp. of America)

through the shunt element transistors Q1 and Q2 varies with the load current of the input voltage. The current variations are reflected across resistor R1 in series with the load, so that output voltage V_{out} is maintained nearly constant. This regulator circuit can provide a regulated output voltage within plus or minus 0.5 percent at 28 Vdc, with a current capability of 500 milliamperes, for inputs of 45 to 55 Vdc. The zener diode is rated at 27 volts.

With a 28-volt output, the reverse-bias connected reference diode CR1 operates in the breakdown-voltage region. In this region, the voltage drop across the diode remains constant (27 volts) over a wide range of reverse currents through the diode. The output voltage will tend to rise with an increase in either the unregulated input voltage or the load-circuit impedance. Under these conditions, the current through resistor R2 and the reference diode increases. However, the voltage drop across the

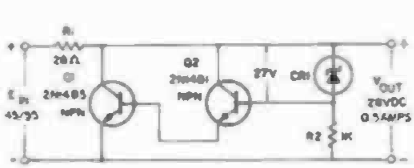


Fig. 21-122. Shunt-type transistor voltage regulator. Regulation is 0.5 percent. (Courtesy, Radio Corp. of America)

diode remains constant at 27 volts, and the increase in output voltage is developed across R2. The voltage drop across R2 is directly coupled to the base of Q2, thereby increasing the forward bias on Q1, and the current through this transistor increases.

Since the increased current of both transistors flows through R1 in series with the load impedance, the voltage drop across R1 becomes a larger proportion of the total applied voltage. In this manner, any tendency for the output voltage to increase is immediately reflected as an increased voltage drop across R1; thus, the output voltage remains constant.

Under a condition where the output voltage decreases, the voltage drop across the diode still remains constant, and the full decrease occurs across R2. This action results in a decrease in forward bias for both transistors; therefore, less current flows through R1. The resultant decrease in the proportional amount of the input voltage across R1 immediately cancels any tendency for a decrease in output voltage, thus resulting in a constant output voltage.

Resistor R1 is indicated as 28 ohms. Its value is to include the source resistance (transformer and rectifier). If 1.5-percent regulation is satisfactory, transistor Q1 may be omitted, and the emitter of Q2 returned directly to ground.

Resistor R2 is then changed to 50 ohms. Other circuit components remain the same.

21.123 Describe the circuitry of a transistor-regulated power supply.— Transistor-regulated power supplies are rather complicated in their design. They generally have regulation characteristics of 1 percent or less, and must function under many different conditions of use. Fig. 21-123A shows a model PBX regulated constant-voltage supply, manufactured by Kepco and rated 21 volts dc at 1 ampere, with 0.01-percent regulation. Six such units, rack-mounted, are shown in Fig. 21-123B, and the circuitry is shown in Fig. 21-123C and D. Basically, the device consists of four major sections—the main source with pass element and driver, error amplifier, comparison bridge, and the auxiliary voltage supplies.

Referring to Fig. 21-123D, the main power is derived from the lower winding of transformer T101 and is rectified by a conventional full-wave center-tap rectifier circuit consisting of diodes CR204 and CR205 and capacitor C203. This supply delivers the operating voltages for the pass elements, as well as the output current. The pass elements (series control transistors) Q101 and Q102 are in parallel, and are in series with the plus output terminal so that the unregulated voltage from the rectifier circuit is divided between them and the load. By changing the equivalent resistance of the series control element, the voltage drop across it is made to change in such a manner as to maintain a constant output voltage. The base drive current for the pass elements to effect this change is supplied by transistor Q203 which, in turn, is driven by the control signal from the error amplifier.

The error signal amplifier is dc-coupled and consists of Q208, Q207, and Q202. Its function is to amplify the signal from the comparison bridge to a level suitable for driving Q203. The comparison bridge is a four-arm bridge circuit and is the regulating and control element. It consists of a zener diode

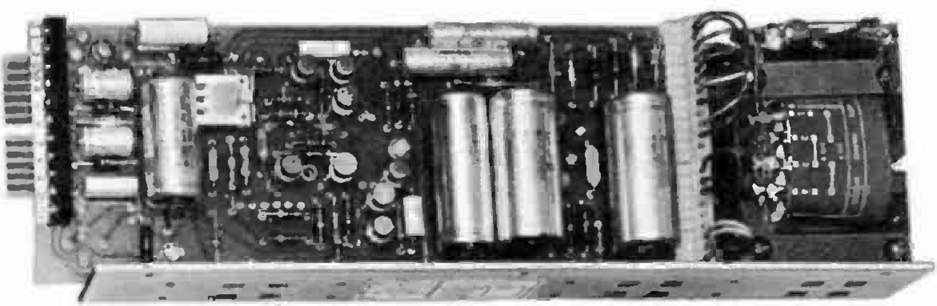


Fig. 21-123A. Interior view of Kepco Model PBX regulated power supply. The solid-state elements are all silicon.

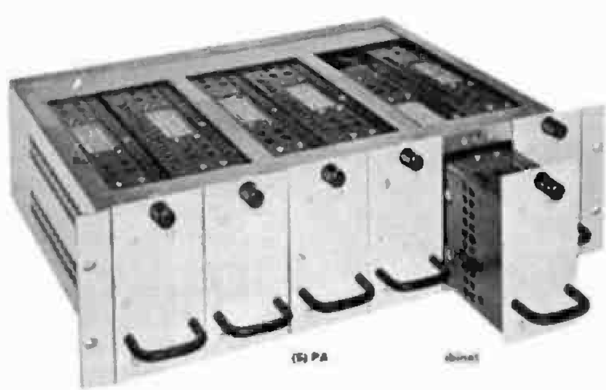


Fig. 21-123B. Six Kepco Inc. Model PBX power supply modules rack mounted.

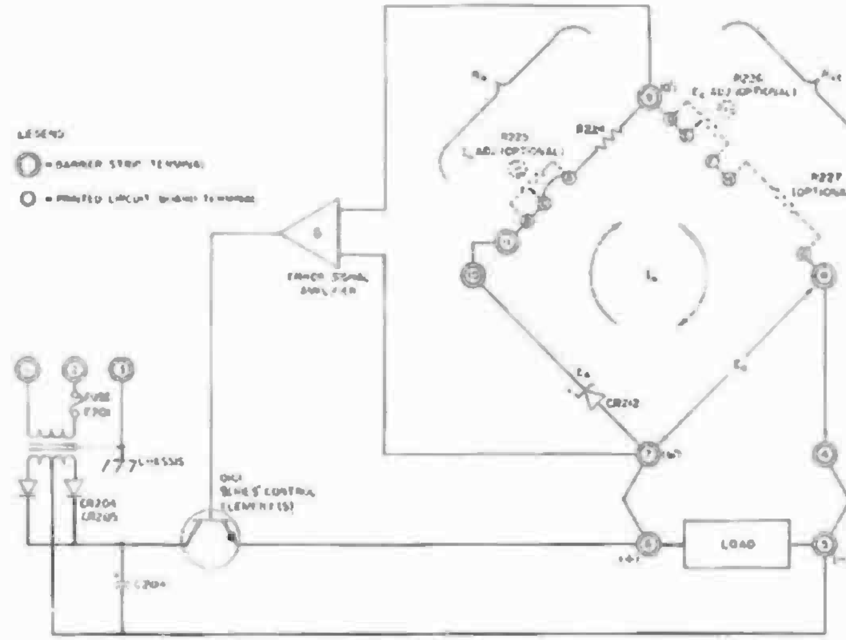


Fig. 21-123C. Basic diagram for Kepco 0.01-percent regulated power supply feedback-regulating system.

CR212 (V_R), resistor R224 (R_R), output voltage (E_O), and the voltage control resistance (R_{VC}), which is applied externally between terminals 8 and 9. As shown in Fig. 123C, a reference voltage (E_R) in series with the reference resistance (R_R) is continuously compared with output voltage (E_O) in series with voltage control (R_{VC}). At a condition of balance, a constant current (I_b) flows through the bridge, keeping the error signal at bridge terminals (a′) and (b′) at approximately zero volts. Any deviation from the preset output voltage will tend to change (I_b) in the sensing half of the bridge, and thereby will produce an error signal at the bridge terminals. This dc error signal is then amplified and acts as a control signal for the series-regulator transistor, changing the voltage drop across it to compensate for the change in output voltage.

The current-limiting circuit consists of Q205 and Q206, connected as a differential pair. The base of Q206 is referenced to the current-limit potentiometer R221, while the base of Q205 senses the voltage across current-sensing resistor R219. As long as the voltage drop across R219 (due to the output current) does not exceed the value selected by the setting of R221, transistor Q206 is conducting and keeping Q204 in a cutoff condition. If an overcurrent causes a rise of voltage across R219, the base of Q205 becomes more positive and Q205 conducts, while Q206 is driven toward cutoff. Consequently, Q204 will start conducting and pull Q202 and Q203 toward a direction tending to cut off the pass elements Q101 and Q102 (in parallel), thus sharply limiting the output current.

The amplifier power supply consists of a full-wave, filtered dc source comprising T101 (topwinding), silicon rectifiers CR201 and CR202, and filter capacitor C201. A series-regulator stage, followed by two zener diodes CR209 and CR210, provides operating voltages for both the amplifier and comparison bridge. Driver voltage is supplied by a half-wave rectifier CR203 and a capacitor C202, delivering the voltage to the collector of Q203. The I_{co} supply provides the turn-off bias voltage for the series pass elements, especially at higher ambient temperatures. The circuitry encompasses CR206, CR207, and capacitor C204.

Specifications for this supply are: an output voltage of 21 volts at 1 ampere, with a regulation of 0.01-percent output voltage change for line voltages of 105 to 125 volts; less than a 0.01-percent or 1-millivolt change for no load to full

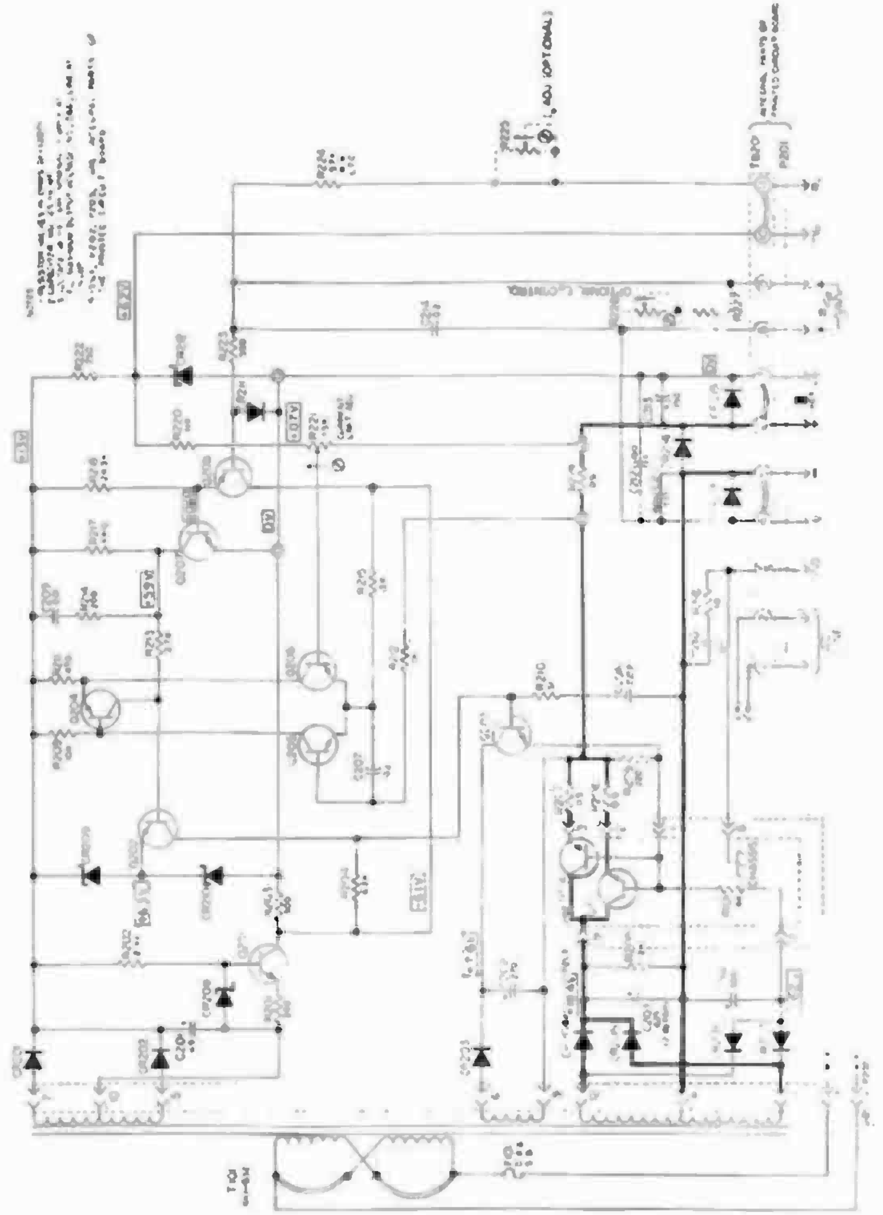


Fig. 21-123D. Schematic diagram for Kepco Model PBX 21-Vdc 1-ampere power supply having 0.01-percent regulation.

load change at any output voltage within the specified range; ripple less than 0.10 millivolt; an output voltage change of 0.01 percent over a period of eight hours; recovery time of 50 microseconds; ambient temperature operating range, minus 20 degrees to plus 65 degrees centigrade (case); and an isolation voltage, 500 volts between the chassis and either output terminal.

21.124 *Define the terms "open-loop gain," "closed-loop gain," and "loop gain," as associated with regulated power supplies.*—The terms "open-loop gain," "closed-loop gain," and "loop gain" are associated with the negative-feedback circuitry used for control purposes in regulated power supplies.

Open-loop gain is a measure of the gain without negative feedback, and is the ratio of the voltage at the supply terminals to the causative voltage required at the input of the null junction of the bridge circuit. (Causative voltage is the voltage required to cause the operational amplifier to function.) Open-loop regulated power supplies utilize voltage or current-sensing devices that automatically change the internal resistance of the power supply in such a manner that the load voltage or current remains constant. This may be accomplished by the use of either a gaseous voltage-regulator tube (VRT) or a zener diode.

Closed-loop (or operational) gain is the measure of the gain with negative feedback and is ratio of the voltage appearing at the output of the power supply to the causative voltage required at the input of a dc control amplifier. Closed-loop regulating circuits also employ an error-sensing circuit. In this instance, the output voltage (or current) is controlled by the error signal. This is the result of comparing a sample of the output voltage (or current) with a reference voltage. The reference voltage may be supplied by a zener diode or a VRT.

Loop gain is a measure of the negative feedback in a closed-loop system and is equal to the ratio of the open-loop gain to the closed-loop gain in decibels. The magnitude of the loop gain determines the error attenuation (control) and the performance of the control amplifier in the power supply regulating circuits.

21.125 Explain the basic design principles of a heat sink. — Heat sinks are used to radiate heat from solid-state rectifying devices. They are generally made from extruded aluminum or copper and are painted black, except for the areas in which the rectifying device is mounted. The size of heat sinks will vary with the amount of heat to be radiated, ambient temperature, and the maximum average forward current through the rectifying element. Several different types of heat sinks are pictured in Fig. 11-159.

The overall effectiveness of a heat sink is dependent to a great extent on the intimacy of the contact between the

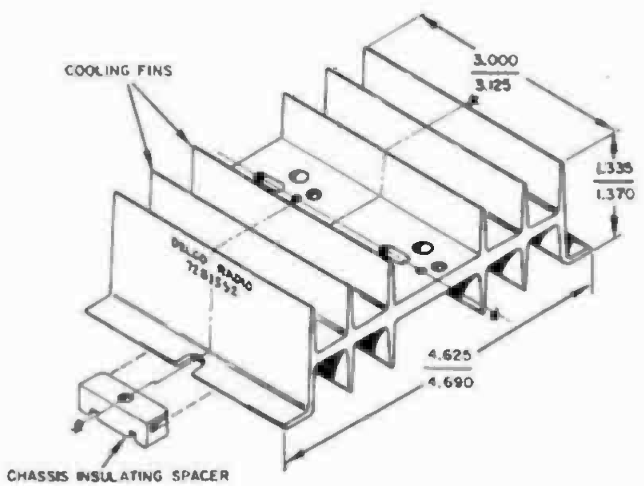


Fig. 21-125A. Typical heat sink for mounting two transistors. (Courtesy, Delco Radio Corp.)

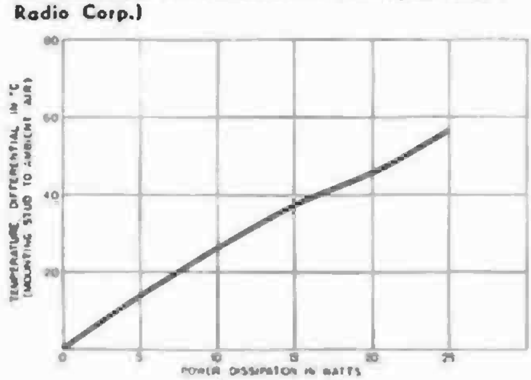


Fig. 21-125B. Thermal characteristics for the heat sink shown in Fig. 21-125A, with convection flow of air. (Courtesy, Delco Radio Corp.)

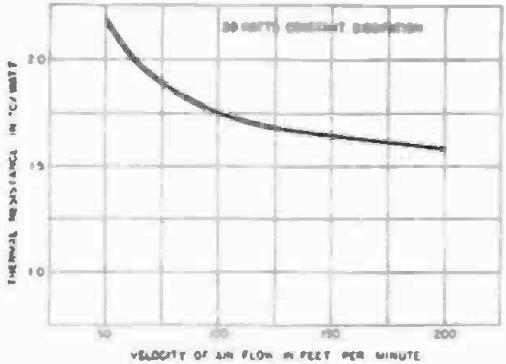


Fig. 21-125C. Thermal characteristics for the heat sink shown in Fig. 21-125A, with forced air currents. (Courtesy, Delco Radio Corp.)

device to be cooled and the surface of the sink. Intimacy between these two is a function of the degree of conformity between the two surfaces and the amount of pressure which holds them together. The application of a silicone oil (see Question 21.126) to the two surfaces will help to minimize any surface unevenness. However, the use of a mica washer between the base of the device to be cooled and the heat sink will add as much as 0.5 degree centigrade per watt to the thermal resistance of the combination. Therefore, it is recommended that (whenever possible) an insulating washer be used to insulate the entire heat sink from the chassis to which it is to be mounted. This permits the solid-state device to be mounted directly to the surface of the heat sink (without the mica washer). In this way, the thermal resistance of the mica washer is avoided. A typical heat sink manufactured by the

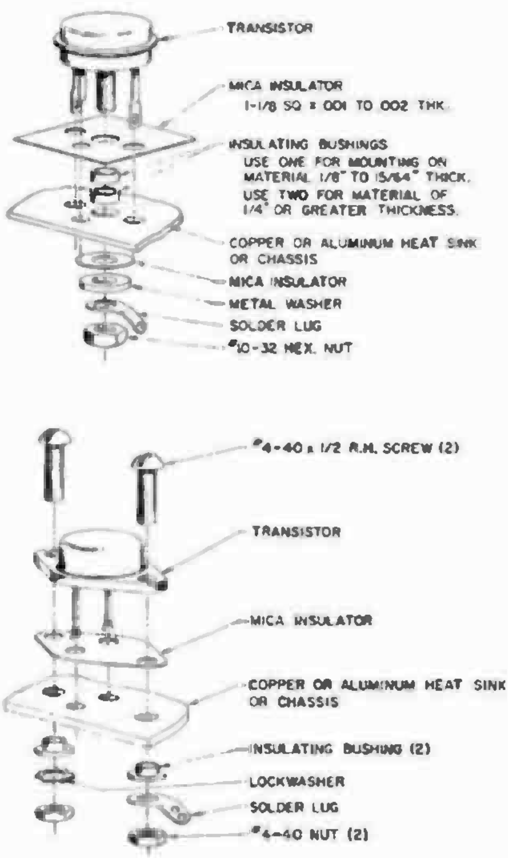


Fig. 21-125D. Transistor mounting kits for heat-sink operation. (Courtesy, Delco Radio Corp.)

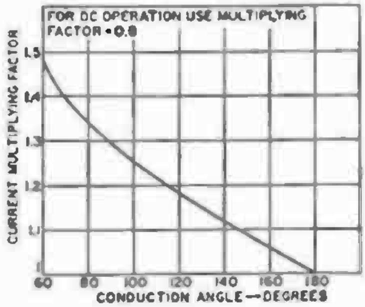


Fig. 21-125E. Current multiplying factor chart for heat-sink design. (Courtesy, Radio Corp. of America.)

Delco Radio Corp. is shown in Fig. 21-125A. This sink has 165 square inches of radiating surface.

The graph in Fig. 21-125B shows the thermal characteristics of a heat sink with a transistor mounted directly on its surface. A silicone oil is used to increase the heat transfer. This graph was made with the heat-sink fins in a vertical plane, with air flowing from convection only. Fig. 21-125C shows the effect of thermal resistance with forced air blown along the length of the fin.

Transistor mounting kits designed for type TO3 and TO36 cases are shown in Fig. 21-125D. However, if possible, it is always good practice to mount the solid-state device directly on the surface of the sink, then to insulate the sink from the supporting chassis (for reasons previously discussed).

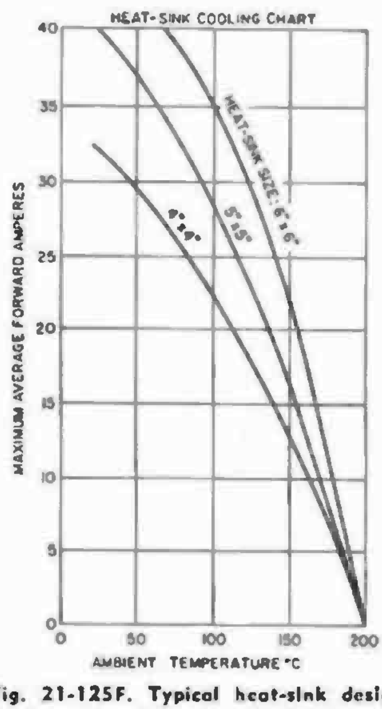


Fig. 21-125F. Typical heat-sink design cooling chart. (Courtesy, Radio Corp. of America.)

The size of the heat sink for a given application depends on the ambient temperature and the maximum average forward current of the rectifier element. The size must be calculated for each application. The calculation of size can be greatly simplified by the use of two charts: a current-multiplying chart shown in Fig. 21-125E and a heat-sink cooling chart shown in Fig. 21-125F. The first chart applies to all rectifier types for both polyphase and dc operation. The second chart differs for various types of rectifier types, and the calculation requires four steps.

From Fig. 21-125E, the current-multiplying factor is determined for the applicable conduction angle (the fraction of the ac input cycle during which forward current is expected to flow in the application). For dc operation of a silicon rectifier, a multiplying factor of 0.80 is generally specified. The desired output current (in amperes) is divided by the number of current paths. The actual number depends on the rectifier configuration and is determined from the table below.

Type of operation	*Number of current paths*
Single-phase, full-wave center-tap	2
Single-phase, full-wave bridge	2
Three-phase, wye-connected	3
Three-phase, double-wye	6
Three-phase, bridge	3
Six-phase, star	6

The resulting figure is the average forward current of the rectifier. The average current is then multiplied by a current-multiplying factor from Fig. 21-125E. The resulting figure now represents the adjusted average forward current of the rectifier.

This adjusted current is applied to Fig. 21-125F, to determine either the maximum or minimum allowable ambient temperature. The following example illustrates the method of calculation for the minimum heat-sink size for a three-phase, half-wave, wye-connected circuit. The conducting angle is 120 degrees, the output current is 90 amperes, and the ambient temperature is 90 degrees centigrade. Referring to Fig. 21-125E, the current-multiplying factor for a conduction angle of 120 degrees is 1.18. For three-phase, half-wave operation, the number of current paths is three; the average forward current therefore is 90/3 or 30 amperes. This average forward current is then multiplied by the current factor of 1.18 to provide an adjusted forward current of 35.4 amperes. From Fig. 21-125F, the minimum heat-sink size is found to be 6 × 6 inches square.

21.126 What type silicone fluid is used for seating transistors and diodes in heat sinks?—Several different types of silicone fluids are available for this purpose. Among them are Dow Corning Corp. Type-200, Wakefield Thermal compound, and CG Electronics Z5 Silicone compound. The fluid is applied between the base of the transistor and the surface of the heat sink, or if the sink is insulated, between the base and the mica washers. For diodes pressed into a heat sink, the silicone fluid is applied to the surface of the diode case before pressing it into the heat sink. The purpose of the silicone fluid is to provide a good heat seal between the two units. (See Question 21.125.)

21.127 What is the average emissivity for different materials used in heat sinks?—The thermal capacity of a cooling fin or heat sink must be large compared to the thermal capacity of the rectifier cell and have good thermal conductivity across its entire area. Since the surface conditions create differences in the emissivity, it becomes important to select or create a surface that provides the greatest emissivity. The average emissivity is expressed in watts per degree centigrade, per square inch. The average emissivity for different materials and surfaces is;

Aluminum anodized	0.8 (black)
Aluminum paint	0.50
Aluminum painted	0.9 (black)
Aluminum polished	0.05
Copper oxidized	0.7
Copper painted	0.9 (black)
Copper polished	0.05
Steel painted	0.9 (black)
Steel sheet	0.65

In the selection of a heat-sink material, thermal conductivity of the material must be considered. This determines the thickness required to eliminate thermal gradients and the resultant reduction in emissivity. An aluminum fin must be twice as thick as a comparable copper fin, and steel must be eight times as thick. (See Questions 21.125 and 21.126.)

21.128 Describe a solid-state variable low-voltage power supply.—A variable source of regulated low-voltage dc is quite useful for experimental work, particularly with solid-state devices. A solid-state, regulated low-voltage power supply, Model IP-20, developed by Heathkit, is shown in Fig. 21-128A. Referring to the simplified diagram in Fig. 21-128B, it will be noted that transistors Q4 and Q5 are connected in parallel and function as a series regulator. When a high current demand is presented to the output terminals, the regulated dc output voltage tends to decrease. As the voltage decreases, the voltage between the base and the emitter of Q2 decreases because the reference voltage remains constant. (Both the reference voltage and bias-voltage circuits have been shown as a battery to simplify the explanation.) This causes a reduction of current in Q2, causing the base current through Q3 to increase proportionally, because the current from the base supply is constant.

Fig. 21-128A. Heathkit Model IP-20 regulated power supply (solid-state).

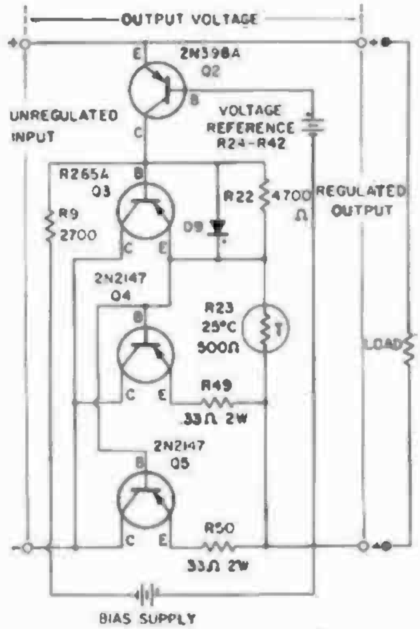


Fig. 21-128B. Basic circuit for Heathkit transistor regulated power supply Model IP-20.

As the emitter current in Q3 increases, the resultant increase in current flow appears in the base of Q4 and Q5. This increases the load current and restores the output voltage to its original value, that action occurring in the matter of a few microseconds. If the output voltage increases, the procedure is reversed, and the output voltage is held at a constant value. The purpose of diode D9 is to prevent transistor Q3 from being back-biased when the load current is removed.

Positive feedback is used to assist in the voltage regulation. Referring to Fig. 21-128C, assume the output voltage drops because of an increase in load current. This produces a voltage drop across resistor R16 of the polarities shown. This voltage drop aids in further reducing the current through transistor Q2 which, in turn, produces more out-

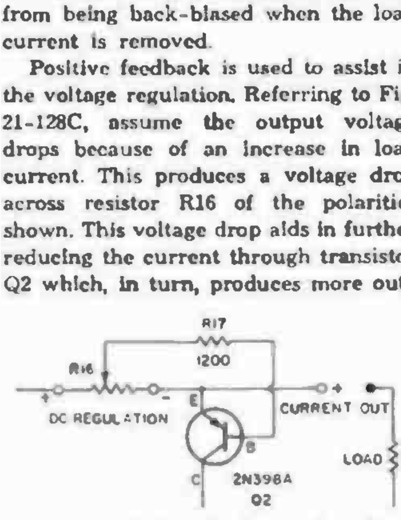


Fig. 21-128C. Feedback circuitry for a transistor regulated power supply.

put current. Thus, regulation is improved. When adjusting resistor R16 for a voltage of 15 volts at the output, from no-load to full-load, all other voltage settings will remain within 15 millivolts (0.015 volt), from no-load to full-load.

Referring to Fig. 21-128D, a fixed voltage reference source is obtained from a separate winding S2 on the power transformer, rectified by diode D8, filtered, then passed through a 6-watt, 120-volt ballast lamp (pilot light), and applied to V1, an OB2 gaseous voltage regulator tube, which supplies a reference voltage of plus 105 Vdc. The reference voltage is then applied to D1, a 56-volt zener diode, and a divider network on the voltage range switch, VR1. A small amount of voltage from V1 is also applied to the base of transistor Q2 through 2700-ohm resistor R9. For a fine control of the output voltage, control R42 provides a means of adjusting the voltage between the steps of the voltage range switch.

Transistor Q1 is a current limiter and normally operates in a saturated state because of the bias current from the current-limiter, diode, D6 (Fig. 21-128E). As the load current increases through the current-limiting control resistor R2, a voltage is developed across diode D6 and it starts to conduct. As soon as conduction starts, the emitter to base voltage of Q1 is fixed and limits the current to the value being delivered at the output. Capacitors C4 and C6 (Fig. 21-128D) reduce the internal output impedance to a value of 0.10 ohm at 10,000 Hz, and 0.5 ohm at frequencies above 10,000 Hz. The transient response is on the order of 25 milliseconds. The regulation for line voltage changes of 105 to 125 volts is less than 0.005 percent. Ripple and noise is approximately 150 microvolts. If desired, by proper adjustment of resistor R16, the output voltage may be set for a no-change condition, between a full-load and no-load condition.

With a heavy load current of a transient nature, such as might be demanded by an amplifier, the current-limiter tube Q1 tends to clip the peak of the current waveform being delivered to the load terminals. This characteristic may be reduced or eliminated by the connection of a 100-μF capacitor across the load terminals. This capacitor must

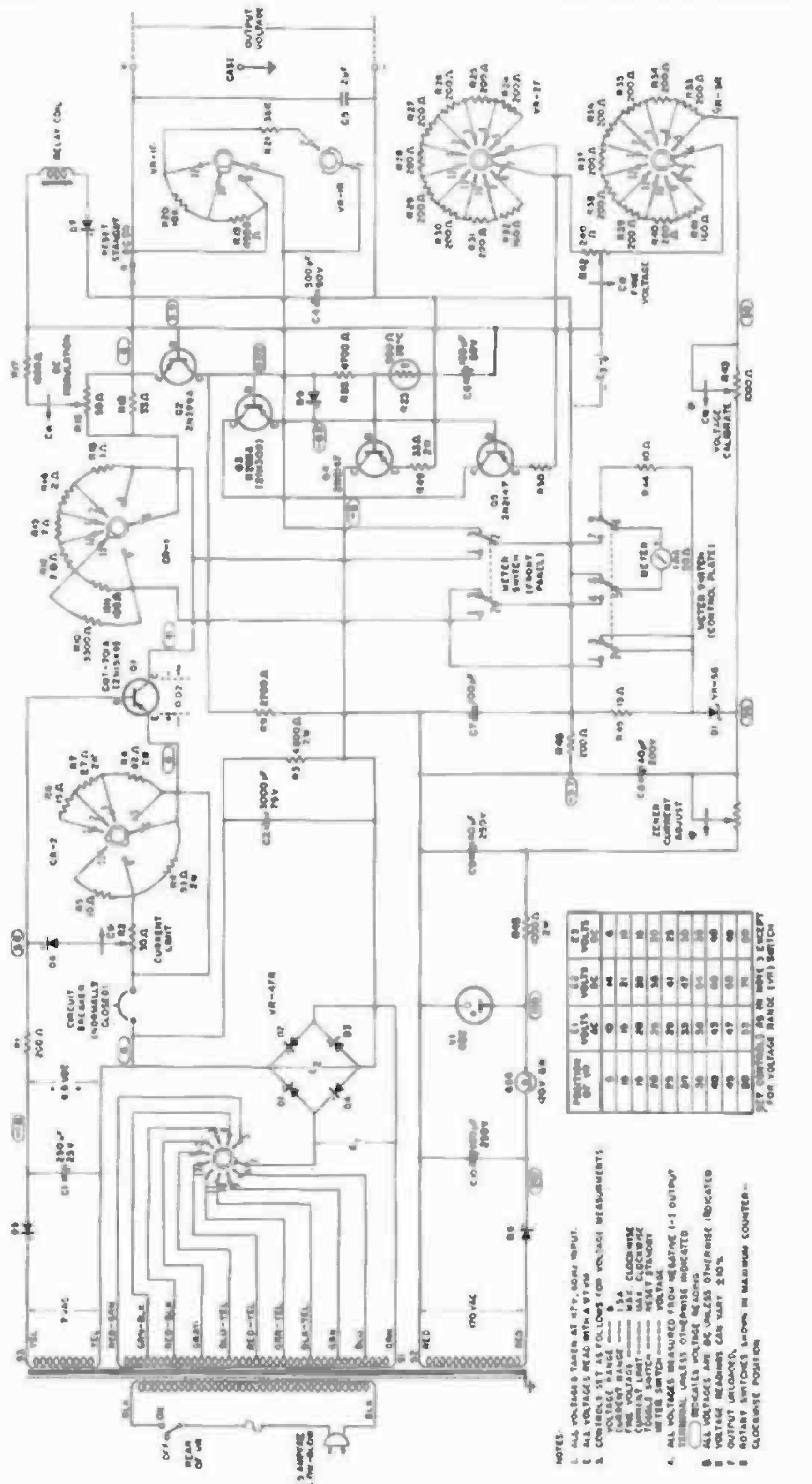


Fig. 21-128D. Schematic diagram for Heathkit Model IP-20 transistor regulated power supply.

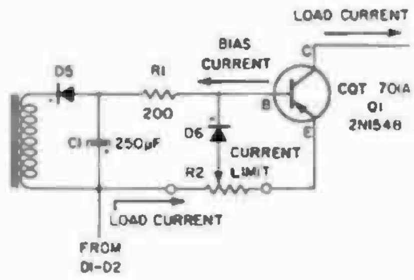


Fig. 21-128E. Current-limiting transistor and circuitry.

have low internal inductance and must be located as close to the output terminals as possible in order to reduce the inductive effect of the connecting leads. (See Question 21.115.) If the peak current demanded by the load does not exceed the current-limiting control setting, this characteristic will not be evident.

21.129 Give a circuit for constant voltage or current regulation, using a transistor.—Two basic circuits are given in Fig. 21-129. Referring to circuit A, the transistor is connected in a grounded-collector configuration, which means that the potential of the emitter is approximately equal to the base potential. If the emitter potential becomes more negative than the base, the transistor will be turned off, the current will drop, and the emitter potential then becomes more positive than the base, turning the transistor on. Therefore, the emitter potential will always remain almost equal to the base potential.

This effect is taken advantage of in the constant-current regulator in B except that a separate 1.5-volt supply is connected in series with a 1500-ohm resistor.

Assuming that a no-load condition exists between output terminals 1 and 2, and that no voltage drop exists between the emitter and the base of the transistor, the cnrrent is equal to:

$$I = \frac{1.5 \text{ volts}}{1500 \text{ ohms}} = 1 \text{ milliampere}$$

If a 500-ohm load is connected to terminals 1 and 2, the total resistance will be approximately 500 plus 1500 ohms, not taking into consideration the emitter-to-collector resistance.

Under the above conditions, a current in the outside loop will try to approach 6 mA. As soon as the current increases to more than 1 mA, the IR drop across the 1500-ohm resistor will become greater than 1.5 volts. The emitter then becomes negative with respect to the base, and the transistor is turned off. By substituting a variable resistor for the 1500-ohm resistor, the current in the load can be controlled (as long as the current rating of the transistor is not exceeded).

21.130 Describe a transistor motor-speed control unit.—The circuit shown in Fig. 21-130 can be used to provide motor-speed control and regulation under changing loads for both ac and dc universal motors. The motor should have a current rating of not more than 2 amperes using a 2N3228 SCR, or up to 12.5 amperes using a 2N3669 SCR. The motor speed will be controlled from cutoff to its full rated speed. The circuit provides smooth skip-free operation at reduced speeds.

The speed of the motor is determined by the time that the SCR conducts during each half-cycle of the ac input signal. The time cycle is controlled by the manual adjustment of variable resistor R2. At minimum resistance, the rectified current from the four 1N2860 diode rectifiers charges capacitor C1 rapidly to the triggering point of the

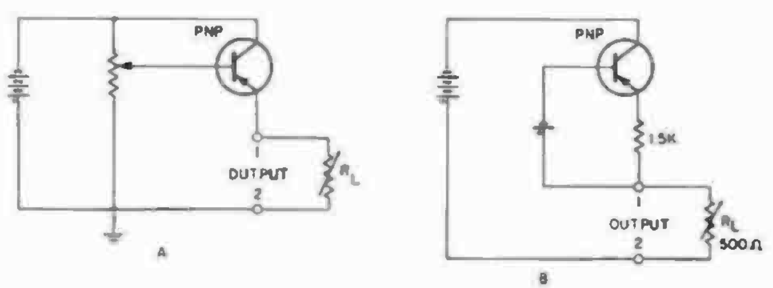


Fig. 21-129. Transistor circuits for constant-current control.

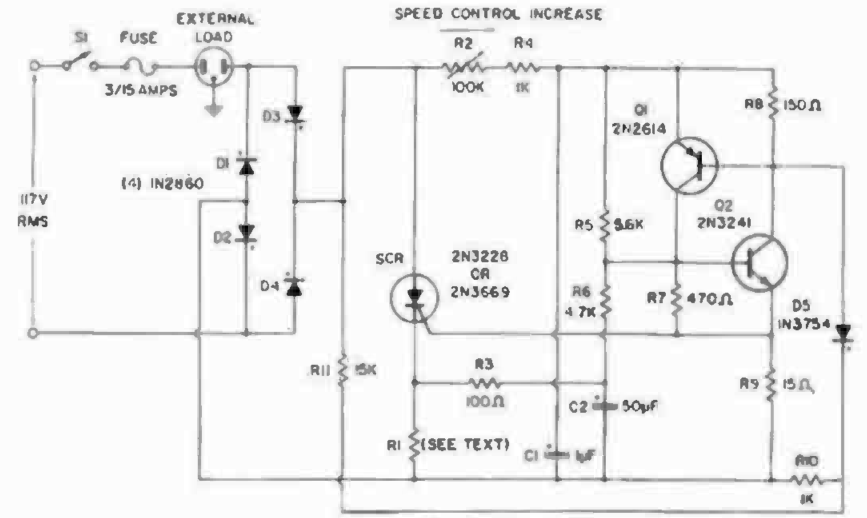


Fig. 21-130. *Universal transistor motor-control circuit. This circuit is capable of controlling load currents up to 12.5 amperes. (Courtesy, Radio Corp. of America.)*

transistor regenerative-switch, Q1 and Q2 (the circuit shown is preset for 6 volts). The switch is triggered into conduction early in each input half-cycle. When Q1 and Q2 conduct, capacitor C1 discharges through the series circuit of the transistors and the gate electrode of the SCR. This discharge current triggers the SCR into conduction, and the load current then flows until the end of the input half-cycle. This operation repeats for each succeeding half-cycle of the ac input signal, and the motor speed is maintained at maximum. When increasing the value of R2, capacitor C1 charges more slowly and the SCR is triggered later in the input half-cycle, or not at all if the charge on C1 falls short of 6 volts. Thus, the speed of the motor is reduced to cut-off completely.

A feedback circuit consisting of resistors R1, R2, R6, and capacitor C2 maintains the speed of the motor essentially constant under changing load conditions. When a load is applied to the motor, the speed momentarily decreases and the current through the motor and the SCR increases. Resistor R1 in series with the SCR develops an increased voltage drop, and the charge in capacitor C2 is increased. The increased charge produces a current increase through R6, and less current is then required through R5 and the regenerative transistor switch. As a result, the SCR is triggered earlier in the next half-cycle of the input ac voltage. The increased conduction time results in a corresponding increase in motor speed. Resistor R9 performs an additional function by shunting out commutator noise and eliminates premature triggering of the SCR.

The circuit can also be used for the control of lighting circuits, drill motors and other equipment, with a capability of 240 watts using the 2N3228 SCR, and 1500 watts using the 2N3669 SCR. For lighting-control use, resistors R3, R6, and capacitor C2 are not required. The circuit operation is essentially the same. The value of resistor R1 is obtained by dividing 2 volts by the load current. The wattage is then calculated and a 50-percent safety factor is added.

21.131 Describe a nonelectronic bridge circuit for constant-voltage or constant-current operation. — Fig. 21-131A shows a simplified circuit that may be used as a source of constant voltage or constant current. It was devised by James Spencer of Westinghouse Electric and Manufacturing Co. The basic circuit is the well-known Wheatstone bridge, consisting of two noninductive resistors R1 and R2, and several 25-watt, 117-volt lamps. The internal resistance of the lamps will vary, while the resistors R1 and R2 will remain constant. The voltage at the output of the bridge circuit will remain constant over a range of 75 to 130 volts ac at the input, provided the load current remains constant at the output of the bridge.

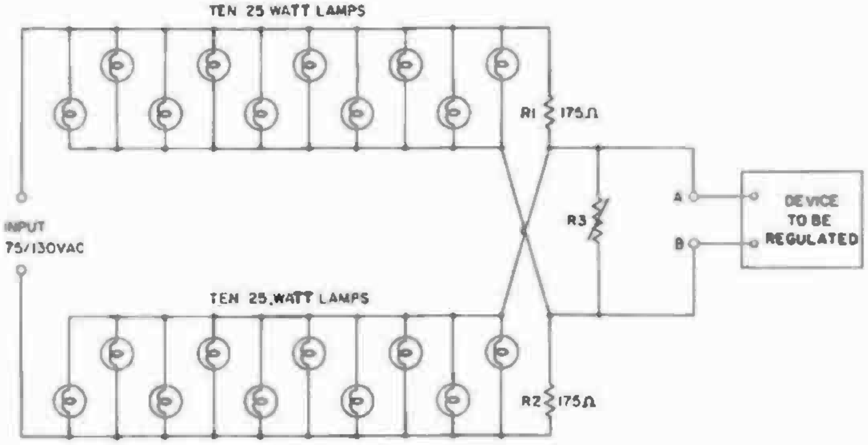


Fig. 21-131A. Constant voltage- or current-regulating circuit (after Spencer).

If the load current is fairly constant, the circuit is also a good voltage regulator and may be substituted for a constant-voltage transformer. If the voltage at the output of the bridge is high, it may be reduced by loading the output side with a heavy wattage resistor R3, or by imposing an autotransformer between the output and the input of the device being regulated.

This circuit was originally developed for the calibration of ac voltmeters and ammeters. In the instance of ammeter calibrating (Fig. 21-131B), a 20:1 current ratio transformer is connected at the output, and a standard meter connected in parallel with the meter under calibration. For voltmeter calibration, (Fig. 21-131C), a transformer of the proper voltage rating is used with a standard meter in parallel with the meter to be calibrated, and a load resistor of such value to load the transformer to about 20 percent of its rated current-carrying capacity.

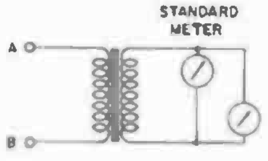


Fig. 21-131B. External circuit for calibrating an ammeter.

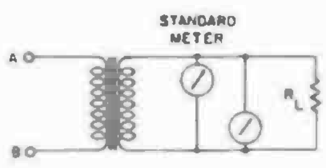


Fig. 21-131C. External circuit for calibrating voltmeters.

A second lamp bridge devised by Kelly, is given in Fig. 21-131D and is designed to supply a constant output of 1 volt, with 0.25-percent regulation for line-voltage variations of 105 to 125 volts, over a frequency range of 25 to 800 Hz. Looking into the output terminals, an internal impedance of 60 ohms is seen. When maximum output is required, the 500-ohm resistance R1 is replaced by the external load. The three lamps are No. 48, 2-volt, 60-mA pilot lamps.

21.132 How is leakage current between a power supply and ground measured?—By connecting a 1000-ohm resistor between the chassis and a good water-pipe ground as shown in Fig. 21-132. With the ac power plugged in and turned on, the voltage drop across the resistor is measured using a sensitive vacuum-tube voltmeter. The current through the resistor is calculated by simple Ohm's law, $I = E/R$. This

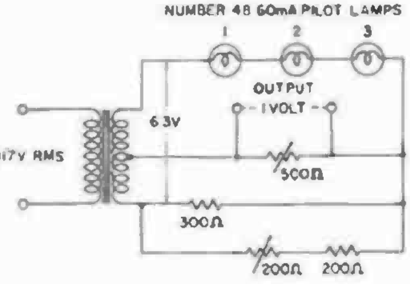


Fig. 21-131D. Small voltage lamp bridge, for a constant 1-volt output (after Kelly).

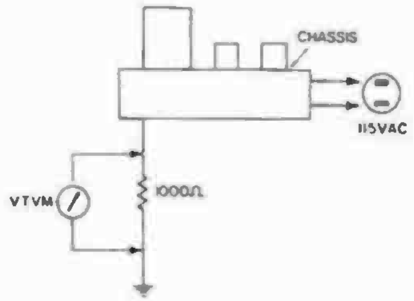


Fig. 21-132. Measurement of leakage current between an amplifier chassis and ground.

measurement must be made by measuring the leakage for both directions of the plug in the 115-volt outlet.

Normally, the leakage current between the chassis and ground will amount to only a few microamperes. If the leakage current runs to several milliamperes, it is quite possible for the chassis, under certain conditions, to become dangerous.

21.133 Show the schematic diagram for a simple vacuum-tube variable-voltage supply.—The circuit given in Fig. 21-133 is quite simple and is capable of delivering 150 milliamperes at an output voltage of 300 volts dc, with a fairly low ripple voltage. The two 6L6 tubes are used as rectifiers, with the control grids returned to a potentiometer connected across the dc output terminals. To control the level of the output voltage, the bias voltage on the 6L6 tubes is varied by potentiometer P1. The voltage may be varied from zero to full output. Any type tubes similar to the 6L6 may be used, provided that they will pass 150 milliamperes or more of current. This circuit is not a regulated one and is subject to the ills of unregulated power supplies.

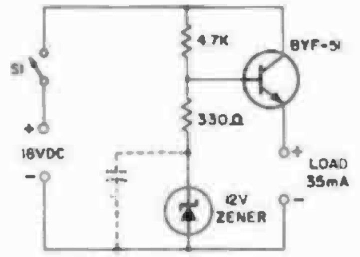


Fig. 21-134. Solid-state voltage regulator suitable for use with batteries.

21.134 Describe a solid-state voltage regulator suitable for battery regulation.—When a battery voltage is not the exact value required, this difficulty may be overcome by the use of the circuit given in Fig. 21-134 using a battery voltage somewhat greater than that required.

The arrangement utilizes only four components and will, as shown, deliver 12 volts from an 18-volt battery supply. The internal output impedance of the circuit is comparable to a new 12-volt battery. To prevent damage to the transistor when used with a resistive load having a large value of capacitance in parallel, a second capacitor of higher value than the first is connected between the collector and emitter elements of the transistor. The two capacitors then charge in series, without drawing a heavy charge current through the transistor.

When the circuit is used with a continuous charging circuit (trickle-charger), a certain amount of ripple voltage is always present. In this instance, a low value of capacitance is connected in parallel with the zener diode. Switching the batteries off discharges the capacitor through the base-emitter circuit in series with the load.

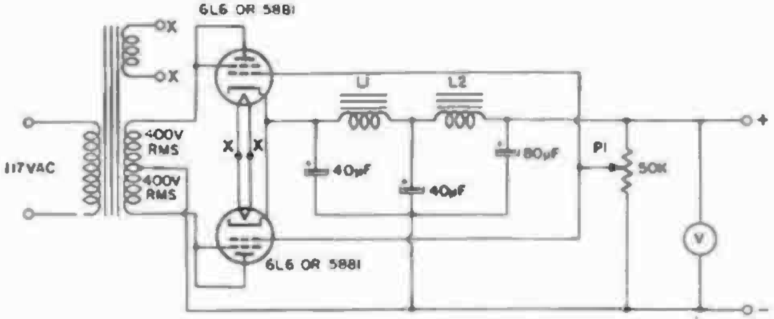


Fig. 21-133. Variable high-voltage supply using a bias control on the rectifier tubes.

When the batteries are switched on, the output voltage slowly builds up at the same charging rate of the second capacitor.

Overload protection is afforded by increasing the value of resistors R1 and R2. During overload, the base-emitter current increases the voltage drop across R1. Thus, the base voltage drops below zener cutoff. The circuit no longer regulates, but prevents an excessively high I_c.

The circuit shown is not confined entirely to batteries or the voltages mentioned, as any voltage output can be obtained by the selection of the proper diode. Such a circuit can be quite useful for supplying a 9-volt portable radio from a 12-volt car battery.

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Section 22

Test Equipment

The greatest advancement in test equipment has come about through the use of semiconductor circnitry. Many large, cumbersome instruments, heretofore limited to laboratory operation, are now available as compact portable devices. Although the basic circuitry of test equipment has not changed in principle, graphic recorders and computer techniques have added speed and accuracy to their operation. A wider scope of performance is available with multiple and complex operations included in one instrument, such as the multiple-trace oscilloscope with both long- or short-persistence CRT's, and with bandwidths of 50 megahertz and greater. Test instruments, such as random-noise (white-noise) generators, wave analyzers, distortion-factor meters, and phase and octave-band analyzers, are some of the items covered in this section.

22.1 What is a thermocouple element and how is it used?—A thermocouple element consists of two dissimilar metals, such as copper and constantan wire, joined together to form a junction. The thermojunction is placed in contact with a heater wire as shown in Fig. 22-1A. When used, the heater wire is connected in series with the source of current to be measured. The current in passing through the heater wire generates heat (watts eqnal I^2R) which is transmitted to the thermojunction. The application of heat to the junction causes it to generate a small dc voltage which is measured by a sensitive dc measnring instrument, such as a millivoltmeter. Within the limits of the thermocouple element, the greater the current throngh the heater element, the greater the dc voltage generated by the thermojunction. (See Fig. 22-1B.)

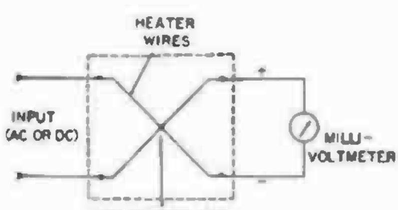


Fig. 22-1A. Method of connecting a thermocouple element to a dc meter movement.

The quantity of heat generated by the heater element will be exactly the same regardless of whether the sonrce current is alternating or direct. Waveform complexity and frequency do not enter into the measurement. This makes the thermocouple ideal for measurements of frequencies up to several megahertz.

Thermocouples may be calibrated accurately by the use of dc voltages and will hold their calibrations indefinitely. They may be used to measure either alternating or direct current. A typical

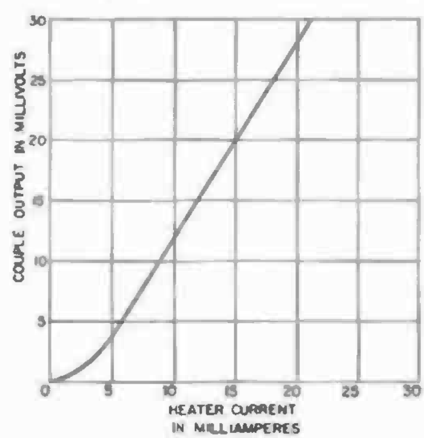


Fig. 22-1B. Typical calibration curve of a thermocouple element, showing the relationship between heater current and output in millivolts.

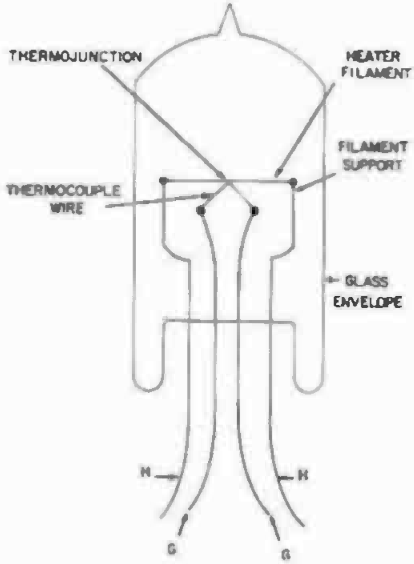


Fig. 22-1C. Cross-sectional view of a typical thermocouple element in a glass envelope.

thermocouple element, enclosed in a vacuum-tube glass envelope, is shown in Fig. 22-1C. Thermocouple elements may be obtained having a variety of internal resistances. Typical values range from 0.30 ohm to 1100 ohms, with the junction side ranging from 3 to 50 ohms. A typical example is the Western Electric thermocouple Type J, which has a heater resistance of 600 ohms and a couple resistance of 12 ohms. To produce an open-circuit voltage at the couple terminals of 0.005 to 0.015 volt requires a current through the heater element of 0.002 to 0.005 ampere, respectively.

22.2 ***Why are the calibrations on a thermocouple meter crowded at one end?*** — Because thermocouple meters and hot-wire meters both respond to the power in the circuit, although they may be calibrated in voltage, which varies with the square of the power passing through the thermocouple element.

22.3 ***What is a hot-wire ammeter?*** —An indicating meter which depends on its operating force being generated by the current in the circuit under measurement flowing through a hot wire. The internal construction of a typical hot-wire ammeter used for measuring high-frequency currents is shown in Fig. 22-3. The current in passing through hot wire A in the meter

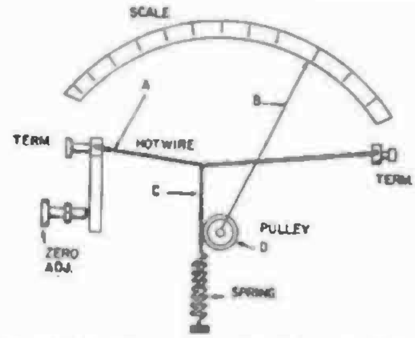


Fig. 22-3. Internal construction of a hot-wire ammeter for measuring high-frequency currents.

causes the wire to expand and contract. Pointer B of the meter movement is caused to move by means of a second wire C connected to the hot wire and wrapped around a small pulley D to which the pointer is attached. As the hot wire expands and contracts, the pointer is caused to move across the scale by the rotation of the pulley. The heat generated in the hot wire is proportional to the current passing through the wire.

22.4 ***Are hot-wire or thermocouple meters affected by phase shift in a complex waveform?***—No, they are not affected by phase shift or waveform.

22.5 ***What is a dynamometer-type meter movement?*** — A dynamometer meter movement is based on the principle of magnetic repulsion between two or more electromagnetic fields. When two coils are used, one is fixed and one is made movable. The meter hand or pointer is attached to the movable coil. (See Fig. 22-5.) When

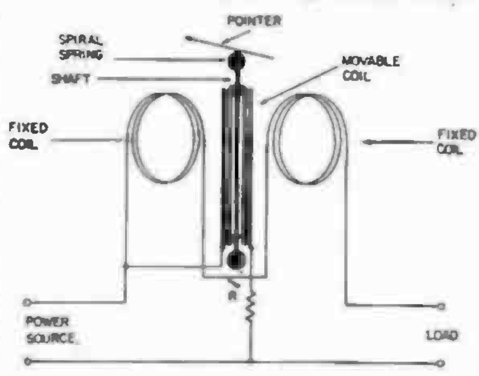


Fig. 22-5. Dynamometer-type meter movement used in a wattmeter. Fixed coils carry the load current and movable coils are connected in parallel with the line.

current is passed through the coils, magnetic fields are created. The movable coils are repelled by the fixed coils when the magnetic fields are of the same polarity.

Dynamometer meter movements are used for the measurement of ac and dc current, voltage, and power and are the type of movement most commonly used in the design of wattmeters. When one is used as a wattmeter, two fixed coils of high resistance and many turns are connected in series across the voltage source. The movable coil is of low resistance and is connected in series with the power source. The movable coil swings a distance (carrying the pointer) proportional to the combined strength of the two magnetic fields. Therefore the movement is proportional to both the current and voltage of the circuit. Since the power in the circuit is proportional to the current and the voltage, the meter can be calibrated to read directly in watts.

22.6 *What is an electrostatic voltmeter?* —A meter movement used for the measurement of very high voltages which draw very little current. The movement consists of two plates of a variable capacitor, one of which is movable and carries a pointer.

When two plates of a capacitor are charged electrically, a mechanical force, called an *electrostatic force*, is created between the two plates. If the charge on the plates is opposite in sign, the plates are attracted to each other; if of the same sign, they are forced apart. Advantage is taken of this force to move the free plate and the pointer.

The electrostatic force created by the charge may be calculated:

$$\text{Force (oz)} = \frac{1.59 \times K \times A \times E^2}{L^2 \times 10^{11}}$$

where,

K is the dielectric constant,
A is the area of the plates in centimeters,
L is the distance between the plates in centimeters,
E is the potential between the plates in volts.

Electrostatic voltmeters may be used on either alternating or direct current.

22.7 *Describe the construction of a taut-band meter movement.* —Since the introduction of the original D'Arsonval meter movement, electrical indicating and measuring instruments have generally been restricted to the pivot-jewel and hair-spring suspensions, with the exception of the string galvanometer movement. With the advent of many new electronic devices requiring the use of meters, the need for a more sensitive movement as can be supplied by the hair-spring type is required. A

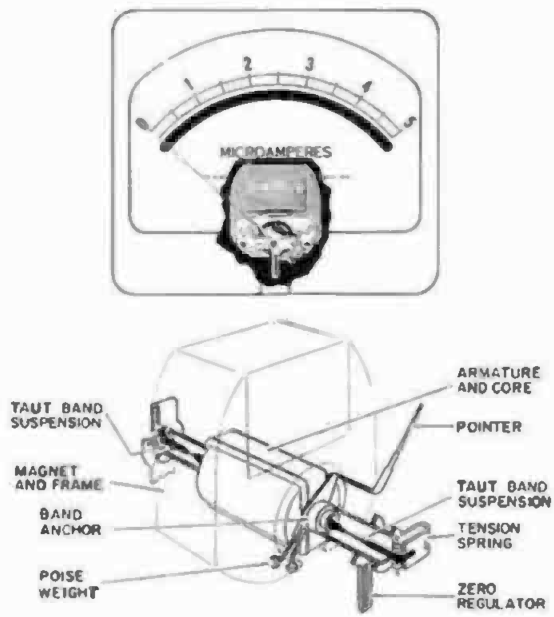


Fig. 22-7. Interior view of a taut-band meter movement, and a cross-sectional view of its construction. (Courtesy, Stark Electronic Instruments, Canada)

movement utilizing a taunt-band or strip principle has become practical by the use of a new alloy in the form of a thin metal strip supporting the movement, carrying the current, and providing the restoring torque.

The suspension concept eliminates the bearing pivot hair spring, with its attendant reduction in sensitivity. Taut-band suspension permits sensitivities of ½ microampere (0.0000005 amp) for 100 angular degrees of pointer deflection, with a power consumption of only 0.00000001 watt. Taut-band suspension may be employed for the construction of ac and dc meter movements, both iron-vane and dynamometer types.

Such movements are particularly useful where high impacts are encountered. Damping can be controlled to a much finer degree than heretofore. With these movements it is possible to measure ac voltage and current at frequencies between 0.6 and 5 Hz through a simple damping arrangement. In Fig. 22-7 are shown two views of the movement mounted in the meter case, and a cross-sectional drawing of its construction, as manufactured by Stark Electronic Instruments (Canada).

22.8 ***What is a meter shunt?*** — A low resistance connected in parallel with a meter movement for the purpose of measuring current. An ammeter generally consists of a sensitive meter movement drawing only a few milliamperes of current. By connecting a shunt of proper resistance in parallel with the meter movement, large amounts of current can be measured. Ammeter shunts range from a few thousandths of an ohm to several ohms in value, the exact value depending on internal resistance of the meter movement.

The value of a shunt may be calculated:

$$R_s = \frac{R_m \times I_m}{I - I_m}$$

where,

I_m is the normal full-scale current range of the meter,
I is the new range of current,
R_m is the internal resistance of the meter movement,
R_s is the value of the shunt resistance.

A typical meter shunt is shown in Fig. 22-17.

22.9 ***What is a meter multiplier?***— A resistance connected in series with a meter movement to permit it to be used for the measurement of voltage. Generally, a multiplier resistance has a value of many hunderds of times the internal resistance of the meter movement. As an example, a voltmeter rated at 1000 ohms per volt employs a 1-milliampere movement, with 1000 ohms connected in series with the movement for every volt required for its full-scale deflection. Thus, a 1000 ohms per volt meter with a full scale of 10 volts has a 10,000-ohm resistor in series with the movement.

The value of the multiplier resistor may now be calculated:

$$R_{mult} = R_m (N - 1)$$

where,

R_m is the internal resistance of the meter movement,
N is the new full-scale range divided by the original full-scale reading of the movement.

22.10 ***Describe a quasi-rms responding meter.***—It is a metering circuit used by General Radio Co. in their sound-level and octave-band noise analyzers for reading the output signal. Prior to the development of this meter circuit, sound-level meters employed a meter movement driven from a full-wave rectifier, operating in the low-density region to approximate square-law operation. Basically, these meters were average-reading meters. Although they met the requirements for a two-tone rms response, their indications were much closer to average than rms on most other types of tests. This meter circuit, though not a true rms indicator, will indicate more closely an rms value than did the previous meter movements. Fig. 22-10A shows the characteristics and circuit of this meter, which combines an average-reading circuit with a peak-reading circuit, to create what has been termed a "quasi-rms indicating circuit." Since the rms value of most waveforms falls between the average peak indication and the actual peak value, adding a portion of the peak indication to the average value of a given waveform should result in an rms iudication. However, different waveforms require different amounts of peak indication to give rms values.

In the circuit shown, the ratio of R2 to R1 determines the amount of peak contribution, and one set of values will

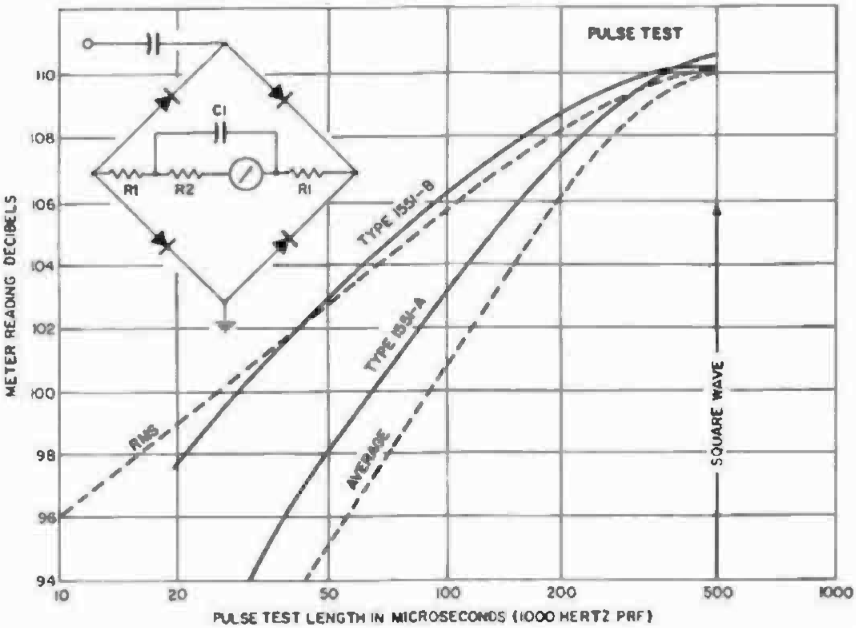


Fig. 22-10A. Quasi-rms meter circuit used by General Radio Co. in their sound level meters, and octave-band noise analyzers.

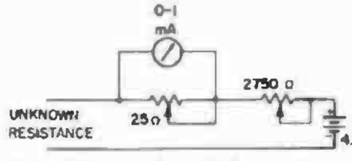
dB fluctuation With Phase Changes at 30% Third Harmonic	*For Two-Signal Addition*	*For Square Waves*	*For Noise*
0.45	+0.5	+0.1	−0.25
1.8	−1.0	+1.0	−1.0
1.7	−0.4	+0.6	−1.0

Fig. 22-10B. Difference in meter indication and rms decibels.

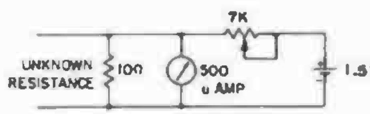
give an indication of many signals close to rms voltages. The characteristic for the quasi-rms meter circuit and an average-reading meter are given in Fig. 22-10B. The solid curves of Fig. 22-10A show the response of the quasi-rms meter and its predecessor to pulses of constant height but varying length. As will be observed, the quasi-rms meter indicates the rms value (upper dotted curve) within plus or minus 1 dB, until the pulse duration becomes as short as 1/25 that of a square wave.

22.11 Describe a volt-ohmmeter.—Volt-ohmmeters are service instruments designed for servicing of all types of electronic and electrical equipment. Basically, the circuity consists of a voltmeter with several ranges, generally from 1 volt to 1500 volts, for both ac and dc. Also included are resistance-measuring circuits. A basic circuit for measuring resistance above 10 ohms is given at (a) of Fig. 22-11A, and for resistance values below 10 ohms, the shunt circuit in (b) is used.

A vacuum-tube volt-ohmmeter manufactured by Allied Radio Corp. is shown in Fig. 22-11B. The principal advantage of the vacuum-tube volt-ohmmeter is its high input impedance,



(a) *Circuit for high resistance.*



(b) *Circuit for resistances below 10 ohms.*

Fig. 22-11A. Basic ohmmeter circuits.

as it absorbs only a minute amount of power from the circuit being tested. This feature permits it to be used for measuring vacuum-tube and transistor circuits generally without affecting their operation. Vacuum-tube volt-ohmmeters as a rule include all the scales of the nonelectronic volt-ohm-meter, and additional peak-to-peak ac voltage scales. Since the instrument illustrated employs a full-wave peak-to-peak rectifier for ac measurements, the meter actually indicates peak-to-peak values. In actual practice the rms scales are used exclusively, and the peak-to-peak scales only when a peak-to-peak measurement is necessary.

Dc voltages may be read full scale from 0.5 to 1500 volts with a full-scale accuracy of plus or minus 3 percent, and the ac rms voltages from 1.5 to 1500 volts, plus or minus 5-percent full-scale accuracy. Peak-to-peak voltages may be read from 4.2 to 4000 volts full scale. The frequency response is plus or minus 1 dB, 30 Hz to 3 MHz, and plus or minus 3 dB 3 Hz to 5 MHz. Decibels are read on a separate scale. By using a high-frequency probe, the range may be extended to 250 MHz. Ohmmeter scales in multiples of ten from 1 to 1,000,000 ohms provide a wide range of resistance measurement. The input impedance for all voltage

Fig. 22-11B. Allied Radio Corp. Model KG 620 vacuum-tube volt-ohmmeter.

ranges, both ac and dc, is 11 megohms. By the use of an external high-voltage probe, the voltage range may be extended to 25,000 volts. The meter movement is knife edge, with a fluorescent coating to improve viewing. An on-off position on the function switch provides a short circuit across the meter movement to prevent it from swinging and possible damage during transportation.

The schematic diagram for this instrument is given in Fig. 22-11C. The heart of the circuit is a vacuum-tube bridge circuit, composed of a 12AU7 dual triode and a 200-microampere meter. With the bridge circuit properly balanced, the voltages at the plates of V2A and V2B are equal. Since the meter is connected between the two plates, it will indicate zero. Variable resistor R20 connected between the plates of V2 is a front-panel balancing control, and is adjusted (after a short warm-up period) for a zero indication on the meter.

When a positive voltage is applied to the first control grid (pin 7) of V2A, the current through V2A increases, resulting in a lower voltage at the plate of V2A. Since the current of V2A flows through common cathode resistor R22, the voltage drop across R22 increases. Thus, the control grid of V2B is biased more negatively because of the increased voltage drop across the cathode resistor. This causes the current through V2B to decrease and the voltage at the plate (pin 1) to rise.

Current now flows through the meter, because of the difference in potential between the plates. The meter is calibrated to indicate the magnitude of the voltage applied to the control grid of V2A. Different voltage ranges are developed by switching precision resistors in series with the grid circuit of V2A. In practice when one is measuring ac voltage, the voltage is first rectified by V1, a 6AL5 dual diode functioning as a full-wave peak-to-peak rectifier. Instantaneous positive peaks of the ac voltage being measured will cause VIA to conduct, charging capacitor C2 to the positive value of the peak input voltage. As the ac signal voltage swings in a negative direction, VIA ceases to conduct and the charge is retained on capacitor C2. The negative peak of the input voltage is added in series with the charge on C2 and is

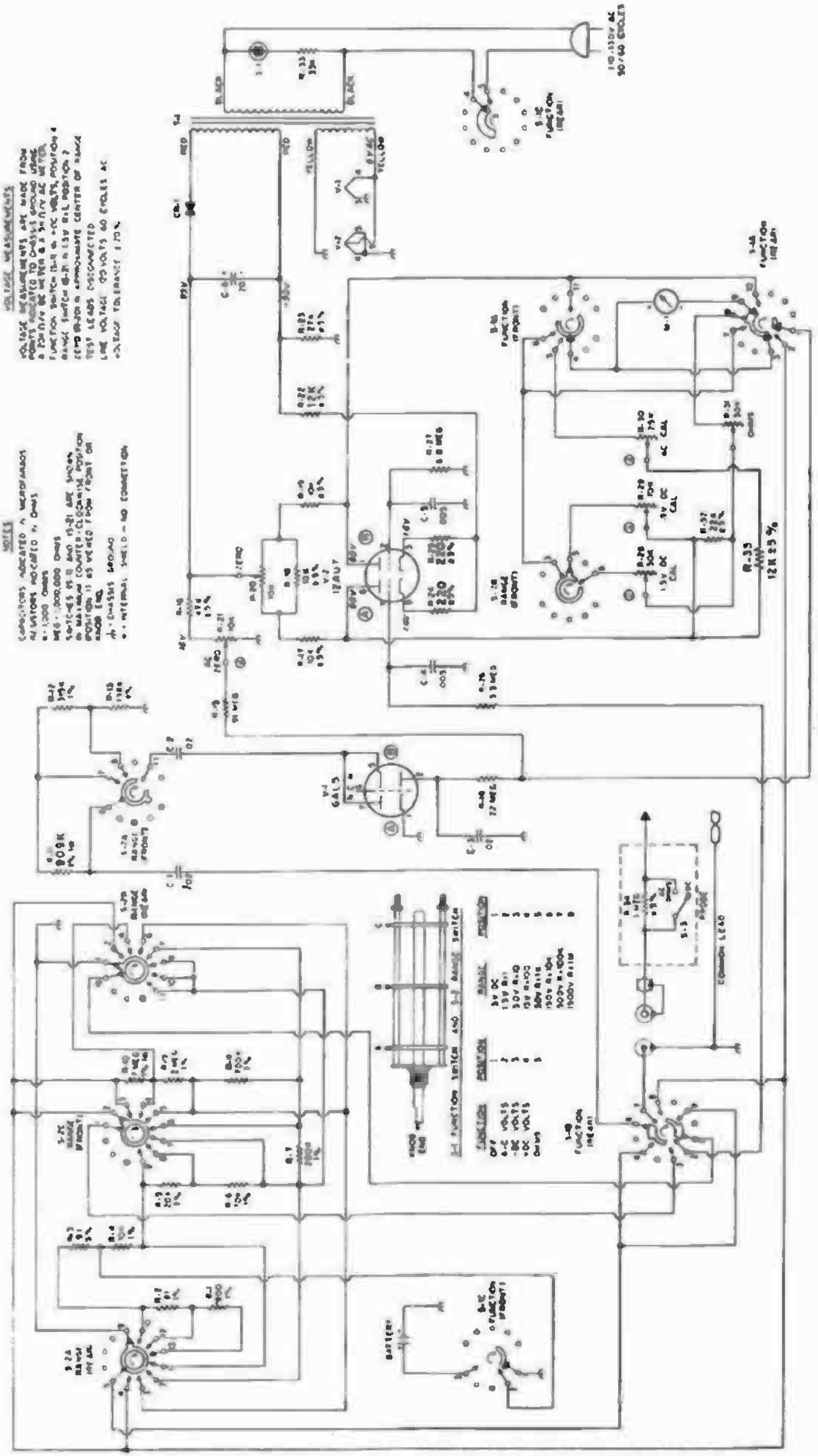


Fig. 22-11C. Schematic diagram for the Allied Radio Corp. KG 620 vacuum-tube volt-ohmmeter.

Fig. 22-11D. Transistor multipurpose meter, Model 427A manufactured by Hewlett-Packard.

applied to VIB. VIB now conducts and charges capacitor C3 to a value equal to the sum of the positive and negative peaks. The charge across C3 (or a portion) is applied to the bridge circuit and the meter calibrated in terms of peak-to-peak and rms voltage.

For resistance measurements, a 1.5-volt battery is connected through a series of precision resistors and the unknown resistor. These two resistors form a voltage-dividing network across the battery. The resulting voltage is applied to the bridge circuit and the meter, which is also calibrated to read ohms.

The function switch chooses the mode of operation. The range switch selects the required series resistance for the various ranges of measurement. An ac zero adjustment compensates for contact potential developed on the element of V1.

The probe circuit shown at lower center of the diagram is provided with a switch which connects a 1-megohm resistor in series with the input circuit in the dc voltage position to act as an isolation resistor for dc measurements. When ac voltage is measured, the resistor is switched out of the circuit.

Pictured in Fig. 22-11D is a transistor multipurpose meter, Model 427A, manufactured by Hewlett-Packard Co. The meter is capable of making dc measurements from 1 millivolt to 1000 volts, ac measurements from 0.3 millivolt to 300 volts at frequencies of 10 Hz to 1 MHz megacycle, and resistance measurements from 0.2 ohm to 500 megohms. A decibel scale is also provided, permitting measurements from minus 40 dBm to plus 50 dBm. The frequency response is shown in Fig. 22-11E.

Referring to the schematic diagram of Fig. 22-11F, at the lower left is a dc amplifier for regulating the battery or external source of operating voltage. It provides a plus 6.7 and minus 6.7 volts. Amplifier A1 is a high-impedance amplifier used to amplify dc and the resistance-measuring inputs, and also serves as a preamplifier for the ac signals. Metering amplifier A2 amplifies the ac signal from the preamplifier, converts it to a dc signal (proportional to the average ac), and feeds it to the meter. The meter displays the rms value of the ac signal.

In the dc mode of operation, the dc input voltage to be measured is applied to the dc range attenuator, where it is attenuated 10 dB for each step of the attenuator. The output of the attenuator goes to the function switch and then a dc filter network R6, R7, C2, and C3, rejecting any ac superimposed noise voltage that may be present on the dc input voltage under measurement. Superimposed peak ac voltages, 100

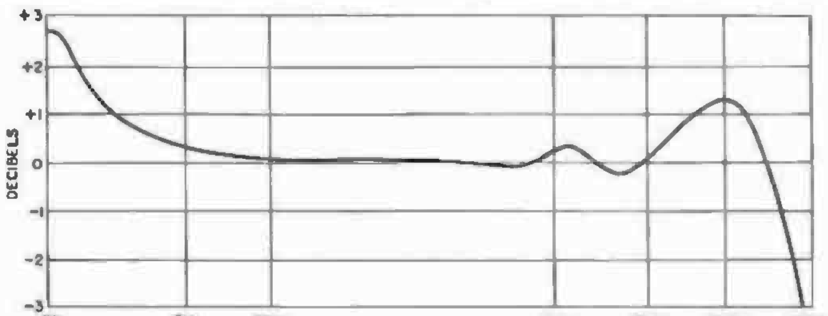


Fig. 22-11E. Frequency response of Hewlett-Packard Model 427A multipurpose meter of Fig. 22-11D.

times greater than full scale, effect the reading less than 1% for 60 Hz and above. The dc output of the filter network goes to amplifier A1 and then to the meter. Direct-current amplifier A2 matches the high impedance of the attenuator to the low impedance of the meter. The dc calibrate circuits are resistive, in series with the meter, and are used to calibrate the meter current on the lower ranges.

In the ohmmeter position with the input open, resistors R16 and R17 (amplifier A1) form a voltage divider. The voltage across R17 causes full-scale current to flow through the meter. Ohms calibrate circuit R37 adjusts the meter current for an indication of infinity. When the unknown resistor being measured is equal to R17, the total resistance from the ohms terminal to ground will be half of R17 resistance, the voltage across R17 and the unknown resistor will be halved, and the meter indication will be half scale. The circuit is designed so that the full range setting will be displayed at the center of the scale. For an example, 1 ohm on the times 10 range is a center-scale reading.

For ac measurements, the input signal is applied to the ac range attenuator. On the 1-volt range aud below, the signal is not affected by the attenuator. However, on all the higher ranges the signal is attenuated 50 dB. Variable capacitor C3 at the ac range attenuator is used to adjust the frequency response of the attenuator at 100,000 Hz on the 3-volt range. The signal from the ac attenuator goes through amplifier A1 and to the ac post attenuator, where it is attenuated 10 dB for each step of the range selector. The dc amplifier matches the low impedance of the post atteuuator to the high impedance of the range attenuator, acting as a preamplifier. The metering circuit contains both a feedback-stabilized ac amplifier and an averaging meter circuit. The meter circuit converts the ac signal to a dc voltage which is proportional to the average of the ac amplifier output. Resistor R17, between the meter and ground, adjusts the current through the meter to read in rms volts.

22.12 What is a hook- or clamp-on meter? — A service electrician's test meter that may be used to measure the current in a conductor without physically connecting the instrument to the circuit under measurement.

The meter consists of a sensitive movement and a single-turn loop that is clamped around the conductor in which the current is to be measured.

A Model AK-5 hook-on volt-ammeter, manufactured by General Electric, is shown in Fig. 22-12A. The range of this instrument is from 5 to 350 amperes and from 150 to 750 volts ac. This meter may be used to measure current surges of short duration, such as motor-starting currents, in either single- or multiple-phase circuits. The proper current range is selected by means of a small knob at the lower portion of the handle. Test leads are plugged into pin jacks near the knob for voltage measurements. The circuit of a typical clamp-on meter is shown in Fig. 22-21B.

22.13 If a meter is rated 2-percent, full-scale accuracy, what is its accuracy below full-scale?—Two-percent accuracy means that the guaranteed accuracy will only be obtained at full-scale deflection. The accuracy below full-scale deflection will vary, depending on the accuracy of the calibration and tracking of the meter movement. As a rule, in small meters the scale is printed and the accuracy depends on the accuracy of the meter movement or its percentage of deflection for a given voltage or current. If the scale is hand calibrated, full-scale accuracy may be maintained throughout the whole deflection length of the calibrations.

It is the practice of some instrument manufacturers to machine-calibrate each individual meter scale, to eliminate tracking error on mass-produced meter movements. The system used by Hewlett-Packard custom calibrates and photographically prints the meter scale to match exactly the linearity characteristics of each meter movement. The basic movement is the taut-band construction, which is thoroughly discussed in Question 22.7.

22.14 What is the purpose of a mirror behind the pointer of a meter?— To eliminate the error due to parallax when reading the meter scale. The most accurate reading is obtained when the pointer is lined up with its reflection in the mirror. This type of scale is quite common with precision meters such as shown in Fig. 22-14.

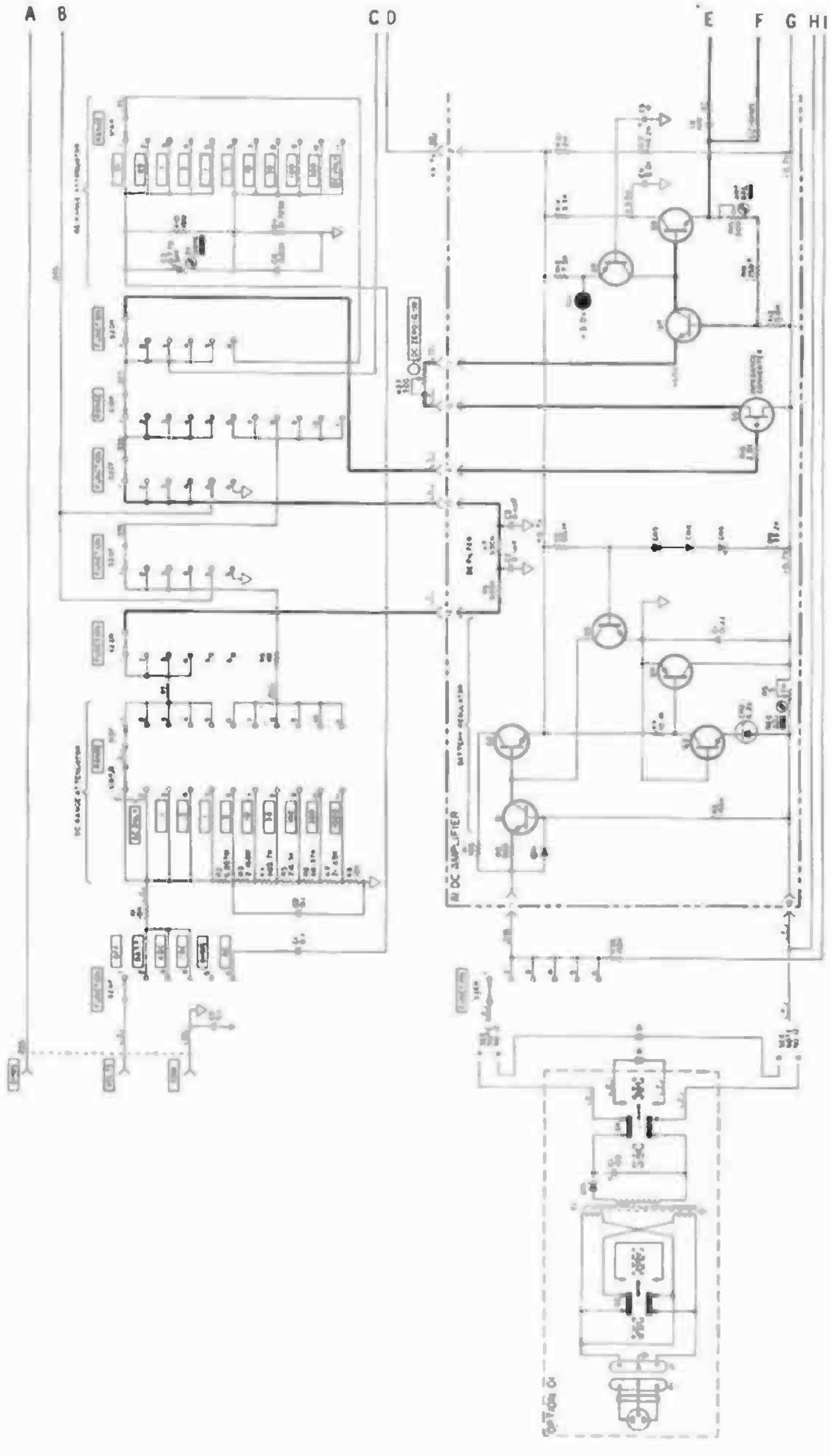
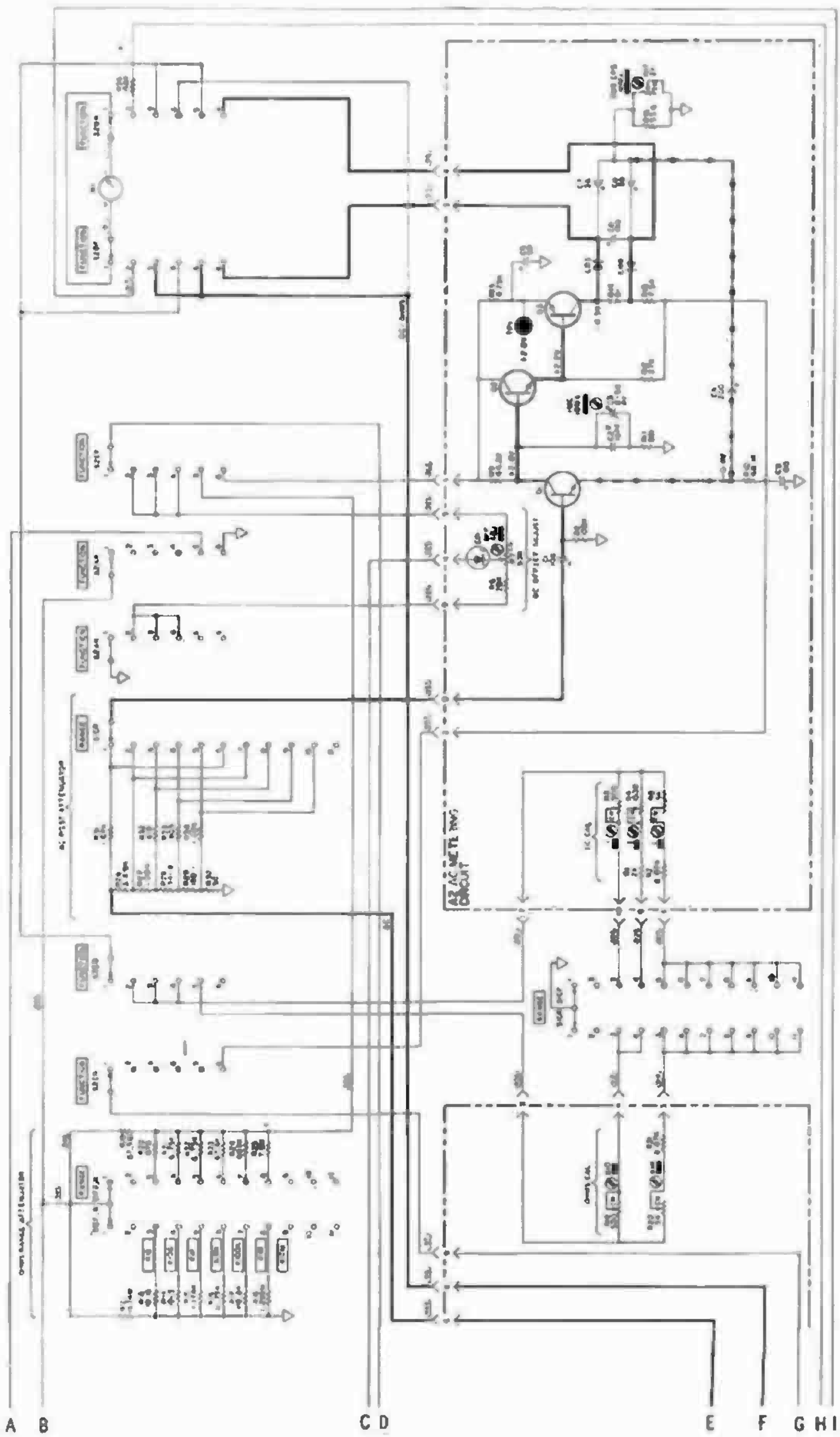


Fig. 22-11F. Schematic diagram for Hewlett-



Packard Model 427A multipurpose meter.

Fig. 22-12A. The General Electric Model AK-5 hook-on volt-ammeter being hooked on a cable.

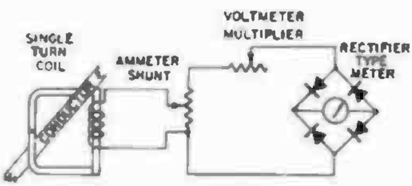


Fig. 22-12B. Circuit for a typical clamp-on volt-ammeter.

22.15 *What type of voltmeter is recommended for measuring transistor operating voltages?*—A meter with at least 10-megohms input resistance, for either dc or ac voltages. This high input resistance is necessary to prevent loading the circuits and upsetting the voltage divisions.

22.16 *How may the balance of a meter movement be checked?*—By noting its reading in both the vertical and horizontal positions. The reading should also be checked at a 45-degree angle.

22.17 *How are current shunts for an ammeter calibrated?*—They are calibrated in amperes and millivolts drop across the shunt. A typical example is the shunt in Fig. 22-17 designed to have a 50-millivolt drop across it when 10 amperes of current are flowing through it. The meter movement is, in reality, a sensitive voltmeter reading only the voltage drop across the shunt.

22.18 *Will a thermocouple meter read the same as a vacuum-tube voltmeter for a complex waveform?* — No. The thermocouple meter will read higher because it is not affected by waveforms. Therefore, the reading is

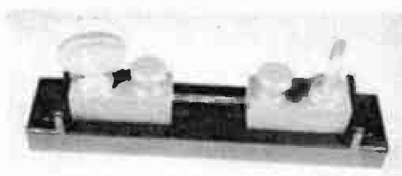


Fig. 22-17. A 10-ampere, 50-millivolt ammeter shunt.

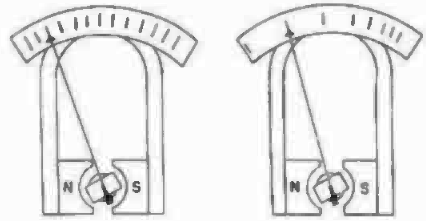
Fig. 22-14. The Sensitive Research Instrument Corporation multitester. It is designed for measuring both ac and dc voltage and current.

a true indication of the current and requires no correction for waveform.

22.19 ***How is a logarithmic characteristic obtained in a D'Arsonval meter movement?***—By shaping the magnetic pole pieces in the magnetic structure of the movement. This type meter movement is employed in the Ballantine vacuum-tube voltmeter described in Question 22.102. A linear movement is shown at (a) and a logarithmic one at (b) in Fig. 22-19.

22.20 ***What is an instrument transformer?***—A special-type transformer used with switchboard indicating instruments to isolate them from a high-voltage source. The secondary side of the transformer is provided with taps for different voltage or current ratios. When it is used with an ammeter, it is called a *current transformer*, and when used with a voltmeter, it is called a *potential transformer*. As a rule, the transformer is polarized and operates with a gronnded secondary to protect operating personnel.

22.21 ***What is a Wheatstone bridge?***—An instrument used for measuring resistance. Although the circuit used in a Wheatstone bridge was invented by Christie, Wheatstone received credit for its invention because of his adaptation of the circuit for the measurement of resistance.



(a) *Linear.* (b) *Logarithmic.*

Fig. 22-19. Linear and logarithmic meter movement obtained by shaping the pole pieces.

A commercial model of a Wheatstone bridge manufactured by the Leeds and Northrup Co. is shown in Fig. 22-21. The device pictured may be used for measuring resistance and locating and measuring the distance at which a ground or short circuit on a transmission line or cable has occurred. Either the Murray or Varley loop test may be used. Resistance valnes are measured by the aid of a four-dial decade box, a ratio dial, and a sensitive galvanometer.

22.22 ***What is a null indicator?***—The term null means a point of no voltage or current. Null indicators generally consist of headphones, an oscilloscope, or a vacuum-tube voltmeter or sensitive dc meter connected in a circuit to indicate the point of lowest

Fig. 22-21. The Leeds and Northrup Model 5430-AM Wheatstone bridge.

voltage or when a balance has been achieved by a zero indication. The instrument used for indicating a condition of balance in a bridge circuit is called a *null indicator*.

22.23 Show the various bridge circuits used for the measurement of resistance, capacitance, and inductance.—The most widely used bridge circuits for the measurement of resistance, capacitance, and inductance are shown in Figs. 22-23 through 22-31. The basic circuit of all bridges is the Wheatstone (Christie) bridge illustrated in Fig. 22-23. The circuit consists of four fixed or variable resistance arms, R_1, R_2, R_3, and R_4, a source of direct current, and a sensitive (zero center) galvanometer designed to swing in either direction.

The current from the battery enters the circuit at A. At this point the current divides into two parts, one part through R_2 and R_1, the other through R_4 and R_3. Both currents combine at point B and return to the battery. A dc galvanometer is connected from point C to point D to indicate when a balance or null has been reached.

The operation of the bridge circuit may be thus explained. Assume R_3 to be an unknown resistance. Resistors R_1, R_2, and R_4 are variable resistors and are adjusted with the battery connected until a null is indicated by the galvanometer. The circuit is now considered to be balanced. Points C and D are now at the same potential. If there were a potential difference between points C and D, a current would flow and would be indicated by a deflection of the galvanometer.

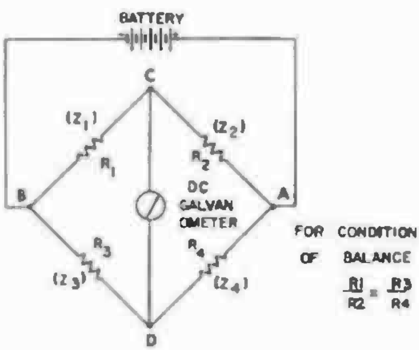


Fig. 22-23. Basic circuit for a Wheatstone resistance bridge. If reactive elements are substituted for resistances, the circuit becomes an impedance bridge.

Consider now the voltages across the individual resistors. It was stated that the current divided into two parts, with one part through ACB and the remainder through ADB. The current through ACB will be designated I_a and that through ADB as I_b. The voltage drop across points AC is $I_a \times R_2$, and across points AD it is $I_b \times R_4$. Points C and D are at the same potential, since the galvanometer indicates a null. Therefore the voltage drop from A to C is equal to the voltage drop across A to D. The voltage drop across C to B must be equal to the voltage drop across B to D. Under the foregoing conditions:

$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

The bridge may now be said to be in balance.

In the practical Wheatstone bridge, the arms R_2 and R_4 are ratio arms, adjustable in decades which are varied simultaneously. R_1 is a variable resistor group which may be varied from 1 ohm to several thousand. The unknown resistor is connected in place of R_3.

This bridge circuit may also be used for the measurement of impedance, inductance, and capacitance, as will be explained in the following questions.

22.24 What is a slide-wire bridge?—A simplified version of a Wheatstone resistance bridge. The basic slide-wire bridge circuit is shown in Fig. 22-24. It consists of two fixed resistors, R_1 and R_3. A slide-wire resistor in the form of a large rheostat provides the two lower arms, R_2 and R_4. The unknown resistor is substituted for R_1 or R_3. The slide-wire rheostat is adjusted for a null. The value of the unknown resistor may be calculated:

$$R_x = \frac{R_4}{(R_2 + R_4) - R_4} \times R_1$$

where,

R_1 is a standard resistor,
R_x is substituted for R_3.

22.25 What is a Maxwell bridge?—A form of Wheatstoue bridge used for the measurement of inductance and capacitance. The dc galvanometer is replaced by a pair of headphones or a vacuum-tube voltmeter. The battery is replaced by an alternating-current signal source.

The Maxwell bridge circuit is generally used for measuring inductance,

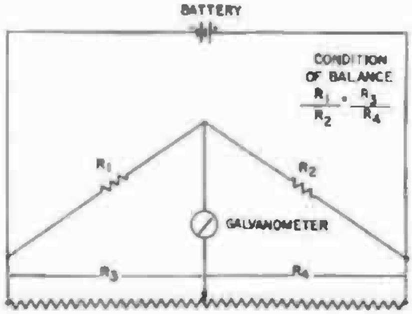


Fig. 22-24. Slide-wire type Wheatstone bridge circuit.

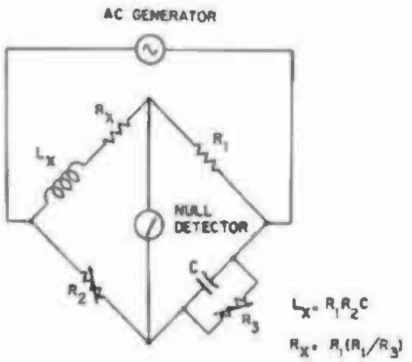


Fig. 22-25. A Maxwell bridge circuit for measuring inductance.

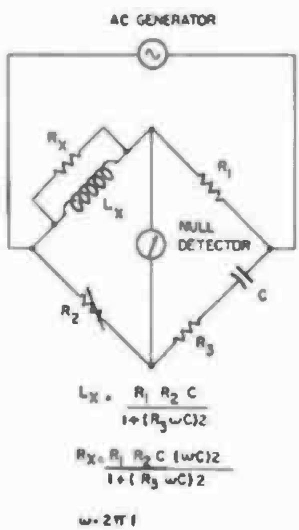


Fig. 22-26. The Hay bridge circuit used for measuring inductance.

as it will permit an unknown inductance to be measured, using a capacitor as a standard. This bridge may also be used to measure coils of high Q and coils having considerable series resistance (low Q).

The bridge is balanced in a manner similar to that used for balancing the dc Wheatstone bridge in Fig. 22-23, except the null point is indicated by a minimum signal in the headphones or minimum indication on a vtvm. The value of inductance may be calculated from the equation given in the diagram.

22.26 ***What is a Hay bridge circuit?***—A bridge circuit used for the measurement of inductance with a Q of over 10. The Hay bridge, Fig. 22-26, is often used for measuring incremental inductance or coils with a dc current through them.

22.27 ***What is a Schering bridge?***—A bridge used for the measurement of capacitance and the characteristics of electrical cables. (See Fig. 22-27.) An advantage of the Schering bridge is that it may be used for the measurement of electrolytic capacitors while a dc potential is applied to the capacitor, without the direct current circulating throughout the bridge.

22.28 ***What is an Owen bridge?***—The Owen bridge circuit, Fig. 22-28, is somewhat similar to that of the Maxwell bridge and is used principally for the measurement of inductance. Unlike the Maxwell bridge, the Owen bridge requires two capacitors, one in each of the upper two arms.

22.29 ***What is a Wien bridge?***—A bridge circuit which permits the measurement of capacitance in terms of frequency and resistance. It may also be used as a frequency-selective filter. The circuit is shown in Fig. 22-29.

The Wien bridge is the basic circuit used in resistance-tuned oscillators and

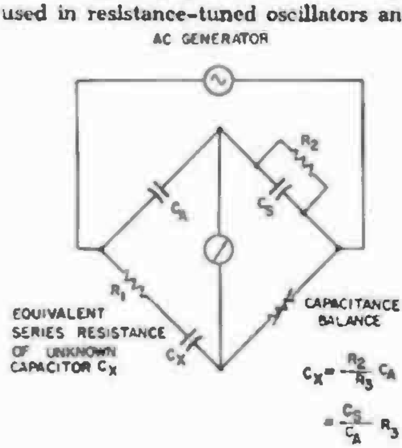


Fig. 22-27. The Schering bridge circuit used for measuring capacitance and cable characteristics.

other test equipment. The audio-frequency meter described in Question 22.36 is a Wien bridge calibrated to read directly in frequency.

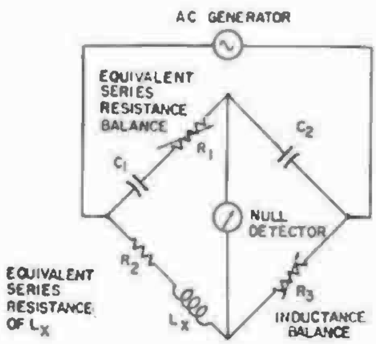
22.30 ***What is a resonant-frequency bridge?***—A bridge circuit as shown in Fig. 22-30. The unknown inductance, L_x, is connected in series with a capacitor C to form a series-resonant circuit at a frequency of the signal applied to the bridge. At the resonant frequency the inductive and capacitive reactances, being equal in magnitude and opposite in sign, cancel. This bridge may also be balanced if pure resistance is substituted for the reactive elements.

Because the capacitor must cover quite a large range, this type of bridge is seldom used for the measurement of inductance.

22.31 ***What is a Kelvin bridge?***—A bridge circuit devised by Lord Kelvin for the measurement of resistances below 0.10 ohm, such as might be encountered when measuring an ammeter shunt. Contact resistance between the connecting leads of the bridge and the unknown resistance can introduce errors up to 20 percent, using a conventional bridge circuit.

Measurements may be made of shunts having only a few thousandths of an ohm resistance with a high degree of accuracy using the Kelvin bridge circuit shown in Fig. 22-31.

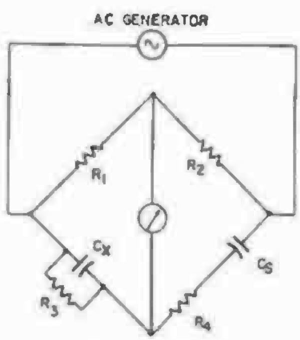
If ratio arms R_3 and R_4 equal R_1 and R_2, the contact resistance between unknown resistance R_x and R_5, the standard, is eliminated. The contact resistances at points A and B have no effect, because the current through the galvanometer is zero when the bridge is balanced.



$$L_X = R_1 R_3 C_2$$

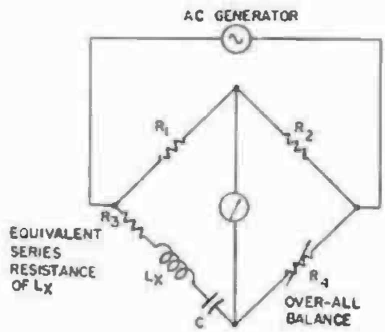
$$R_2 = \frac{C_2}{C_1} R_3$$

Fig. 22-28. The Owen bridge circuit used for measuring inductance.



$$C_X = \sqrt{\frac{R_2 R_3 - R_1 C_S}{R_1 R_3^2 R_4 (2\pi f)^2}}$$

Fig. 22-29. The Wien bridge circuit which permits measuring capacitance in terms of frequency and resistance. It may also be used for frequency rejection and measurement.



$$L = \frac{1}{(2\pi f)^2 C}$$

$$R_3 = \frac{R_1}{R_2} R_4$$

Fig. 22-30. Resonant-frequency bridge.

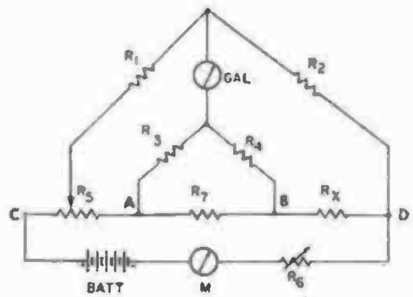


Fig. 22-31. The Kelvin bridge circuit used for measuring resistances below 0.10 ohm.

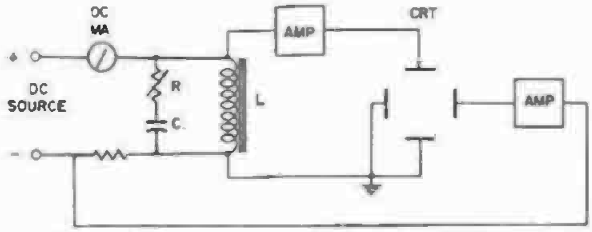


Fig. 22-32. Basic circuit for an incremental inductance bridge.

Contact resistance at points C and D is negligible compared with ratio arms R_1 and R_2, both of which are reasonably high. Resistor R_4 is adjusted for a current indicated by ammeter M for the required sensitivity. Resistor R_3 is small in value, about 1 to 10 ohms.

For a condition of balance:

$$\frac{R_2}{R_4} = \frac{R_1}{R_3}$$

22.32 ***What is an incremental inductance bridge?*** — An instrument designed specifically for the accurate measurement of the inductance of iron-core inductors, filter chokes, transformers, plate reactors, and similar devices. The incremental inductance bridge permits measurements to be made with direct current through the inductance under test. It is possible, therefore, to measure the inductance at any direct current or core saturation condition.

The inductance under test is connected so as to form a parallel resonant circuit consisting of L, C, and R, as shown in Fig. 22-32. The bridge circuit is supplied with an ac voltage at a frequency of 120 Hz superimposed on a dc voltage. The magnitude of the direct-current component is set to a desired value by adjusting the output of the dc power supply. The value of the inductance is obtained by the adjustment of resistor R, until resonance of the parallel circuit is indicated by a null on the cathode-ray oscilloscope tube. Resonance of the parallel circuit is obtained when:

$$L = CR^2$$

The value of the inductance is read from the resistor R which is calibrated to read directly in henries.

22.33 ***Describe a bridge circuit suitable for the measurement of resistance, capacitance, inductance, and impedance.*** —Such a circuit, developed by Heathkit, is given in Fig. 22-33A. This circuit is designed for the measurement of resistance, capacitance, inductance, and impedance. By the use of a single switch, four separate bridge circuits may be selected, including the Maxwell and Hay inductance bridges, Wheatstone resistance bridge, and a capacitance comparison bridge. An internal 1000-Hz oscillator supplies a signal for ac measurements. The dc for resistance measurements is taken from an internal power supply, which also supplies a source of voltage for the detector (bridge balance) amplifier. External binding posts are provided for the sample under test and for connection of an external oscillator and detector when required.

When the bridge is set for a given mode of measurement, the proper bridge circuits are automatically selected. One main dial, called the CRL dial, is employed to balance the bridge for all types of measurements, except for ac measurements. For these measurements, two additional dials are used in the balanacing procedure, and are termed the D and Q dials. Dial D represents the dissipation factor and Q its reciprocal storage factor. The equations for these two factors are:

$$D = \frac{1}{Q} = \frac{R}{X_L} \qquad Q = \frac{1}{D} = \frac{X_L}{R}$$

where,

R and X_L are the series resistance and reactance of an inductance or capacitor under measurement.

The dissipation factor is directly proportional to the energy dissipated per cycle; the storage factor is directly proportional to the energy stored per cycle. The term "dissipation factor" is commonly used with capacitors as it varies with loss, while "storage factor" is used for inductors, since it indicates the merit (*Q*) of a coil.

Resistance measurements are made with the Wheatstone bridge. As this is a

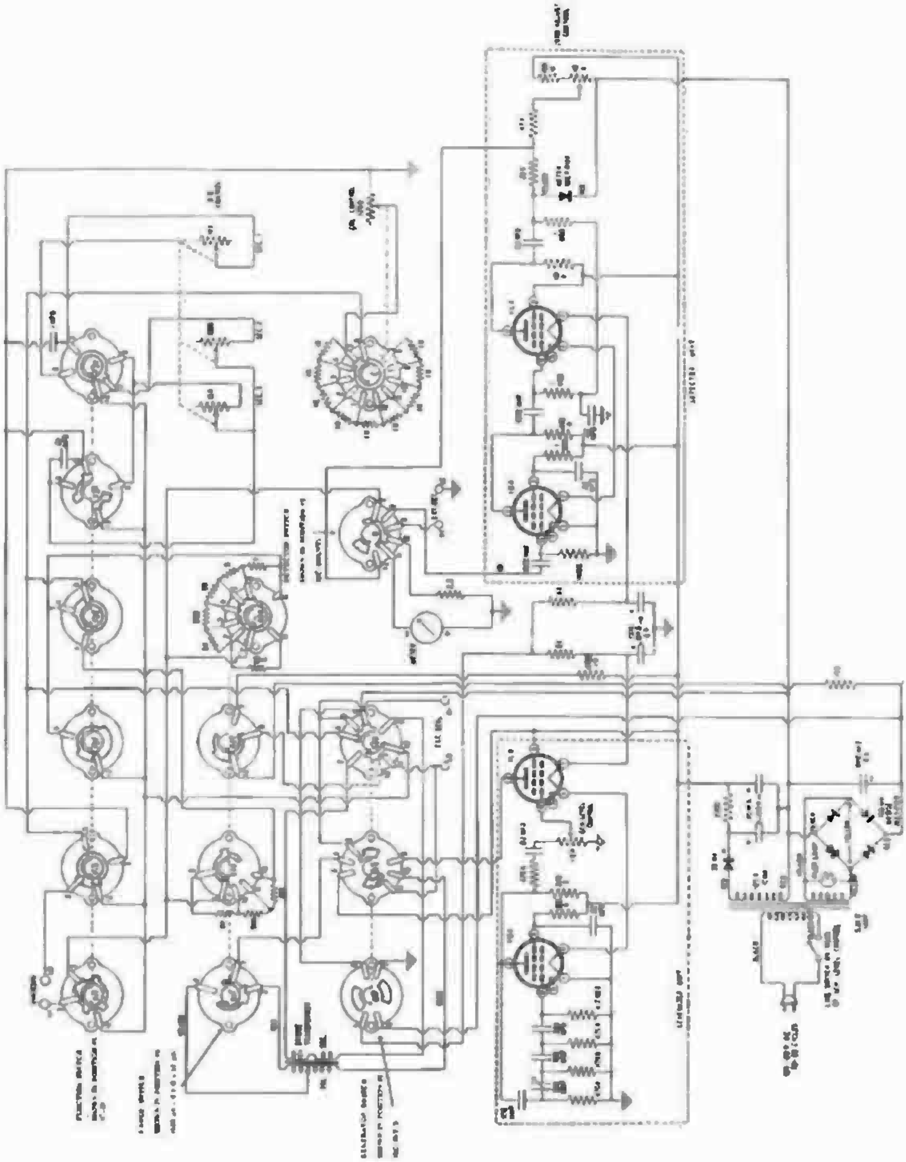


Fig. 22-33A. Schematic diagram for Heathkit Model IB-2A impedance bridge.

dc measurement, the bridge is balanced for zero indication on the zero-center meter. When the meter indicates zero, the bridge is in balance and the value of the sample in ohms is that indicated by the CRL dial multiplied by the range dial. The complete instrument is shown in Fig. 22-33B.

A laboratory-type impedance bridge, Model 1650A, manufactured by the General Radio Co., is shown in Fig. 22-33C. It is designed for measurement of resistance, capacitance, inductance, and impedance. Five bridge circuits are used. The Hays and Maxwell inductance bridges and series and parallel capacitance-comparison bridges provide a wide coverage of the *D* and *Q* ranges. Full use of these ranges at low-*Q* and high-*D* values is achieved by means of an "Orthonull" balancing mechanism. Both dc and ac measurements may be made with the bridge, as it has no internal phase balance.

The Orthonull balancing mechanism is a mechanical device that improves the bridge balance convergence when low-*Q* inductors or high-*D* capacitors are measured. Those who have tried to balance low-*Q* components on the con-

Fig. 22-33B. Heathkit Model IB-2A impedance bridge.

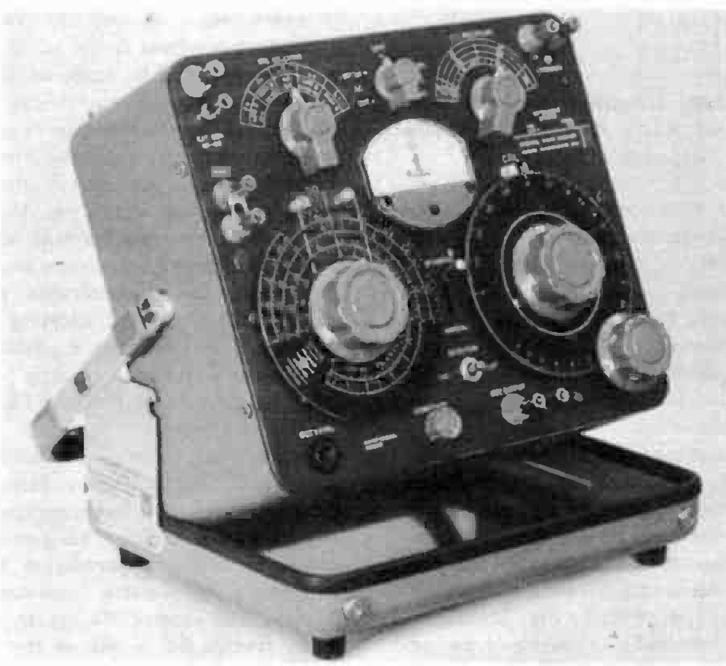


Fig. 22-33C. General Radio Co., impedance bridge Model 1650A.

ventional bridge have experienced a sliding balance, which is a slow balance convergence, resulting from the interdependence between the two balance adjustments. Thus, when a Q of less than 2 is encountered, the balance becomes tedious, and it is extremely difficult for values below 0.5. The Orthonull ties together mechanically both the CRL dial and the *D* dial, which are one and the same for this bridge. The Orthonull gangs the two adjustments nonreciprocally in such a manner as to cancel their interdependence, leaving the two adjustments independent of one another. As a result, advantage is taken of the ful *Q* range and the balance is simplified.

Contained internally within the case is a 1000-Hz transistor oscillator and a three-stage amplifier, which can be made either selective, with over 20-dB

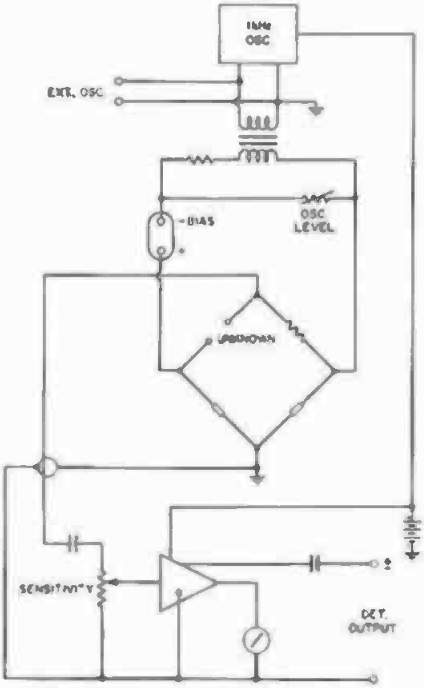


Fig. 22-33D. Simplified block diagram for General Radio Co., Model 1650A impedance bridge.

reduction of second harmonics or flat for measurements at other frequencies. The amplifier drives the panel meter to give a visual indication, thus eliminating the need for headphones. Terminals are provided for an external oscillator, detector, dc bias for polarizing the element under measurement, and headphones, if desired.

The resistance ranges are from 1 milliohm to 10-megohms; capacitance ranges from 1 pf to 1000 μF; inductance ranges from 1 μH to 1000 H, with accuracy of plus or minus 1 percent. An elementary block diagram of its internal connections is given in Fig. 22-33D.

22.34 ***Describe a service-type capacitor tester.***—Such a circuit is given in Fig. 22-34A. For measurements of capacitance the bridge circuit consists of four arms, potentiometer R1, standard capacitors C1, C2, or C3, and the unknown capacitor under measurement. For resistance measurements the four arms consist of potentiometer R1, standard resistors R2 or R3, and the unknown resistance. Potentiometer R1 is considered to be two arms of a bridge circuit, the slider being the center arm, which is varied to balance the bridge. The balance or null point is indicated on a 6E5 electron-ray tube, sometimes referred to as a magic-eye tube. The circuit shown permits the measurement of electrolytic capacitors with a polarizing voltage applied. For paper, oil, mica, or other types of dielectrics, a polarizing voltage is not required. The dial for potentiometer R1 is calibrated to read directly in terms of capacitance or resistance.

Since electrolytic capacitors have an inherent amount of internal resistance which increases with use and age, their power factor increases and reduces their efficiency. To balance out this resistance during measurement, potentiometer R1 and series resistance R4 are used to balance the bridge.

Control R4 is calibrated to indicate from 0 to 50 percent, in terms of power factor. Thus, the efficiency of the capacitor is readily determined during measurement. Electrolytic capacitors indicating a power factor of 15 percent or greater should be replaced; however, for certain types of electrolytic capacitors the power-factor may read higher and the capacitor still be used. The manufactnrer's data sheet should be consulted before replacing. Capacitors other than electrolytics will not indicate power factor; therefore resistor R4 is turned to its minimum position, which trips switch S2, shorting out resistors R5 and R6. For a 50-Hz power line the power-factor reading is multiplied by 0.84; for 40 Hz, by 0.72; and for 25 Hz, by 0.46.

When electrolytic capacitors are measured, the unknown capacitor is connected between the capacitance terminals of the bridge. The positive terminal of the capacitor must be connected to the positive terminal of the bridge, and control R4 set to electrolytic. Switch S1 is set to the desired measurement range, and potentiometer R1 and power-factor resistor R4 adjusted for the widest possible opening of the 6E5 tube. When the bridge is balanced, the value of capacitance is read from the dial and multiplied by the range position of S1. Power factor is directly read from the setting of R4.

Capacitor leakage is measured by turning S1 to a voltage near the rated dc working voltage of the capacitor and throwing S3 (spring loaded to automatically return to normal position) to leakage, and observing the action of the

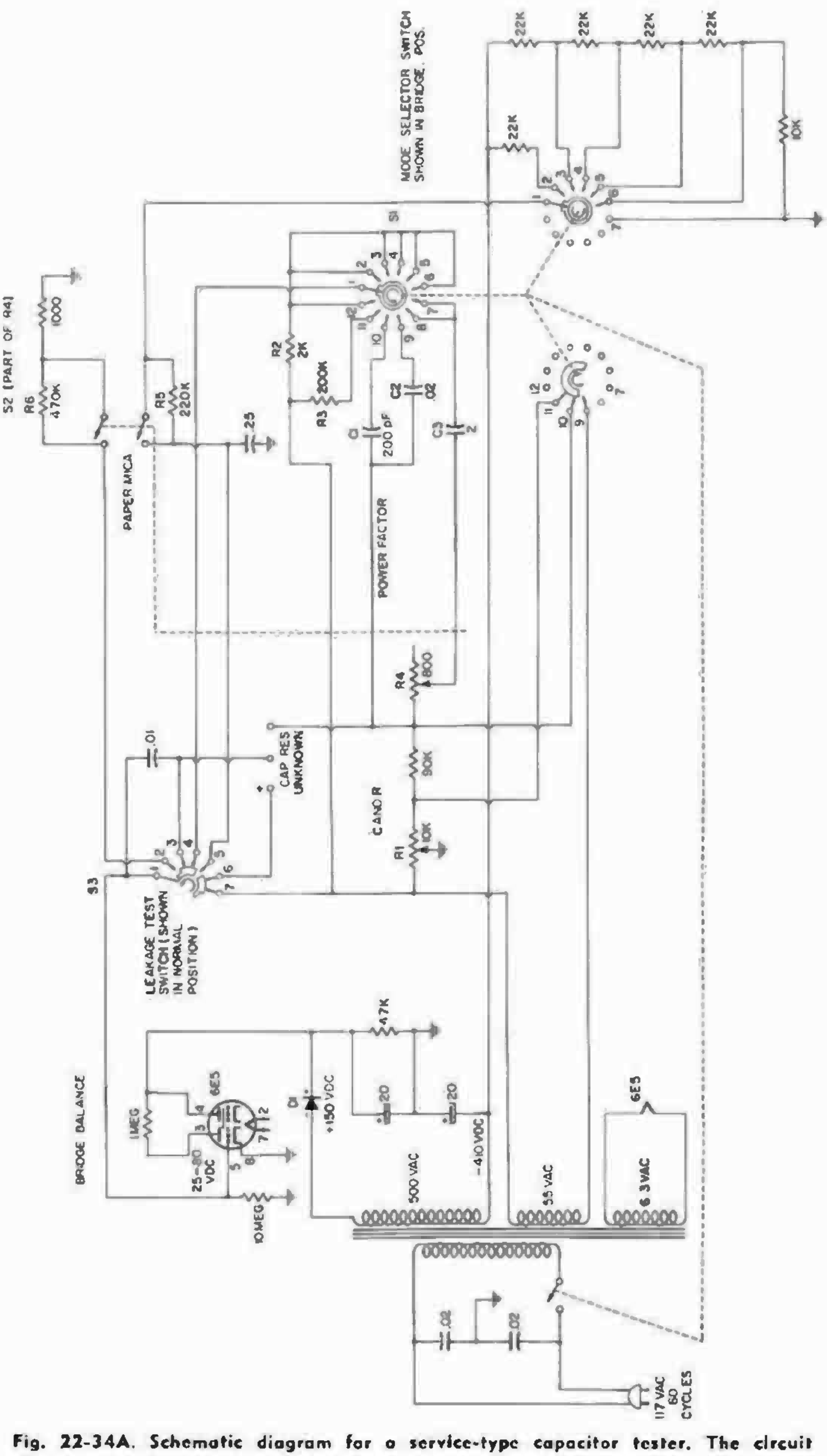


Fig. 22-34A. Schematic diagram for a service-type capacitor tester. The circuit shown will measure capacitance, power factor, and leakage. Resistance may also be measured.

Fig. 22-34B. EICO Electronic Instrument Co., Inc. Model 965 FaradOhm bridge analyzer.

6E5 tube. The amount of leakage will be indicated by the degree of the closing of the eye. If the eye flutters or is intermittent, the capacitor is leaky and arcing over internally. If it is internally shorted, the eye will close completely. Open-circuit capacitors will cause the 6E5 to open at the extreme ranges of capacitance. A normal capacitor will cause the eye to suddenly open and then return to the normal shadow angle.

Resistance is measured by connecting the unknown resistor to the resistance terminals (no polarity need be observed), S2 to the paper or mica position, and S1 to the desired range. The bridge is then balanced, using only potentiometer R1 for a maximum opening of the 6E5 tube.

The circuit shown will measure capacitance from 0.00001 to 1000 μF in four ranges; power factor 0 to 50 percent; leakage 25 to 450 Vdc; resistance 100 to 50,000 ohms. It should be pointed out that this tester circuit must *not* be used for the measurement of low-voltage capacitors used in transistor equipment.

Shown in Fig. 22-34B is a Model 965 FaradOhm bridge analyzer, manufactured by EICO Electronic Instrument Co. This instrument is designed for measuring low-voltage transistor-type capacitors, resistance, and inductance. In addition it will measure the leakage in capacitors, diode reverse current, transistor quiescent current, and insulation leakage. The instrument incorporates a six-range dc vacuum-tube voltmeter (vtvm) and an eleven-range dc vacuum-tube ammeter (vtam). The ammeter has eleven full-scale ranges, with a low scale of 0 to 15 μA, to a high scale of 0 to 15 mA, with an internal voltage drop of 75 mV.

The bridge circuit normally uses only 0.45 Vac at the line frequency. For testing at higher voltages or frequencies, an external source is used in accordance with existing EIA Standards. The bridge detector-amplifier incorporates an automatic gain control (agc) to hold the bridge detector meter on scale during the balancing of the bridge circuit, yet permits a finely defined null at the balance point.

To balance out the equivalent resistance when capacitors are measured, a control is provided, calibrated power-factor zero to 80 percent. A plug-in shield is provided for eliminating stray

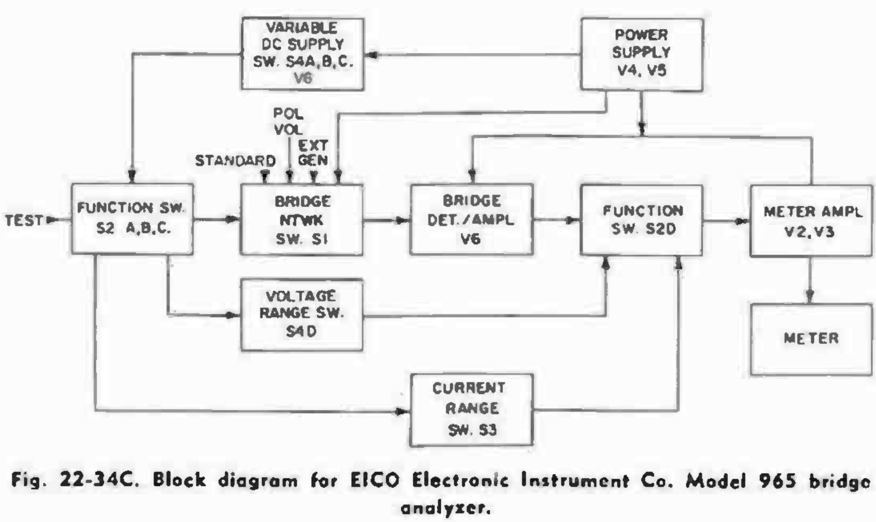


Fig. 22-34C. Block diagram for EICO Electronic Instrument Co. Model 965 bridge analyzer.

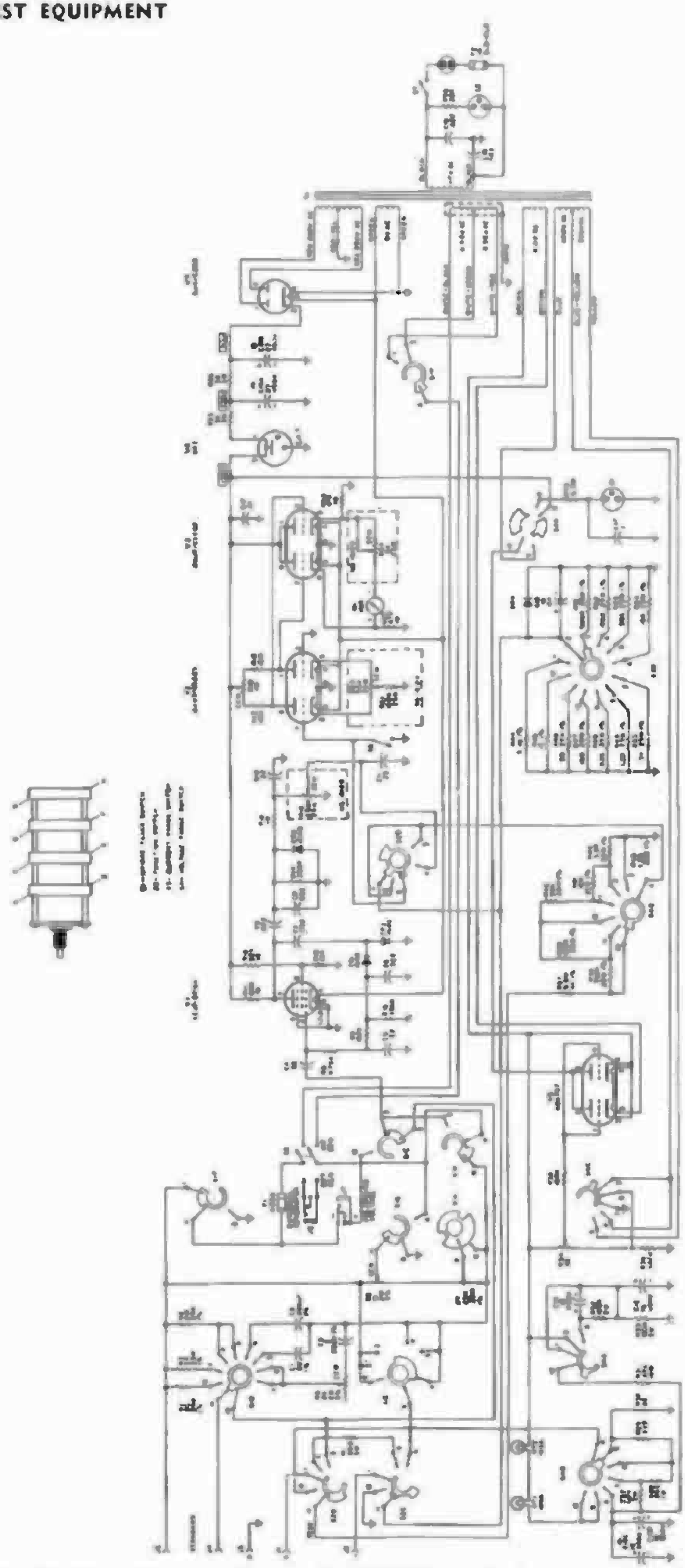


Fig. 22-34D. Schematic diagram for EICO Electronic Instrument Co., Inc. Model 965 bridge analyzer.

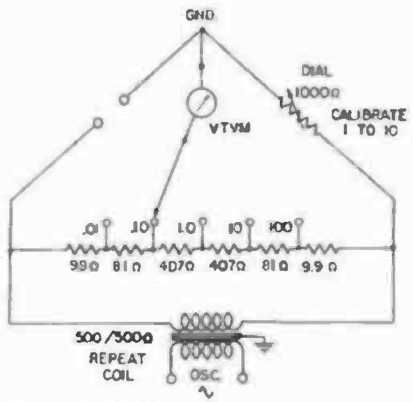


Fig. 22-35. A Z-set or ac ohmmeter used for measuring coil and transformer impedance.

fields when small capacitors or very large resistors are measured.

A six-range 0- to 500-Vdc power supply operates in conjunction with the vtvm and vtam for capacitor-leakage analysis, diode reverse or transistor quiescent current measurements. A warning light indicates when there is a voltage at the test terminals. Capacitance measurements are possible over a range of 5 pF to 5000 μF. Resistances can be measured from 0.5 ohms to 100,000 ohms, and variable dc voltage measurements from 0.4 to 500 volts are possible. A block diagram and schematic diagram are given for this instrument in Fig. 22-34C and Fig. 22-34D.

22.35 ***What is a "Z" set?***—A simplified bridge circuit used for the measurement of impedance. This circuit is sometimes referred to as an ac ohmmeter. (See Fig. 22-35.)

At the bottom of the diagram is a 1:1 coil to isolate the bridge from the signal source, if grounded or unbalanced. Six precision resistors (1 percent) provide the multiplier arms, while the 1000-ohm pot and the unknown impedance complete the bridge circuit. Headphones or a vtvm may be used as a null indicator. The bridge is balanced when a null is indicated on the meter. The variable 1000-ohm resistance is calibrated to read directly in ohms impedance.

22.36 ***Describe a Wien bridge audio-frequency meter.***— Audio-frequency bridges using Wien-bridge circuitry consist of only resistance and capacitance and are used for the precise measurement of audio frequencies between 20 and 20,000 Hz (Fig. 22-36). They may be used to identify frequencies of an oscillator, beat frequency, or other sources in the audio band.

Such instruments are connected at the output of the signal source, with a null detector connected at the bridge output terminals. When the bridge has been balanced for the lowest indication of the null detector, the frequency is read directly from a calibrated dial. A second balancing dial is provided to sharpen the balance and increase the accuracy of the readings. A two-gang potentiometer assembly employing log-taper windings is used to spread the calibration on the dial over a 270-degree range. A multiplier switch pro-

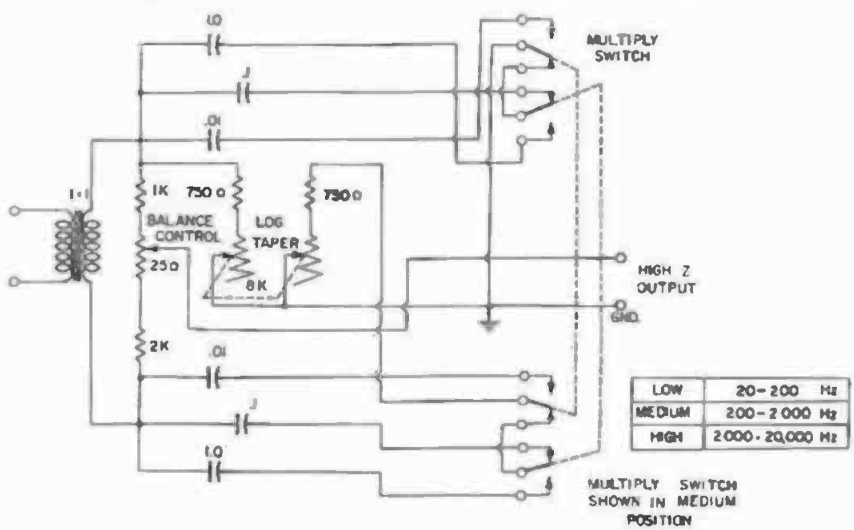


Fig. 22-36. Schematic diagram for Wien bridge audio-frequency meter.

vides three ranges of measurement—20 to 200 Hz, 200 to 2000 Hz, and 2000 to 20,000 Hz.

22.37 ***What is a resonant or tuned-reed frequency meter?*** — A frequency meter employing a group of tuned reeds for measuring the frequency of commercial power sources. A typical resonant or tuned-reed frequency meter, manufactured by the James G. Biddle Co., is shown in Fig. 22-37A.

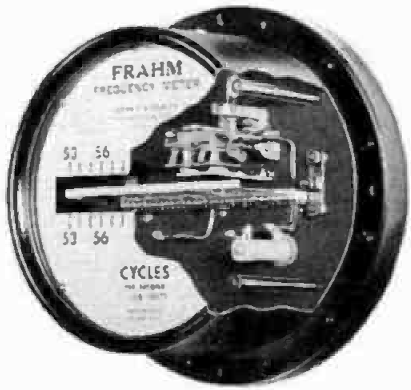


Fig. 22-37A. The James G. Biddle Co. Model 5110 Fram switchboard-type resonant-reed frequency meter. The frequency range is 52.5 to 67.5 Hz in steps of 0.5 Hz.

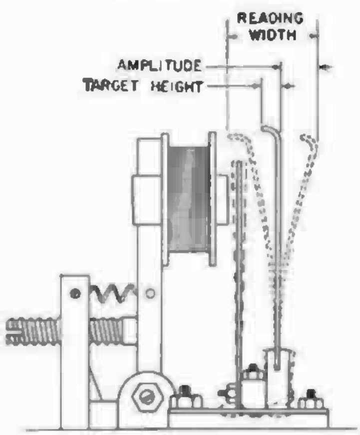


Fig. 22-37B. Cross-sectional view of electromagnet, armature, and tuned-reed assembly of the meter pictured in Fig. 22-37A.

Basically the instrument consists of a group of 31 or more steel reeds with one end bent at a right angle to its length. The bent end is painted white and the other end is clamped in a brass shoe. Each individual reed is tuned to a given frequency within a band of selected frequencies. The complete assembly of reeds is called a *reed comb* and is supported by two flat springs which act as armatures and are vibrated by a magnetic field from an electromagnet with a permanent-magnet core. The windings of the electromagnets are connected in parallel with the line whose frequency is to be measured.

A cross-sectional view of the tuned-reed comb is shown in Fig. 22-37B. This assembly may also be seen in the cutaway illustration of Fig. 22-37A. The amplitude swing of the reeds may be adjusted by decreasing or increasing the distance of the electromagnets to the armatures.

When the instrument is in operation the electromagnets are energized by the power-line voltage. This causes the spring armatures to vibrate at the

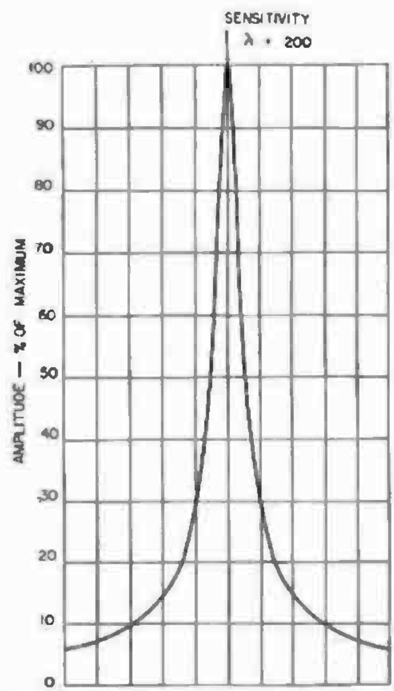


Fig. 22-37C. Resonance curve of a tuned reed used in a Fram frequency meter.

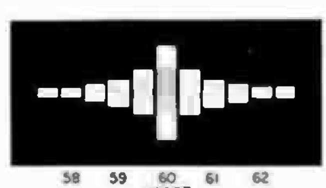


Fig. 22-37D. Appearance of tuned reeds in a Fram frequency meter when actuated by a frequency of 60 Hz.

power-line frequency, actuating the tuned reeds, which are tuned to frequencies falling above and below the normal power-line frequency in steps of one-half cycle each. The motion of the armature and the tuned reeds is shown by the dotted lines. The dotted lines at the base of the reed show how the vibration is transmitted through the shoe to the tuned reed.

The reeds behave at resonance somewhat like a whip. The tip swings through a wide arc while the lower end is anchored by the brass shoe. A typical response curve for a reed resonated at 60 Hz is shown in Fig. 22-37C. Although all the reeds receive the vibrations of the armature, only a small vibration is visible on those not in tune with the actuating frequency. This is illustrated in Fig. 22-37D. As a rule, the frequency range covered by the group of reeds is limited to about 2 percent above and below the normal line-voltage frequency.

A second reed mechanism called the *direct-drive* type is shown in Fig. 22-37E. Here, the reeds are vibrated by the direct action of the electromagnets.

Meters as described in the foregoing are used extensively for adjusting the frequency of the three-phase generating equipment run from batteries in sound trucks employed for recording motion pictures, where frequency is an important consideration. These meters are also used with small battery-driven generators, snch as the one shown in Fig. 3-22.

Frequency meters are connected in parallel with the line, similar to a voltmeter, and may be obtained to indicate frequencies from 30 to several hundred hertz. Each reed is tuned to within 0.30 percent of its designated frequency, which is an error at 60 hertz of less than 0.2 Hz.

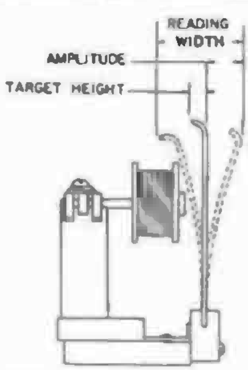


Fig. 22-37E. A direct-drive, tuned-reed mechanism for a From frequency meter.

Frequency meters of the tuned-reed design may also be obtained to cover twice the lowest frequency range in one instrument. A typical scale is 20 to 40 Hz and 40 to 80 Hz. The double range is obtained by using a soft-iron armature in the electromagnet rather than the permanent-magnet core as previously mentioned. With the permanent-magnet core the applied alternating current through the coil increases and decreases the strength of the permanent field once during each electrical cycle. The armature follows the variation in the field strength and vibrates the reed-comb assembly once during each cycle.

If a soft-iron core is used, however, the magnetic field alternates in direction as well as in strength and the armature vibrates twice for each current cycle. As a rule, a soft-iron core is used for frequencies below 30 Hz. Instruments equipped for double-range operation employ two electromagnets, one having a permanent-magnet core for the higher range and the other a soft-iron core for the lower frequency range. The same set of reeds is used for both ranges. The proper magnet is selected by means of an external switch.

22.38 ***What is an impedance bridge?*** —An instrument used for the precise measurement of the dc and reactive components of electrical devices and circuits. Such instruments generally measure capacitance, inductance, and resistance. The circuit diagram of an impedance bridge is shown and explained in Question 22.33.

22.39 ***What is a primary frequency standard?*** — A standard frequency to which other devices may be compared. A primary frequency standard is maintained by the National Bureau of Standards at Washington, D. C. By means of radio station WWV, the Bureau transmits standard frequencies for use by manufacturers and other interested persons on a 24-hour basis. A schedule of transmissions and frequencies may be obtained by writing to the Bureau.

22.40 ***What is a secondary standard?***—It is a highly stable device used as a standard and previously calibrated in terms of a primary standard. It may consist of an inductance, resistor ca-

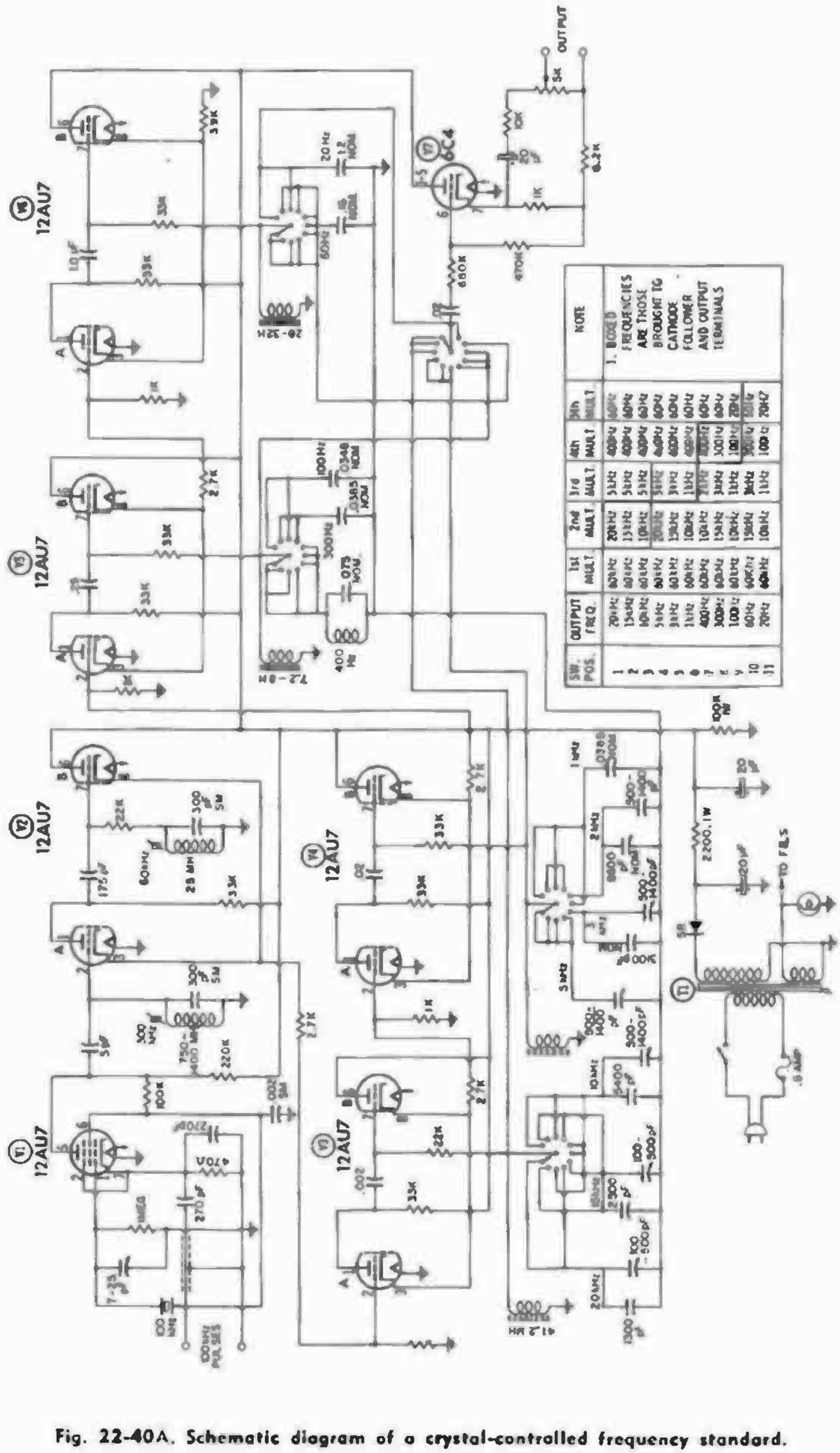


Fig. 22-40A. Schematic diagram of a crystal-controlled frequency standard.

pacitor, frequency generator, or any other electrical or mechanical device used for the calibration of other instruments.

Fig. 22-40A shows a schematic diagram for a secondary frequency standard, designed to generate sine-wave frequencies ranging from 20 to 20,000 Hz. The frequency is controlled by a master crystal-controlled oscillator, followed by five frequency dividers in various combinations. Possible frequencies are given in the table on the drawing.

Since the frequency dividers are locked to the excitation frequency applied to them, all output frequencies are held to the same degree of stability as the 100-kHz crystal oscillator. Each of the output frequencies may be used as a standard frequency signal with a short-term accuracy of one part per million.

The crystal used is of the DT cut, at an adjusted angle, such that the temperature drift characteristics are held to within 20 parts per million over a temperature range from 25°C to 65°C. The crystal is operated in a grid-screen crystal oscillator circuit using a type 6AU6 tube. No tuned circuits are required in the oscillator with the circuit employed. A 100-kHz signal, with high harmonic content, is brought out to the front panel by capacitive coupling to the cathode element of the oscillator tube. This circuit may be used for coupling to a radio receiver for beating of harmonics against a received signal from WWV, for checking calibrations.

A small variable capacitor is provided for setting the crystal oscillator to frequency, if required.

Since the lowest common multiple of the first switched frequency divider (20 kHz, 15 kHz, 10 kHz) is 60 kHz and not 100 kHz, it is necessary that the output frequency of the crystal be converted to 60 kHz before it is introduced into the divider chain. This function is accomplished in two stages. A tuned circuit is introduced into the grid circuit of V2A, the 12AU7 locked multivibrator tube, to pick off the harmonic at 300 kHz. This 300-kHz signal is divided by a factor of five to produce a 60-kHz signal at the output of the multivibrator.

The signal from the first frequency divider is coupled by a low-impedance voltage divider into the control grid of the first switched divider, V3. This divider has one inductor, five frequency-determining circuits, and three values of capacitance which are successively placed across the single-value inductance. With the lowest values of capacitance, the divider is locked at 20 kHz. With the intermediate value, the divider is locked at 15 kHz, and with the highest value, the multivibrator is locked at 10 kHz. The schematic diagram includes a box chart which illustrates the operating frequencies of the successive frequency dividers in terms of the signal output frequency of the complete unit.

The third frequency divider V4 is substantially identical to the second divider, except that there are four values of capacitance which are successively switched across the single inductor. The 5-kHz, 3-kHz, and 1-kHz signals are brought to the front panel, but the 2-kHz signal is used only to provide an integral division factor for the succeeding stage when the succeeding divider is operating at 400 Hz.

The fourth frequency divider V5 delivers an output at 400, 300, and 100 Hz. Due to the relatively low Q obtainable in small tuned circuits at these frequencies, it was necessary to utilize a more favorable L/C ratio at the two ends of the operating range. Hence, for an output of 400 Hz, an additional 2.5-

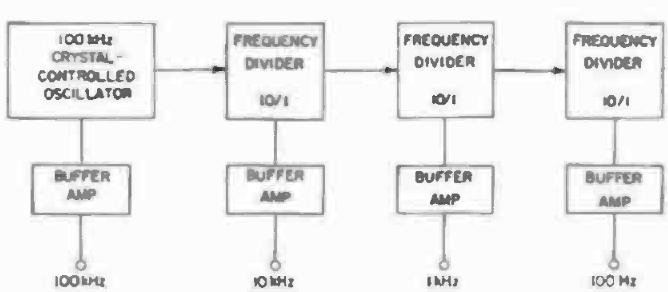


Fig. 22-40B. Elementary block diagram for using three frequency dividers.

henry inductor is placed in parallel with the 7.2- to 8.0-henry inductor on the 400-Hz range only. On the 300-Hz and 100-Hz ranges, the total value of the inductance is used.

The last frequency divider V6 produces an output only at 60 and 20 Hz. The output stage V7 consists of a single 6C4 cathode follower with a resistance voltage divider in the cathode circuit. The power supply is a simple half-wave selenium rectifier with an RC filter. With the circuits properly aligned, the line voltage may be varied over a range of 70 to 135 volts without losing synchronization, and with only a small drop in signal level.

A secondary standard of somewhat different design is given in Fig. 22-40B. Here the circuit also uses a crystal-controlled master oscillator, but uses only three frequency divider circuits in ratios of ten. Thus, only frequencies of 100 kHz, 10 kHz, 1 kHz, and 100 Hz are available. Other frequencies are obtained by the use of an oscilloscope and a second oscillator.

The fundamental divider circuit consists of a modulator-divider tube with a resonant circuit tuned to f/10 and a modulator-multiplier tube with a resonant circuit of 9f/10. The operation of the circuit can be explained by assuming a small voltage in the resonant circuit of the modulator tube. This voltage is applied to the grid of the modulator-multiplier tube. The two voltages mix to supply an output frequency of 9f/10, which is fed to the grid of the modulator tube, where it is mixed with the input control frequency f and results in a frequency of f/10 in the modulator-divider tuned circuit. This action is repeated and the voltage is built up until a stabilized condition is reached. The output of the divider unit is controlled by the input frequency.

By cascading the 100-kHz signal generated by the temperature-controlled, oscillating quartz crystal through three dividers, accurate fixed frequencies of 10 kHz, 1 kHz, and 100 Hz are also made available.

By the use of an external distorting amplifier and mixer circuit, harmonics may be obtained for frequency calibration to 20 MHz or higher, even though the waveform is sinusoidal. The accuracy of the foregoing instrument is 3 Hz per megahertz per degree Centigrade. The crystal-oscillator circuit may be adjusted over a range of plus or minus 8 Hz at 100 kHz to permit its precise adjustment to the primary standard of the National Bureau of Standards radio station WWV. (Also see Question 22.39.) The calibration of instruments in conjunction with secondary standards is accomplished by the use of a cathode-ray oscilloscope and Lissajous figures.

22.41 ***Describe the circuitry and operation of a flutter meter.*** — Flutter bridges or meters are instruments used for the measurement of irregularities in constant-speed drive systems, such as are used in photographic and magnetic recorders, telemetering, disc recording and reproduction, and other devices used for recording and reproducing. They measure both term (drift) and short-term (flutter) variations in the transport system.

Flutter meters are calibrated to read in terms of rms flutter, as specified in the existing standards. However, in some instances such instruments may be calibrated to read both rms and peak to peak and may include a small oscilloscope tube for reading peak indications and for observation of flutter waveforms.

The method used for the measurement of flutter is quite simple. A stable reference frequency is recorded on the media. During playback of this recording any deviations from the reference frequency will be directly proportional to short- or long-term speed errors in the transport system. These frequency errors are detected and read on a meter, displayed on an oscilloscope, or recorded on a graph level recorder. The present state of the art requires that 0.20-percent flutter or less be accurately measured. The measurement is accumulative over a specified bandwidth for each standard tape, film, or disc speed, as tabulated below:

1⅞ ips	0.2 to 313 Hz
3¾ ips	0.2 to 625 Hz
7½ ips	0.2 to 1.25 kHz
15 ips	0.5 to 2.5 kHz
30 ips	0.2 to 5.0 kHz
60 ips	0.2 to 10.0 kHz
120 ips	0.2 to 10.0 kHz

Speed variations slower than 0.2 Hz are generally specified as average speed error, and are generally caused by factors other than rotating mechanical components in the transport system or tape dynamics, which flutter tests are designed to indicate. Flutter signals are equivalent to fm signals of very low modulation index (MI = 0.002); therefore, severe demodulation problems arise in accurately recovering these signals from the playback signal. Under these circumstances, a high playback signal-to-noise level, extreme am rejection, and the separation and identification of tape dropouts are required. These latter subjects are discussed in Section 17. Flutter meters are calibrated to read in accordance with the existing NAB, DIN, SMPTE, IEEE, and USASI (ASA) Standards. Two types of flutter meters are available, a service type and a laboratory type, and they will be discussed in that order.

Pictured in Fig. 22-41A is a portable service-type flutter meter, manufactured by Sentinel, Inc. The block diagram for this instrument and its schematic diagram are shown in Figs. 22-41B and C, respectively. Referring to the block diagram, the instrument utilizes the basic principles involved in the detection of frequency-modulated signals. This includes the use of a limiter amplifier to prevent the amplitude modulation components from producing an indication in the output circuits. For example, amplitude modulation may be introduced by dropouts in magnetic

Fig. 22-41A. Sentinel Inc. Model FL-3D-1 flutter and wow meter.

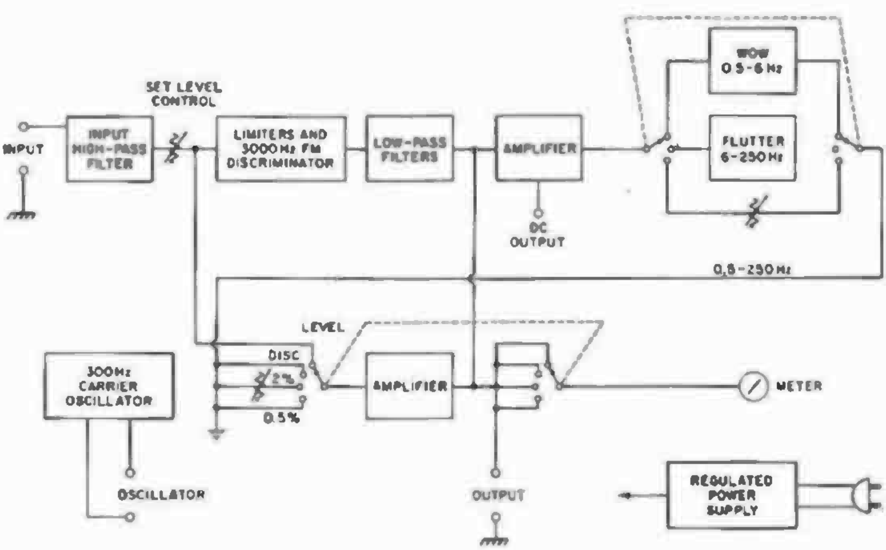


Fig. 22-41B. Block diagram for Model FL-3D-1 flutter meter.

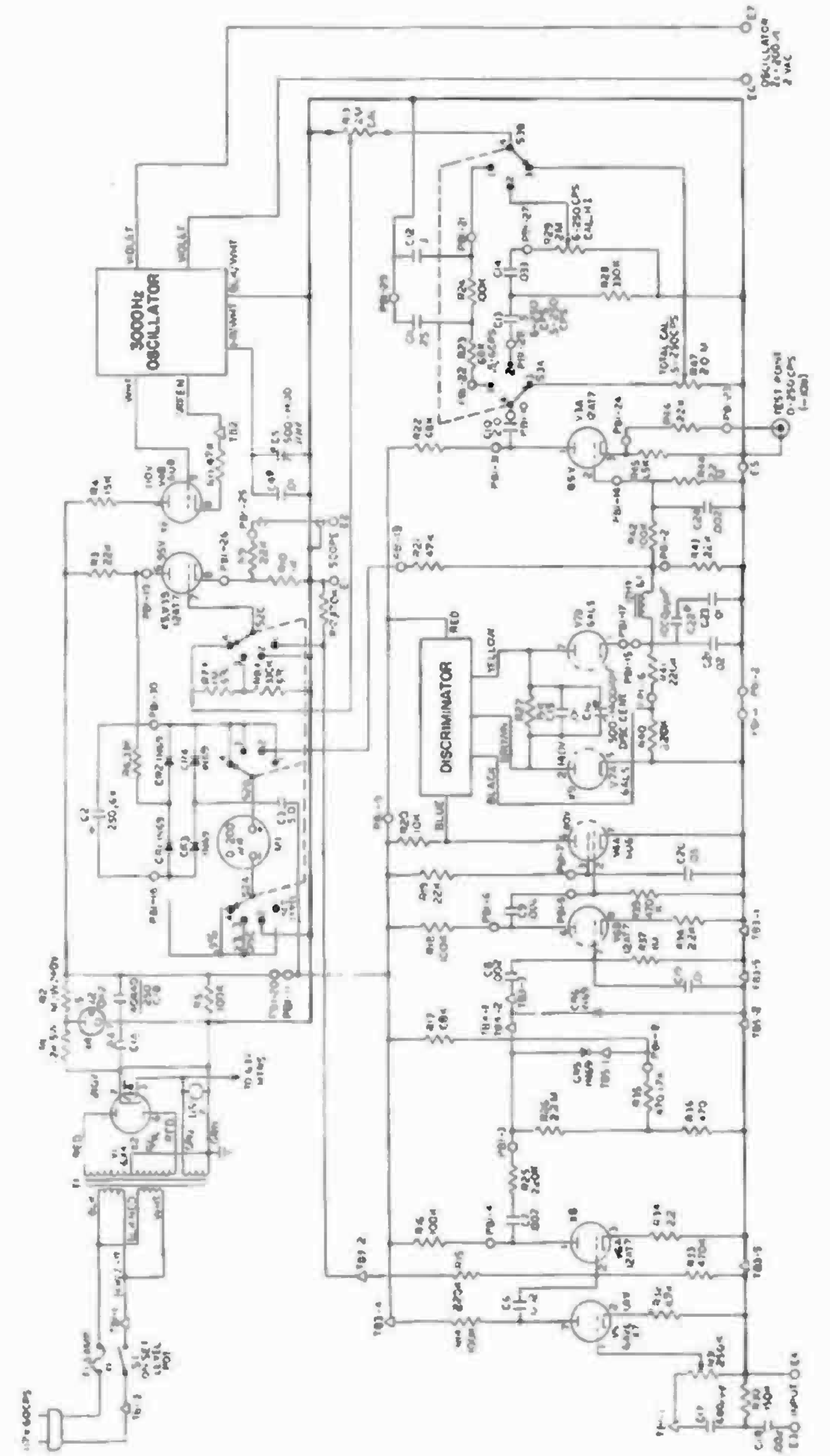


Fig. 22-41C. Schematic circuit for Sentinel Inc. Model FL-3D-1 flutter meter.

tape or film, clicks and pops in phonograph records, and light fluctuations in photographic film recordings. A frequency discriminator demodulates the flutter signals and presents them to an averaging-type meter circuit, calibrated to read the rms value of a sine wave. Suitable filters are provided to examine wow and flutter (wow is a once-around) spectra separately. A regulated power supply and a 3000-Hz internal carrier oscillator complete the circuitry.

Referring to Fig. 22-41C, a two-section, high-pass RC filter connects the input terminals with a set level control. The function of this filter is to remove low-frequency components from the incoming flutter-modulated 3000-Hz tone, reproduced by the machine under test. The network attenuates frequencies below 1000 Hz, and the signal is then amplified and fed to a symmetrical double-diode limiter. This circuit provides symmetrical clipping of the signal to avoid introducing phase-modulation components in the main signal because of changes in the zero-axis crossings associated with nonsymmetrical limiting.

The limited signal is amplified and fed to the pentode section of a 6U8, (V4A) which acts as an amplifier for a Foster-Seely discriminator. The detected flutter signal from a 6AL5 detector (V7B) passes through appropriate filters to remove the carrier signal. A front-panel control, discriminator centering, permits the discriminator secondary tuning to be adjusted for 3000 Hz, the frequency of the internal oscillator. When a prerecorded 3000-Hz signal is used, some adjustment of this control may be necessary. The bandwidth of the discriminator is such that flutter modulation of 250 Hz is attenuated no more than 3 dB from a reference frequency of 15 Hz. Sufficient response is available to identify frequencies to 350 Hz with an oscilloscope.

The demodulated flutter signal is amplified by V3A and fed to selector switch S3B, which permits the wow, flutter, or overall wow, plus flutter components to be measured. The signal is then applied to tube V3B. The cathode of this tube is connected through an isolating resistor to terminals E1 and E2 on the rear of the instrument for the connection of a graph level recorder or oscilloscope.

The filter networks used to separate the wow and flutter components consist of three-section RC filters. The crossover frequency is 6 Hz in accordance with existing standards. After the signal has passed through the filters, it is again fed to diode averaging circuit CR1, 2, 3, and 4. Full-scale meter sensitivities of 2 and 0.5 percent are provided in the meter scale. The selector switch for the flutter range also provides a position for monitoring the input signal for proper level setting and for connecting the meter across the discriminator to indicate the proper center-frequency adjustment.

The 3000-Hz internal oscillator is a standard Hartley oscillator circuit. Through the use of toroid coils and stabilizing techniques, a high degree of stability is had. The output signal of the oscillator is available on front panel terminals for recording purposes. Approximately 2.0 volts are generated at an impedance of 200 ohms.

Fig. 22-41D. Flutter meter Model 8100-W manufactured by Micom Corp.

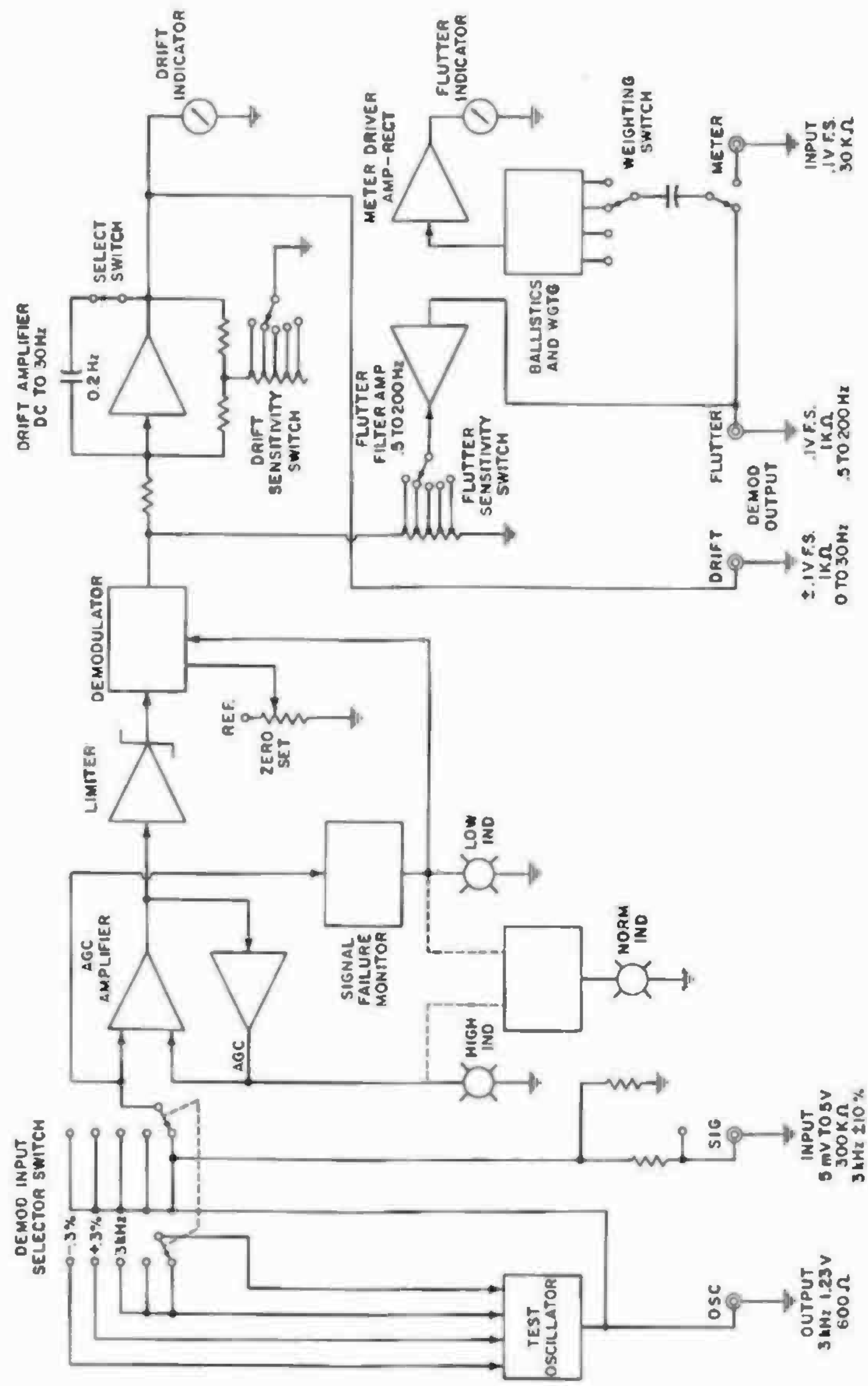


Fig. 22-41E. Block diagram for Micom Model 8100 laboratory-type flutter meter.

A regulated power supply, using an OA2 gaseous voltage-regulator tube ensures stable operation over line-voltage variations, plus or minus 15 percent from 117 Vac. The power transformer is designed to operate from either a 60- or 400-Hz source.

To operate the instrument a 3000-Hz signal is recorded, using the internal oscillator or taken from a special flutter tape disc, or film. With the function switch set to level, the set level control is adjusted until the meter reads midscale (0.25 on half-percent scale, or 1.0 on the 2-percent scale). The function switch is now set to discriminator, and the discriminator centering control is adjusted for a zero indication on the meter. The function switch is then turned to either the 2 or 0.5 flutter position for reading the percent of flutter. By the use of the bandwidth-selectiou switch, the frequency range of the disturbance may be determined. Flutter content is measured in terms of the rms frequency deviation expressed as a percentage of the average signal.

A laboratory-type flutter meter, Model 8100-W, manufactured by Micom, is shown in Fig. 22-41. This instrument is completely solid state and meets the United States and European Standards for flutter measurements. It measures both long-term (drift) and instantaneous variations (flutter). Models 8100 (without a wave analyzer) and 8100-W (with analyzer) are similar in design except for the addition of the analyzer section in the latter model. The major components of a Model 8100 are shown in the block diagram of Fig. 22-41E.

To measure drift a 3000-Hz signal from the device under test is applied to the agc amplifier, through a demodulator input selector switch. The amplified signal, now at a constant level, goes to a limiter and, after frequency doubling, to a demodulator. The demodulator converts frequency to voltage by means of a circuit which causes a precision capacitor to be charged to a reference voltage and discharged completely into a current-to-voltage amplifier once for each zero-crossing of the input signal. Zero output at frequencies other than 3 kHz is obtained by varying the voltage to which the capacitor is charged with a very high resolution zero-set potentiometer. This method of zeroing the instrument causes the output indications to accurately reflect deviations from the actual input frequency and not in terms of a percent-arbitrary 3 kHz.

The filtered output from the demodulator goes to an operational feedback amplifier, whose gain is controlled by changing the feedback factor with the drift sensitivity switch. The high-frequency cutoff of the amplifier is controlled by capacitive feedback and may be switched from 0.2 Hz to 30 Hz. The operational amplifier drives the drift indicator and, after filtering, provides a drift demodulated signal of 0.1 volt for a full-scale reading of the meter.

The output of the demodulator also goes to the flutter sensitivity switch, and then to a flutter amplifier and an active-filter amplifier. The sensitivity switch attenuates the input in the 10-, 3-, and 1-percent ranges and sets the gain by feedback control for 0.3, 0.1, 0.03, and 0.01 positions. The filter-amplifier has a flat bandpass from 1.5 to 200 Hz, with a fast rolloff beyond 200 Hz. The residual noise is less than 0.0005 percent rms at the output of the flutter amplifier.

The flutter-demodulator output goes to the demodulator position of the meter select switch, and then to a weighting switch. This switch controls gain, frequency response, and the mode of driving the flutter meter. In the NAB position, the meter exhibits the standard VU characteristic response and is sensitive to the average of the signal after full-wave rectification. The meter is calibrated in terms of rms value of a sinusoidal waveform having an average output. In both DIN positions, the meter reads plus or minus peaks with appropriate charge and discharge time constants to meet the DIN specifications (the abbreviation DIN refers to the German magnetic tape standard used in Europe).

The meter-driving amplifier is an operational amplifier, with a full-wave diode-bridge output. Feedback compensation effectively removes the nonlinear and thermal effects of the diode bridge, resulting in a linear meter scale at the lower portion of the scale. Weighted positions for both NAB and DIN have a frequency response as given in Fig. 17-144B. Signal monitor lamps indicate whenever the signal voltage drops below 3.5 mV for longer than 20 milliseconds.

A highly stable 3000-Hz test oscillator supplies 1.23 volts from a 60-ohm source. When the demodulator switch is in the test oscillator position, the test oscillator is selected as the input signal. A circuit is included to provide plus and minus 0.3-percent frequency deviations of the oscillator with a high-degree of accuracy. A regulated power

Fig. 22-42A. Allison Model 22 clinical and research audiometer.

supply provides plus 45 Vdc, plus or minus 0.1 volt, and several lower voltage taps. The plus 45 volts is also the reference for the frequency-to-voltage demodulator and is derived from a temperature-compensated zener reference diode.

The wave analyzer, not shown in the diagram, is based on an active bandpass filter, in which the tuning of the filter is proportional to two RC products. The active portion of the circuit is stabilized by feedback. The selectivity is approximately 0.1 octave, with a frequency dial calibrated to read logarithmically. The output of the wave analyzer is indicated on the flutter meter.

22.42 Describe an audiometer?—It is an instrument used for measuring the acuity of hearing. Basically the device consists of an oscillator and calibrated attenuator, so that the ear characteristics may be plotted logarithmically. For the smaller-type audiometers, the test frequencies used are 125, 250, 512, 1000, 2000, 4000, and 8000 Hz. Tests conducted over a period of many years indicate that the sensitivity of the human ear changes with respect to frequency and sensitivity. This is shown graphically in Fig. 1-99.

A clinical and research laboratory-type audiometer, manufactured by Allison Laboratories, Inc., is pictured in

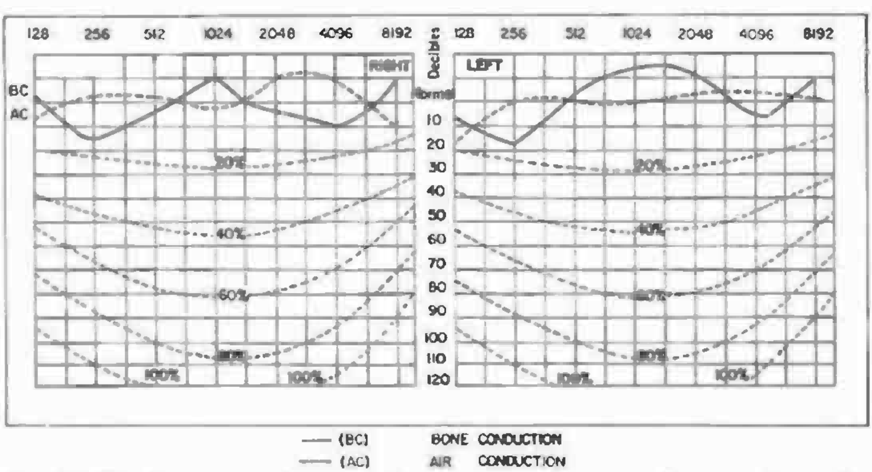


Fig. 22-42B. Human ear characteristics as measured by an audiometer. Curves for conduction (BC) and air conduction (AC) are shown.

Fig. 22-42A. This unit embodies facilities for pure tone and speech testing. An automatic attenuator-drive system permits patient-controlled audiograms, alternate binaural loudness-balance tests, and discrete-frequency audiograms. The instrument also has facilities for special tests, such as delayed speech, special recorded tests, binaural hearing-aid evaluation, white-noise, sound field using warble tone, small-increment sensitivity index, sensorineural acuity level, Rainville test, and many others. Galvanic skin reaction equipment may also be used. A quarter-inch tape recorder/reproducer and record turntable are mounted in the drawers at the right. A talkback system is also included for communication between the patient and technician.

External equipment may be added to provide continuously variable band-pass filters, narrow-band masking filters, headphones, and an automatic graph recorder for use with the automatic attenuator-drive system. Push-button control provides a means for selecting the various functions. Frequencies that are available include 125, 250, 500, 750 hertz, and 1.5, 2, 3, 4, 6, and 8 kilohertz. The instrument is installed in a soundproof room (called a "quiet room") with a glass panel, similar to a recording studio. The console is installed outside the quiet room, at a position where the technician can see the patient.

Typical audiograms for both bone-conduction (BC) and air-conduction (AC) are given in Fig. 22-42B. Calibrations for this instrument meet the ISO Standards.

22.43 ***Describe an audio-frequency microvolter.***—It is an instrument with a calibrated attenuator or voltage divider which, when used with a suitable oscillator, supplies accurately known audio-frequency voltages. The microvolter converts the voltage of an oscillator into a standard signal which may be used for making measurements of gain or loss, frequency characteristics, overloads, and hum level on amplifiers, networks, and other audio equipment. The combination of an oscillator and microvolter is very useful in making injection frequency measurements on pickups, microphones, magnetic recording heads, and other transducers requiring minute voltages. (See Questions 23.51, 23.52, and 23.53.)

Basically the microvolter consists of a constant-impedance attenuator and an input voltmeter by which the input to the attenuator is standardized. A switch controls the output voltage in decade steps, while an individually calibrated dial provides continuous control over each decade.

A Model 546-C audio-frequency microvolter, manufactured by the General Radio Co., is shown in Fig. 22-43A. Only two controls are provided: the output

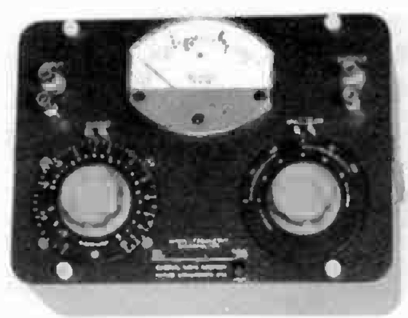


Fig. 22-43A. General Radio Co., Model 546-C audio frequency microvolter.

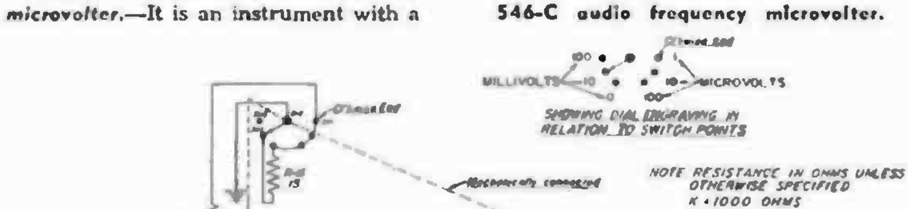


Fig. 22-43B. Schematic diagram for General Radio Co., Model 546-C audio frequency microvolter.

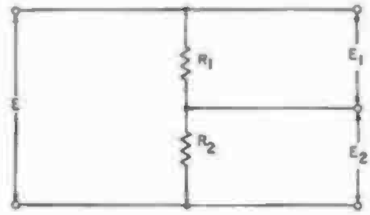
dial and the output multiplier. The output dial is a continuously rotatable knob and a dial which is calibrated in voltage and decibels above 1 microvolt. The voltage scale is approximately logarithmic, and the decibel scale is approximately linear. The multiplier control is a six-position switch, with a dial calibrated in decade steps from 1 microvolt to 100 millivolts, and in 20-dB steps from 0 to 100 dB. The meter indicates in both decibels and voltage. The schematic diagram appears in Fig. 22-43B.

The output voltage ranges from 0.5 microvolt to 1.0 volt, with 2.2 volts at the input. Reducing the input voltage to 1.1 volts reduces the output voltage by half. The input and output impedance is 600 ohms.

To operate the meter a voltage of the desired frequency is applied to the input and the meter is set to indicate 2.2 volts (0 dB). The output and multiplier dials are set so that the product of their voltage settings equals the desired output voltage. For an output voltage of 50 microvolts (0.0005 V) the output dial is set to 5 and the multiplier dial is set to 10 microvolts; for 70 millivolts, the output dial is set to 7, and the multiplier set to 10 millivolts. Output, in terms of decibels above 1 microvolt is denoted by the algebraic sum of the decibel indications on the two dials. Since the reference is 1 microvolt, almost all readings will be positive. When used to supply a 600-ohm load, the output in dBm equals the output in decibels above 1 microvolt, minus 123.8 dB. For the general run of measurements, the output may be considered to be a 600-ohm generator, having open-circuit voltages as indicated by the dial.

22.44 ***What is a decade box?***—An instrument containing resistance, inductance, or capacitance which may be varied in value in decade steps. Decade boxes may be obtained covering a resistance range from a fraction of an ohm to several thousand ohms. Capacitive and inductive decade boxes are also obtainable having values ranging from a fractional amount up to several microfarads or henries.

22.45 ***Describe a high-voltage resistive-type voltage divider?*** — A circuit consisting of a resistive, capacitive, or inductive network, or a combination of such elements. The circuit elements are connected in such a manner that the potential in the circuit is reduced to a given amount. A typical voltage divider circuit is shown in Fig. 22-45A.



$$I = \frac{E}{R_1 + R_2}$$

$$E_2 = E\frac{R_2}{R_1 + R_2}$$

$$E = E_1 + E_2 = IR_1 + IR_2$$

$$E_1 = E\frac{R_1}{R_1 + R_2}$$

Fig. 22-45A. Resistive voltage divider. Voltage division is the same for either ac or dc.

Such devices are used with vacuum-tube voltmeters. A typical resistive-type voltage divider, manufactured by Ballantine Laboratories, is shown in Fig. 22-45B. The divider has a total resistance of 44 megohms and a division ratio of 10:1.

22.46 ***What is the advantage in using a capacitive-type voltage divider over a resistive type?*** — The capacitive type requires practically no current for its operation compared with a resistive voltage divider. A typical capacitive-type voltage divider is shown in Fig. 22-46. Capacitive divider circuits are used only in ac circuits, and they are frequency sensitive.

Fig. 22-45B. High-voltage, 44-megohm, resistive voltage divider manufactured by Ballantine Laboratories.

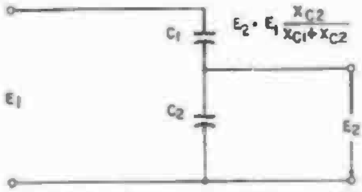


Fig. 22-46. Simple ac voltage divider consisting of capacitance only.

22.47 What are the basic requirements for an audio oscillator? — Audio oscillators are designed to generate a pure or nearly pure sine wave at frequencies of 20 to 20,000 Hz. However, many audio oscillators cover a range from 10 to 40,000 Hz, and in some instances up to 1 megahertz.

Audio oscillators may be of the beat-frequency type (see Question 22.51), Wien bridge (Question 22.50), or parallel-T configuration (Question 22.49). The Wien bridge is sometimes referred to as a "negative-resistance oscillator." Audio oscillators may be designed to cover the audio spectrum in one sweep of the dial or in several decades. Such oscillators must have low total harmonic distortion (THD) and essentially constant output impedance with a constant voltage output. The total harmonic distortion should not exceed 1 percent and preferably 0.25 percent or less.

22.48 Why must audio oscillators have low harmonic distortion? — Audio oscillator cannot be directly subtracted many different devices, such as amplifiers, filters, equalizers, and recording systems. As most of these components are sensitive to harmonic distortion, the test oscillator must have low internal harmonic distortion (THD). As an example, modern high-quality amplifier systems often have less than 0.25 THD; therefore, if the test oscillator has greater distortion than the device under test, an accurate measurement of the distortion characteristics cannot be made. Distortion contributed by an oscillator cannot be directly substracted from the measured distortion. Oscillator distortion, unless unusually high, does not as a rule affect frequency-response measurements.

To illustrate the effect of oscillator distortion, a measurement was made at 400 Hz on a 10-watt amplifier having a THD of 0.35 percent, using an oscillator with a maximum of 0.1-percent THD. A second measurement was then made on the same amplifier, but using an oscillator with 1.3- to 2.5-percent THD. The results of these measurements are given in Fig. 22-48. The low-distortion oscillator remained at 0.10-percent THD for all levels of measnrement; however, the high-distortion oscillator varied from 1.5- to 2.5-percent THD for the different levels of measurement. It will be noted in the second measurement that all that was measnred is the oscillator distortion which changes for different settings of its gain control.

When an oscillator with low-distortion characteristics is not available, a bandpass filter may be connected between the output of the oscillator and the input to the device under test. In this manner a sine wave of low distortion may be generated, provided all impedance matches are satisfied. However, this limits the measurement to one frequency, unless several filters are available.

22.49 Describe the basic principles of a bridged-T (RC) audio oscillator.— The basic circuit of such oscillators is the well-known RC bridged-T configuration. (See Section 5.) The amplitude of the oscillator is stabilized by a large amount of negative and positive feedback in conjunction with a voltage-amplitude-sensitive element in the negative-feedback loop. Various combinations of resistance and capacitance in the bridged-T circuit provide a means of generating a wide range of frequencies.

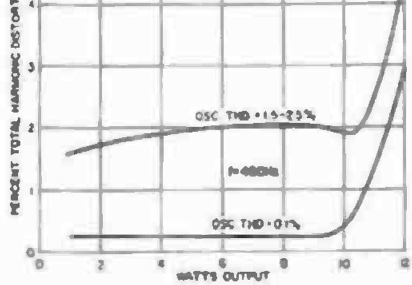


Fig. 22-48. Distortion characteristics of a 10-watt amplifier using two oscillators. The lower plot was made using an oscillator having a THD of 0.10 percent. The upper curve used an oscillator with 1.3 to 2.5 percent THD.

Fig. 22-49A. Front panel view of Waveforms Inc. Model 401B audio oscillator.

A front-panel view of a bridged T audio oscillator, Model 401B, manufactured by Waveforms, Inc., is given in Fig. 22-49A. Referring to the schematic in Fig. 22-49B, the oscillator section consists of a broad-band amplifier stage employing a sharp-cutoff 6AU6 pentode V1, a bridged-T network consisting of a group of resistors connected to range switch S1, and four-section variable air capacitor C1, attached to a front-panel dial and calibrated from 10 to 100.

The oscillator circuit really consists of two amplifier tubes, V1 and cathode follower V2A. Oscillation is caused by feeding a positive feedback voltage (regenerative) from the cathode of V2A through the tungsten filament of a 10-watt, 230-volt pilot lamp (amplitude-sensitive element) to the cathode of V1. A large amount of negative feedback (degenerative) is also taken from the cathode of V2A and applied to the bridged-T network and back to the control grid of V1. The negative feedback prevents V1 from oscillating at any frequency except at the frequency of the bridged-T network. At this one frequency the negative feedback is at a minimum and the phase shift is zero.

Any tendency on the part of the oscillator circuit to produce frequencies of varying amplitude is effectively controlled by the lamp in the cathode circuit of V1. As the output voltage of V1 increases, more current is drawn through the lamp filament increasing the temperature and thus increasing the resistance. The increased resistance decreases the amount of positive feedback coupled to the cathode of V1, thereby reducing the output. When the output signal decreases, the temperature of the lamp is lowered, causing the lamp resistance to decrease. This, in turn, permits more positive feedback voltage to be applied to the cathode of V1, increasing the amplitude of oscillation. The output of V1 is controlled by the amount of resistance in the cathode circuit (which is generally a factory adjustment).

The frequency of oscillation is determined by the values of resistance and capacitance in the bridged-T configuration and consists of the resistors selected by the range switch and the amount of capacitance of the air capacitor. It will be observed that the variable capacitor consists of a four-section unit, with two sections connected in parallel and connected to the opposite end of the T-section of the bridged-T configuration. The common connection of all rotors is connected to the range switch, thereby inserting resistauce in series with the capacitance to form the stem of the bridged-T circuit.

Position 1 of the range switch generates frequencies of 10 to 100 Hz; position 2, 100 to 1000 Hz; position 3, 1000 to 10,000 Hz; and position 4, 10,000 to 100,000 Hz. The rejection frequency for a given value of capacitance and resistance may be calculated:

$$\text{Frequency} = \frac{1}{2\pi RC}$$

where,
C is equal to $\sqrt{C1 \times C2}$,
R is equal to $R1 = R2$.

The output of the oscillator is taken through a 0.10-μF capacitor to the control grid of a dual triode V2A and V2B, connected as a self-driven, single-ended, push-pull cathode follower, providing a low-impedance output circuit. The upper section is similar to a conventional cathode follower, and the lower section is driven out of phase, serving as a dynamic load.

A terminal for external synchronization purposes is provided on the front panel. This connection provides a high-impedance ac sine-wave signal of ap-

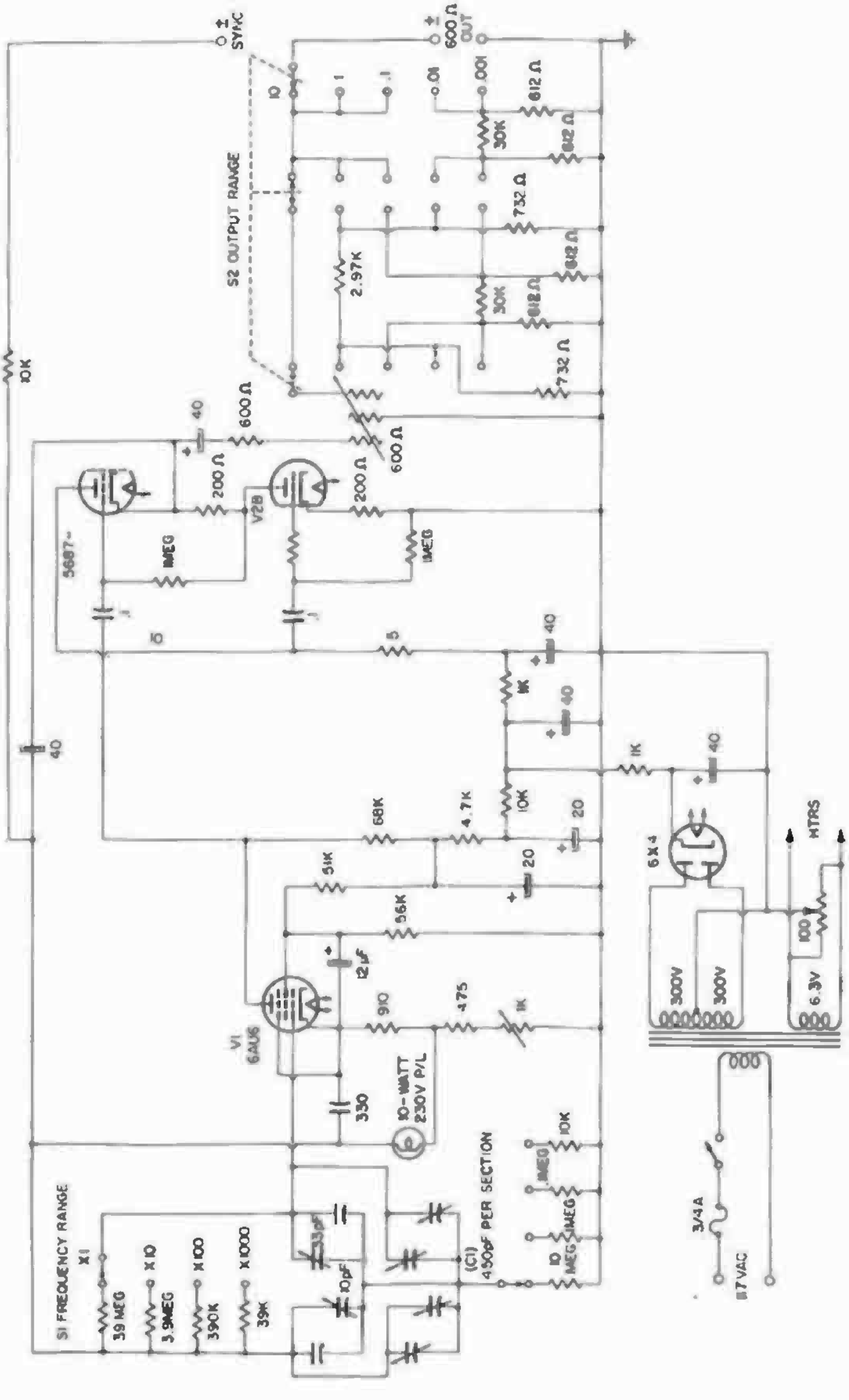


Fig. 22-49B. Schematic diagram for Waveforms Inc. Model 401B bridged-T (RC) audio oscillator.

proximately 24 volts rms, and it is independent of the output controls. This connection must be used only with a high-impedance and resistive loads, since reactive loads will affect the frequency calibration and may introduce distortion. Furthermore, the external circuit to which it is connected must not have an ac component.

The output attenuator section consists of five positions, each position reducing the output by a factor of 10 (20 dB). The maximum output is 22 dBm. Total harmonic distortion is 0.25 percent for 10 to 100,000 Hz. The internal noise and hum is 80 dB below the maximum output. The internal output impedance is 600 ohms resistive.

Although an oscillator may be rated as 600 ohms and will operate satisfactorily with this load, the internal output impedance may be on the order of 35 to 50 ohms. Therefore, when an instrument of this nature is used with equipment that may be impedance sensitive, an impedance-matching or taper pad should be used between the oscillator and the equipment being measured. In some instances a resistance equal to the difference between the oscillator internal output impedance and the input impedance of the device under test may be connected in series with the oscillator output to increase the oscillator output impedance. If the device being tested employs a bridging input impedance, the oscillator may be terminated in 600 ohms and the device fed from this 600-ohm source without difficulty.

22.50 Describe the circuitry for a Wien-bridge or negative-resistance oscillator. — A Wien-bridge or negative-resistance oscillator is composed of a tuned-bridge circuit, consisting of capacitance and resistance, as shown in the basic circuit of Fig. 22-50A. It is the basis of the Hewlett-Packard Model 200AB oscillator, pictured in Fig. 22-50B, and several other models.

Referring to Fig. 22-50A, the oscillator is basically a two-section resistance-coupled amplifier consisting of two tubes, V1 and V2. Two feedback loops are employed around this amplifier—positive feedback to set up oscillation and a negative-feedback loop to reduce distortion and hold the oscillator amplitude constant. The positive-feedback loop contains five resistors and a variable air capacitor so proportioned

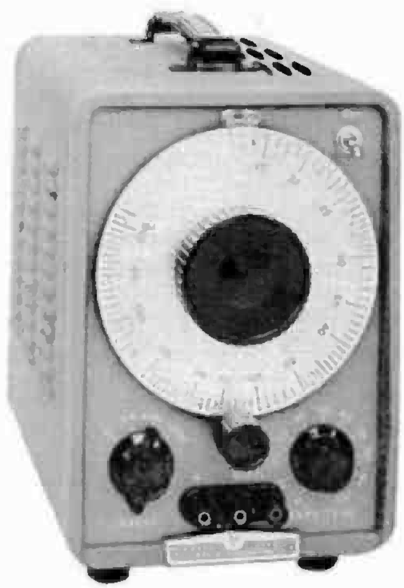


Fig. 22-50B. The Hewlett-Packard Model 200AB Wien-bridge oscillator.

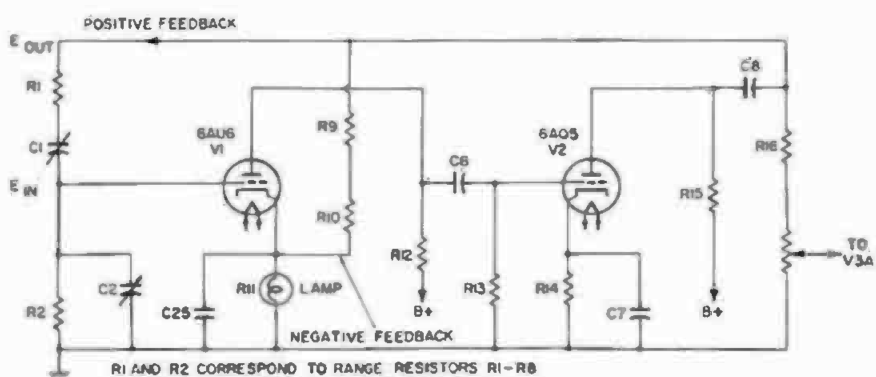


Fig. 22-50A. Basic circuits of a Wien-bridge or negative-resistance oscillator.

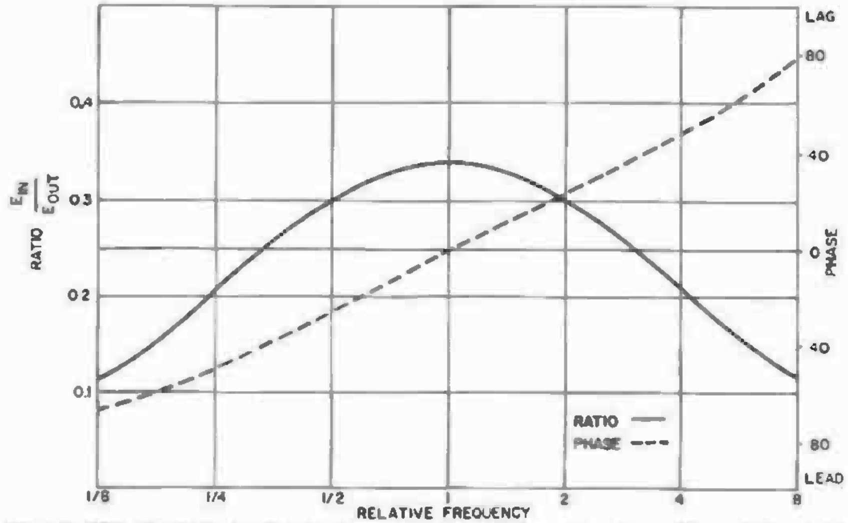


Fig. 22-50C. Positive-feedback characteristics of the network used in a Wien-bridge or negative-resistance oscillator.

that R1C1 equals R2C2. The oscillator output voltage is applied to the network, and the oscillator input voltage is derived from it.

Since both the input and output signal of the two-stage amplifier are in phase, oscillation will occur when the phase shift between the voltage applied to the network and the voltage at the control grid of V1 is zero. The zero phase-shift point is also the point of minimum loss through the feedback network, as indicated by the curves in Fig. 22-50C. The frequency of oscillation (relative to frequency) equals 1.0 and is equal to:

$$½ \pi \sqrt{R1C1R2C2} \text{ or } ½ \pi RC$$

where,

R1 equals R2, C1 equals C2. Cathode bypass capacitors C25 and C7 are for correcting phase shift at the higher frequencies.

The negative-feedback network in the oscillator section minimizes changes in oscillator amplitude with changes in frequency. Incandescent lamp R11 is used as a bias resistor and is also a part of the negative-feedback voltage divider circuit in the first stage of oscillator V1. The lamp (10-watt, 250-volt pilot lamp) has a temperature characteristic such that its resistance will increase in direct proportion to the voltage across it. Changes in resistance of this lamp will change the percentage of negative feedback in the oscillator circuit. As the oscillator voltage rises, more voltage is applied to the lamp. The increased voltage raises the temperature and resistance, which in turn increases the negative feedback in the oscillator circuit. Increasing the feedback tends to decrease the oscillator output voltage to its normal operating point, and vice versa. The thermal resistance of the lamp is such that it does not vary with low-frequency sine-wave oscillations.

Referring to Fig. 22-50D, an overall schematic diagram of the Hewlett-Packard oscillator, the output of the oscillator section is RC coupled to the input of tube V3A, which in turn is direct coupled to a phase-inverter tube, V3B. The phase inverter drives output tubes V4 and V5. Output transformer T1 includes a tertiary winding, which is carried back to the cathode circuit of V3A.

The cathode-bias resistor (R19) of V3A is not bypassed, giving additional negative feedback. Since the total amount of negative feedback in the amplifier section is over 30 dB, little distortion is introduced by this section in the output. The power supply is conventional and requires no explanation. The internal output impedance of the instrument is approximately 35 to 50 ohms up to 10,000 Hz, 75 ohms at 20,000 Hz, increasing to about 250 ohms at 40,000 Hz.

The frequency dial is calibrated to read 20 to 200, operating in conjunction

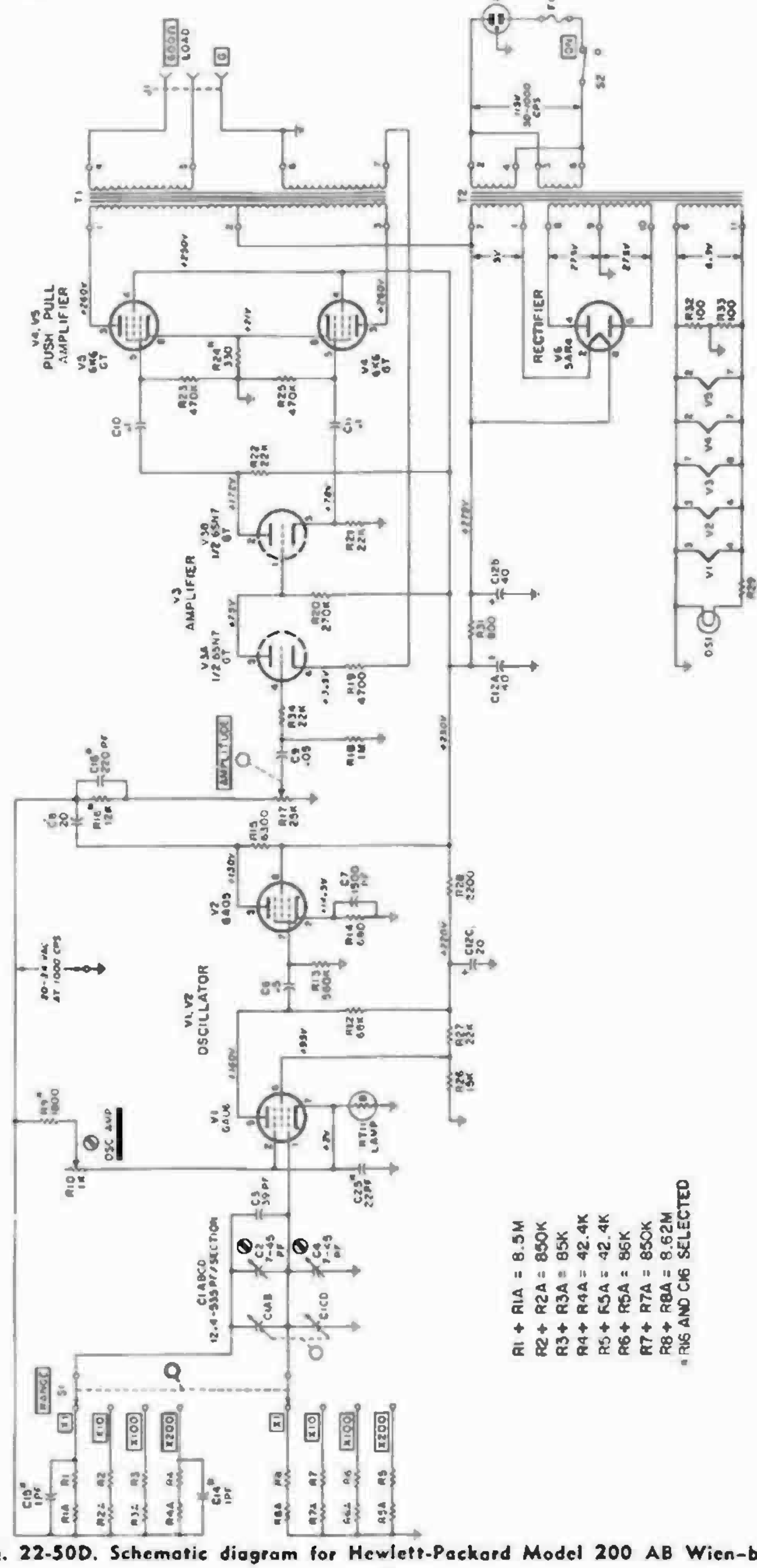


Fig. 22-50D. Schematic diagram for Hewlett-Packard Model 200 AB Wien-bridge oscillator.

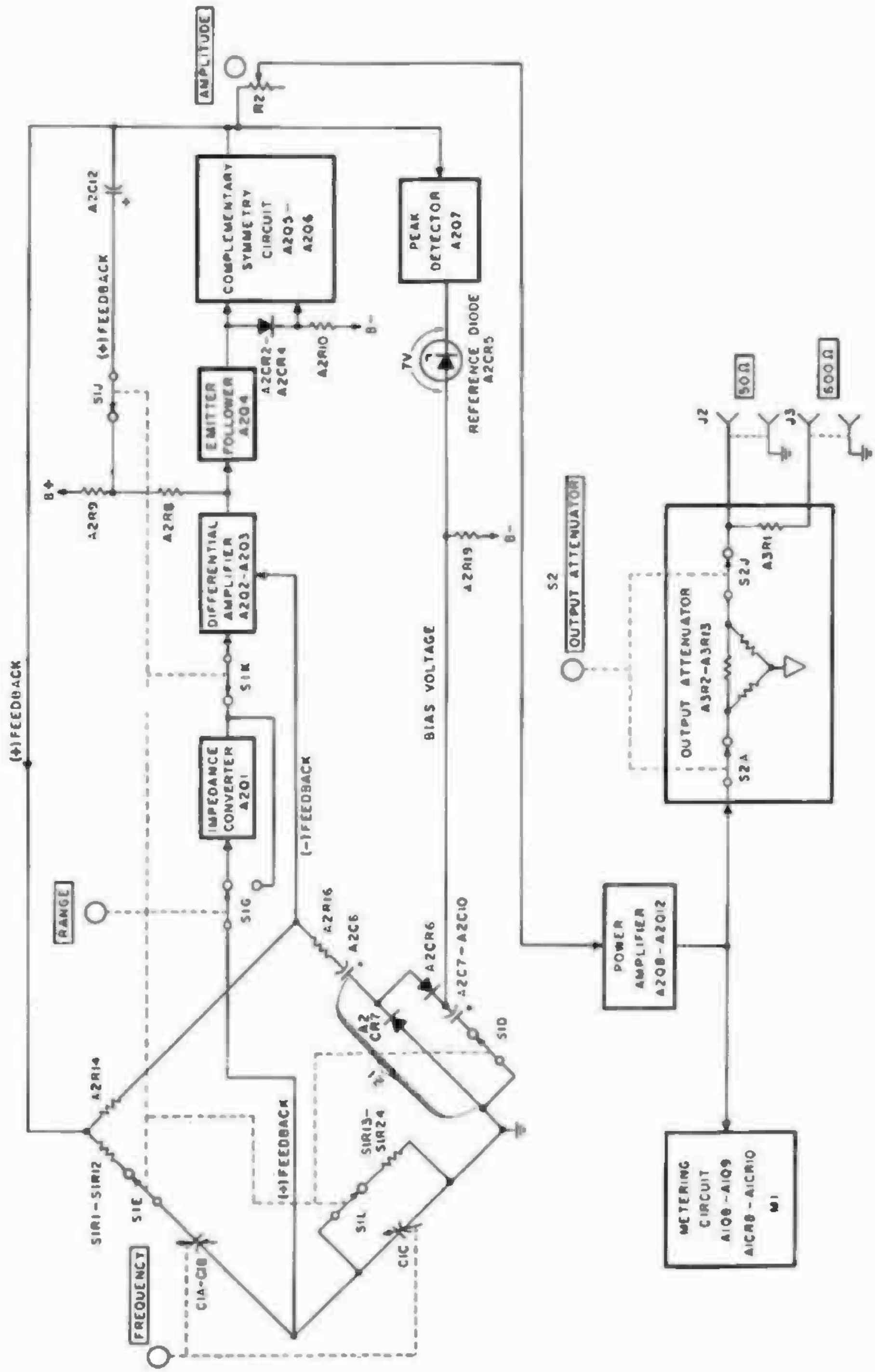


Fig. 22-50E. Simplified block diagram for Hewlett-Packard Model 651 oscillator.

with a four-position range switch. In the first position the frequency range is 20 to 200 Hz; the second, 200 to 2000 Hz; the third, 2000 to 20,000 Hz; and in the fourth position the range switch is multiplied by 200 and reads 400 to 40,000 Hz. The THD is less than 1 percent from 20 to 20,000 Hz. Internal noise is 66 dB below the rated output of 1 watt, or 24.5 volts into a 600-ohm load. The output may be used with symmetrical balanced or unbalanced circuits.

A block diagram for an oscillator with a wide frequency range (10 Hz to 10 MHz), the Model 651, also manufactured by Hewlett-Packard, is shown in Fig. 22-50E. The output signal is adjustable from 10 microvolts to 3.16 volts for either a 50- or 600-ohm output. The internal circuitry is all transistor and includes an oscillator, power amplifier, peak detector, attenuator, and output meter circuit.

This oscillator employs a modified Wien-bridge circuit to generate a low-distortion sine-wave signal, which is applied to a power amplifier circuit using a complementary-symmetry circuit. (See Question 12.251.) A peak-detector circuit provides a degenerative feedback voltage to the oscillator for stabilizing the signal applied to the power amplifier. The output attenuator network provides a means of attenuating the signal in steps of 10 dB each. The metering circuit continuously monitors the signal level to the attenuators. A regulated power supply completes the sections. Because of the current required by the lamp in the cathode circuit of the Model 200AB (Fig. 22-50A), a lamp would be incompatible for transistors; therefore, a peak-detector circuit is used to stabilize the oscillator output in this model.

The oscillator of this instrument generates a sinusoidal signal at the frequency selected by the range switch and the frequeucy dial. The RC bridge network (Fig. 22-50E) is a modified Wien bridge and differs from the conventional Wien bridge in the design of the resistive divider network. This difference is shown in the lower portion of the bridge circuit, where the usual resistor is replaced with an impedance, Z1.

Oscillation at a selected frequency is made possible by the use of both regenerative and degenerative feedback. Positive feedback is provided through a frequency-sensitive RC network to the differential amplifier, A2Q2 and A2Q3; negative feedback is provided to the differential amplifier through a network insensitive to frequency. Only at the selected frequency will the positive feedback exceed the negative feedback voltage to sustain oscillation.

Range switch S1 selects combinations of resistors S1R1 through S1R24 to establish the frequency-sensitive RC networks for the six frequeucy ranges. The frequency dial varies the main frequency-tuning elements, C1A, C1B, and C1C. The RC components maintain relationship of the oscillator output voltage. When X_c equals R, the positive feedback is in phase with the oscillator output voltage and exceeds the negative-feedback voltage. At frequencies other than X_c equals R, the positive-feedback voltage is neither of the right phase nor of sufficient amplitude to maintain oscillation.

Impedance-converter transistor A2Q1 provides high impedance in series with the input impedance of the differential amplifier on the first four frequency ranges (×10 to ×10K). The high impedance added prevents the RC bridge circuit from being loaded by the low impedance of the differential amplifier on the lower frequency ranges. The impedance converter is bypassed on the ×100K and ×1 Meg ranges due to lower resistance values in the RC bridge.

The difference between the feedback voltages from the hridge circuit is amplified by the differential amplifier and is applied to a complementary-symmetry circuit, A2Q5 and A2Q6, through emitter follower A2Q4. Positive-feedback voltage from the output of the complementary-symmetry circuit is applied between resistors A2R8 and A2R9 in the collector circuit of A2Q2 of the first four frequency ranges. The application of the feedback voltage at this point is used to make the effective resistance of the collector load higher than the input impedance of the emitter follower, thus increasing the signal level at the base of the emitter follower. The increase in signal level results in an increase in the loop gain of the emitter follower at the higher frequencies.

The complementary-symmetry circuit provides power gain; also, its low impedance prevents the oscillator output from being loaded by the RC hridge. The transistors in this circuit are in a slight state of conductance to reduce crossover distortion. The output of the oscillator drives the power amplifier section with a constant voltage, which is set by the amplitude control, R2, and held constant by the action of the peak detector.

The peak-detector circuit provides a bias voltage. Peak detector A2Q7 conducts only on the positive peaks of the oscillator signal. When the positive peak exceeds the set level, the peak detector conducts, breaking down reference diode A2CR5. This causes the diode to conduct less, increasing the dynamic resistance and thus increasing impedance Z_1. The increase of Z_1 increases the amount of negative-feedback voltage to the differential amplifier, which results in a reduction of the oscillator output signal. The reduction in signal level compensates for the initial rise in the oscillator output.

The metering circuit monitors the signal level applied to the output attenuator circuit and provides a front-panel readout of the signal in rms volts and

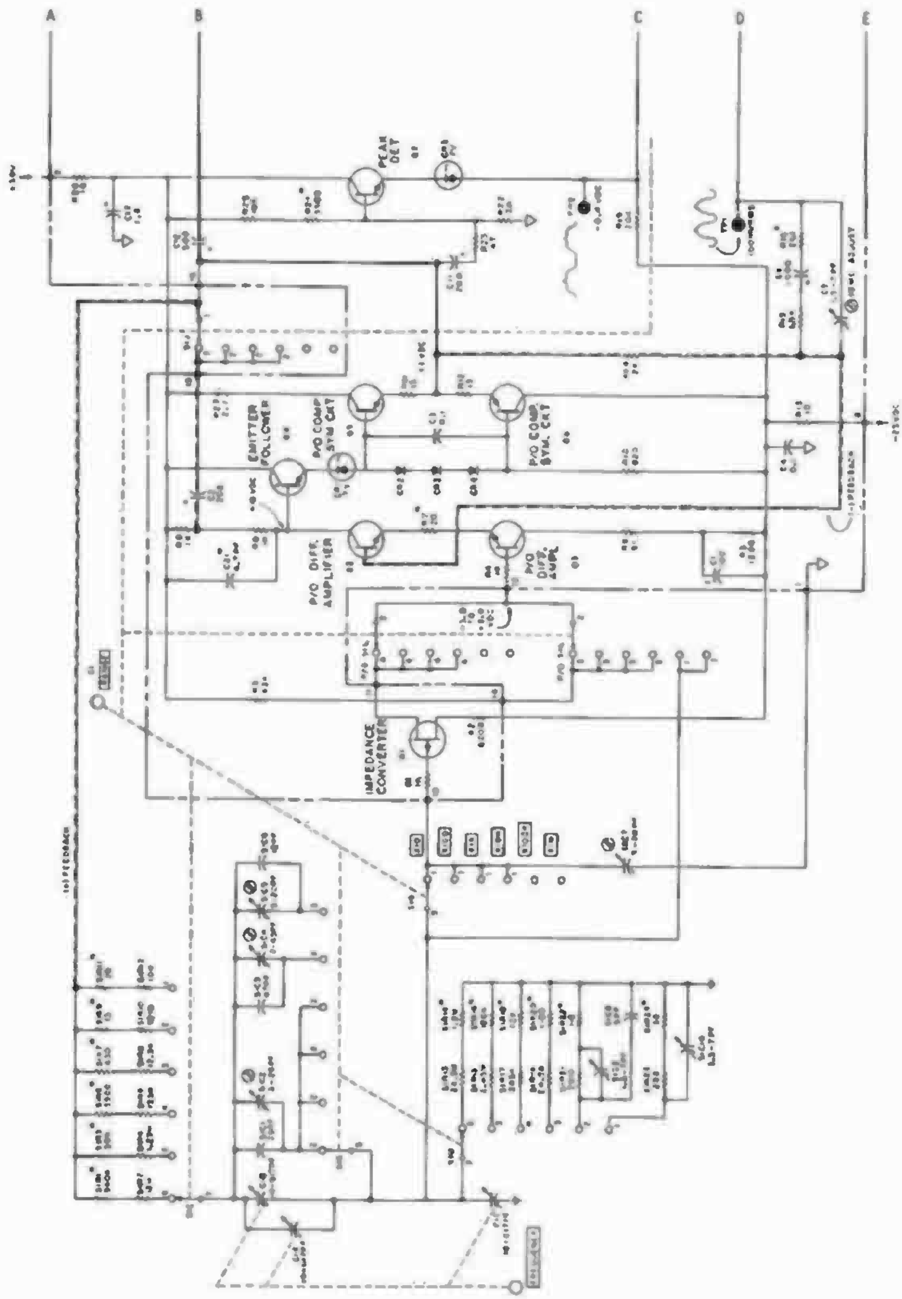
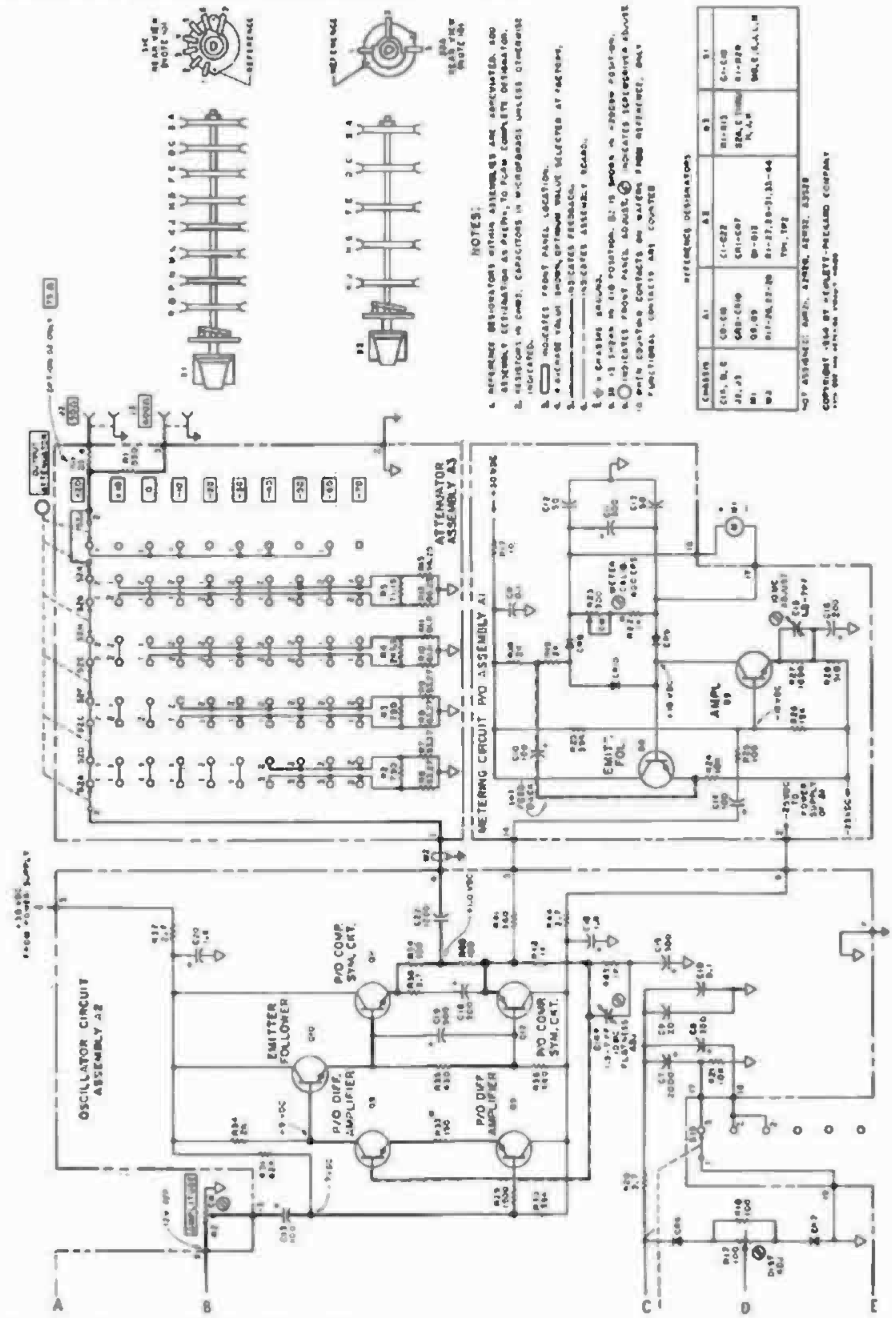


Fig. 22-50F. Schematic diagram for

dBm. Amplifier A1Q9 serves as both an impedance converter between the voltmeter circuit and the power amplifier output circuit, and as a current source to provide full-scale meter deflections. The output attenuator provides a means of attenuating the signal level applied to both the 50- and 600-ohm outputs, by selecting a group of delta networks to produce the desired attenuation in steps of 10 dB. A source of constant voltage for the oscillator and other circuits is supplied from a 30-volt regulated power supply. Similar power supplies are discussed in Section 21. The complete schematic diagram for this oscillator and amplifier section is given in Fig. 22-50F and the power-supply schematic is shown in Fig. 22-50G.



Hewlett-Packard Model 651 oscillator.

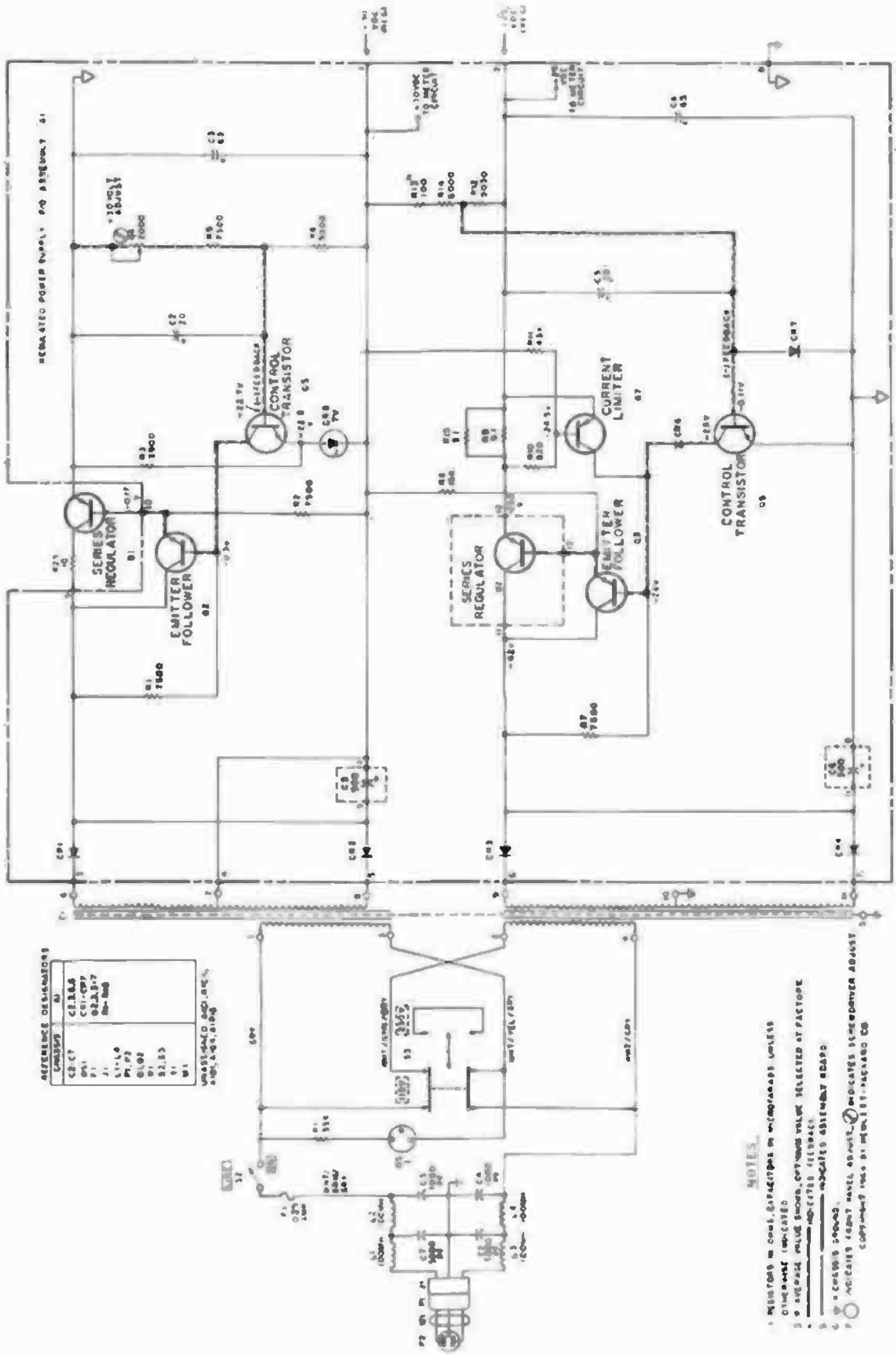


Fig. 22-50G. Schematic of power supply used with Hewlett-Packard Model 651 oscillator shown in Fig. 22-50F.

22.51 ***What is a beat-frequency audio oscillator?***—It is an audio oscillator employing two radio-frequency oscillators, one generating a fixed frequency and the other a variable frequency. The variable-frequency oscillator is connected to a dial on the front panel.

Fig. 22-51A. Brüel and Kjaer Model 1022 beat-frequency oscillator.

In operation, the two oscillators operate on the heterodyne principle and generate a beat frequency which falls in the audio-frequency band. This beat frequency becomes the output signal of the generator. The oscillators generally consist of highly stabilized LC tuned circuits. When the dial of the variable oscillator is rotated its frequency is changed, thereby changing the beat frequency. A comparatively small change in the frequency of the variable oscillator will cause a large change in the beat frequency. The principal advantage of the beat-frequency oscillator is that it may be swept across the whole audio-frequency spectrum, 20 to 20,000 Hz, with one movement of the dial. This makes it desirable for use with graphic recorders, wave analyzers, and other devices that must operate in synchronization with the oscillator frequency.

Pictured in Fig. 22-51A is such an oscillator, manufactured by Brüel and Kjaer of Copenhagen, Denmark. In this instrument are included several features not generally found in instruments of this type. In addition to the main frequency dial, an incremental scale is also provided, allowing an exact frequency selection in the range of plus or minus 50 Hz around any given setting of the main frequency dial. This is accomplished by the use of a coaxial knob on the main dial. A worm-gear attachment, by means of a flexible shaft at the side of the cabinet, permits the oscillator to be connected mechanically to a graph recorder or other mechanically driven equipment.

For acoustical measurements, the instrument contains frequency modulation circuits (variable tone), employing a reactance tube controlled by a saw-tooth oscillator which is switched into the fixed oscillator circuit. Both the amplitude and frequency of the saw-tooth oscillator are adjustable from 0 to 250 Hz, plus or minus the selected frequency. By means of a compressor circuit which may be controlled from an external source of voltage, it is possible to keep voltage, current, or sound pressure constant during acoustical measurements, when the oscillator is used as the sound source. (See Question 22.52.)

The fixed oscillator operates at a frequency of 120 kHz, and the variable oscillator operates from 120 Hz to 100 kHz. The plates in the variable oscillator are so shaped to result in a logarithmic frequency response. In addition, several other controls are provided, including a variable attenuator, impedance-matching transformer with outputs of 6, 60, 600 and 6000 ohms, and a vacuum-tube voltmeter. The attenuator system is adjustable in steps of 10 dB, operated in conjunction with a continuously variable control. The THD varies from 0.1 to 2 percent with frequency and the type of loading and output circuit. Hum and noise are 70 dB below the maximum power output of 2.5 watts.

A beat-frequency oscillator, Model 1305, manufactured by the General Radio Co., is shown in Fig. 22-51C. Referring to the elementary diagram in Fig. 22-51D, at the left are two radio-frequency oscillators, one fixed and one variable controlled by the large dial on the front panel. The fixed oscillator tubes V1A and V1B, deliver a signal of approximately either 190 or 210 kHz to the mixer tube, V3, where the signal is combined with the signal from variable oscillator, V2A and V2B. This latter oscillator is variable from 170 to 190 kHz. The difference signal from the two oscillators is fed to an amplitude control through a low-pass filter which removes harmonics above 40 kHz. Leaving tubes V5A and V5B, the signal is amplified and transformer coupled to the output circuit.

The coils in the oscillator are constructed on a ceramic form for stability. The oscillator tubes are twin triodes, with the plate of the oscillator section

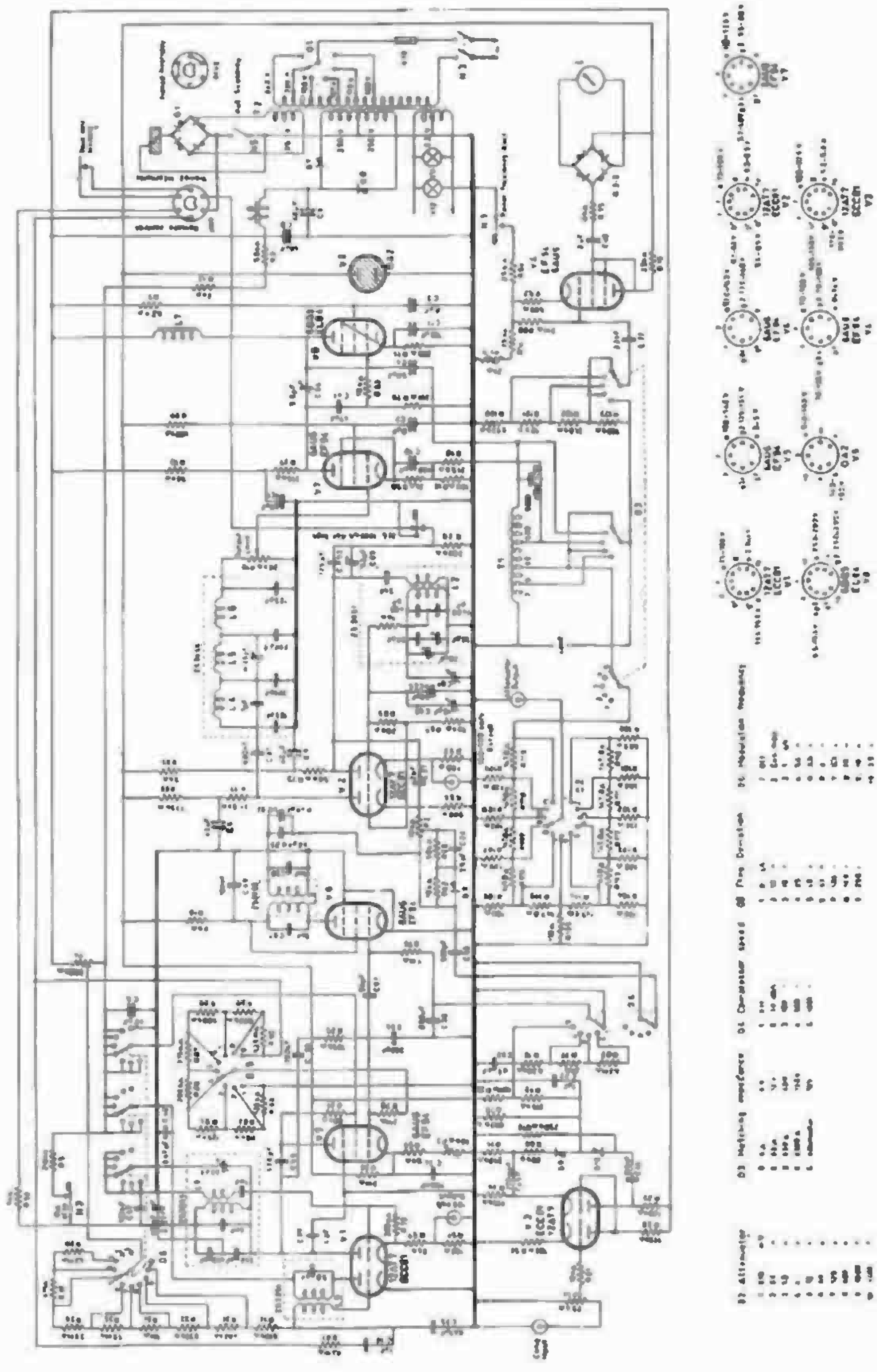


Fig. 22-51B. Schematic diagram for circuits of Brüel and Kjaer Model 1022 beat-frequency oscillator.

grounded and the other section used as a cathode follower isolating amplifier. The mixer tube operates with bias adjustments on grids 1 and 3. This permits the adjustment of voltage and harmonic content of the output signal. This tube is magnetically shielded to prevent pickup from extraneous magnetic fields.

The output stage employs a single-ended push-pull amplifier configuration. One section of the 12AT7 twin

Fig. 22-51C. General Radio Co., Model 1304B beat-frequency oscillator.

triode drives the other section, which is connected as a phase-inverter to drive the output stage in push-pull. The amplifier output is connected through a voltage divider network to the cathode of V5A to introduce negative feedback. The output contains a voltmeter for adjusting the signal amplitude and an attenuator system. The metering system employs an average-reading voltmeter in conjunction with two 1N54 diode rectifiers. The attenuator system consists of T pads, which are switched to provide attenuation in steps of 20 dB, from 0 to 60 dB, thus providing a resistive internal output impedance of 600 ohms.

The instrument may be calibrated to the power-line frequency or by using the zero-beat method. If the power-line frequency is controlled, standardization at the power-line frequency is recommended. The frequency drift for the first hour of operation is less than 7 Hz. For power-line calibration, the large dial is set to line-frequency calibrate, and the zero-adjust control is set for a zero-beat indication on the output meter. For zero-beat calibration, the main dial and the cycles increment dials are both set for zero. The zero-adjust control is then adjusted for a zero-beat indication on the output meter.

The advantage offered by a beat-frequency oscillator over the Wien bridge or bridged T is that it may be swept over the full audio spectrum in one full swing of the main dial. This is very convenient when the frequency characteristics of equalizers, filters, and similar equipment are measured. Such oscillators may also be used in conjunction with graph recorders, by connecting the two instruments together mechanically. A General Radio Co. beat-frequency oscillator and a graph level recorder are pictured in Fig. 22-51E. This oscillator has a THD of 0.25 percent from 100 to 10,000 Hz. Below 100 Hz, the harmonic distortion may reach 0.5 percent. Output power is 1 watt into a 600-ohm resistive load.

22.52 ***Describe a warble oscillator.***—Warble oscillators are used for making acoustic measurements in an enclosure. The frequency of the oscillator is swept slowly from the lowest to the highest frequency while being warbled at the rate of four times per second over a range of 10 percent plus and minus the mean frequency.

Warbling the oscillator frequencies when making acoustic measurements prevents the formation of standing waves within an enclosure. Although it is not necessary to warble frequencies above 1000 Hz, it is the practice to warble all frequencies when making a measurement.

Warble frequency oscillators can be mechanically driven, or the warble frequency can be created electronically. In the mechanical design a motor-driven beat-frequency oscillator, with a four-sided cam mounted on the main dial, produces the warble while the oscillator dial is swept across the frequency spectrum. In the electronic design the warble is produced by electronically modulating the oscillator as the dial is swept across the spectrum. (See Question 22.51.)

22.53 ***What is an audio-frequency sweep generator?***—An audio oscillator which automatically sweeps across the audio-frequency band approximately twenty times per second.

The generator consists of a rotating disc on which is photographed a sine wave starting at a low frequency and

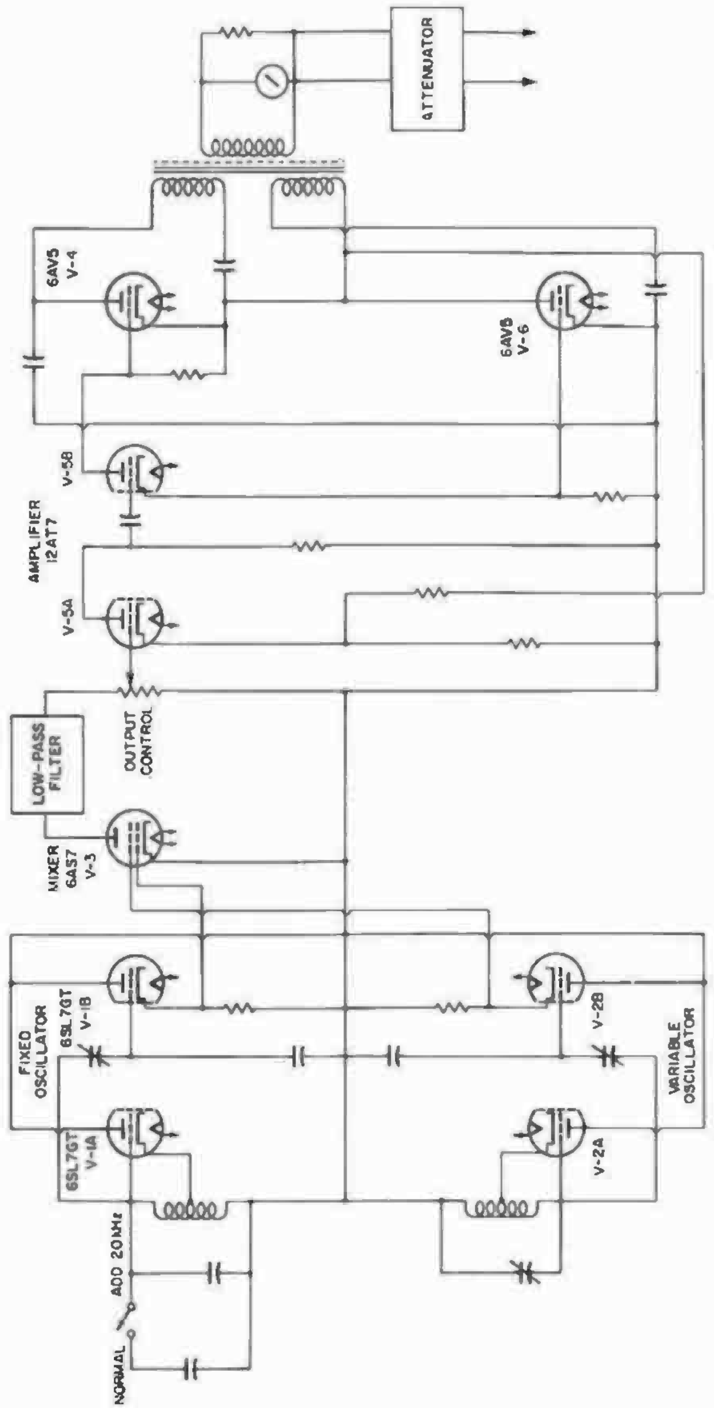


Fig. 22-51D. Elementary diagram for General Radio Co., Model 1304B beat-frequency oscillator.

progressing up to a high frequency. The disc is rotated by a synchronous motor and the image of a sine wave is projected on the target of a photo-electric cell, amplified, and applied to the input of the device being tested. The frequency response is observed visually by means of an oscilloscope connected at the output of the device under test. The same function can be

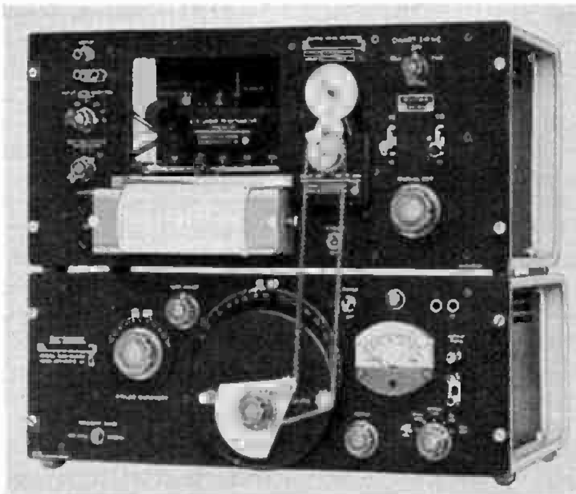


Fig. 22-51E. General Radio Co. graphic level recorder and beat-frequency oscillator.

accomplished electronically by using an audio oscillator and a reactance control tube.

22.54 ***Describe a square-wave generator?***—Square-wave generators may be obtained in two forms—one in which a separate sine-wave oscillator is applied to the input of a device containing clipping and squaring tubes and one in which a sine wave is generated internally and squared. The resulting waveform is a square wave.

The schematic diagram for a transistor square-wave generator to be used

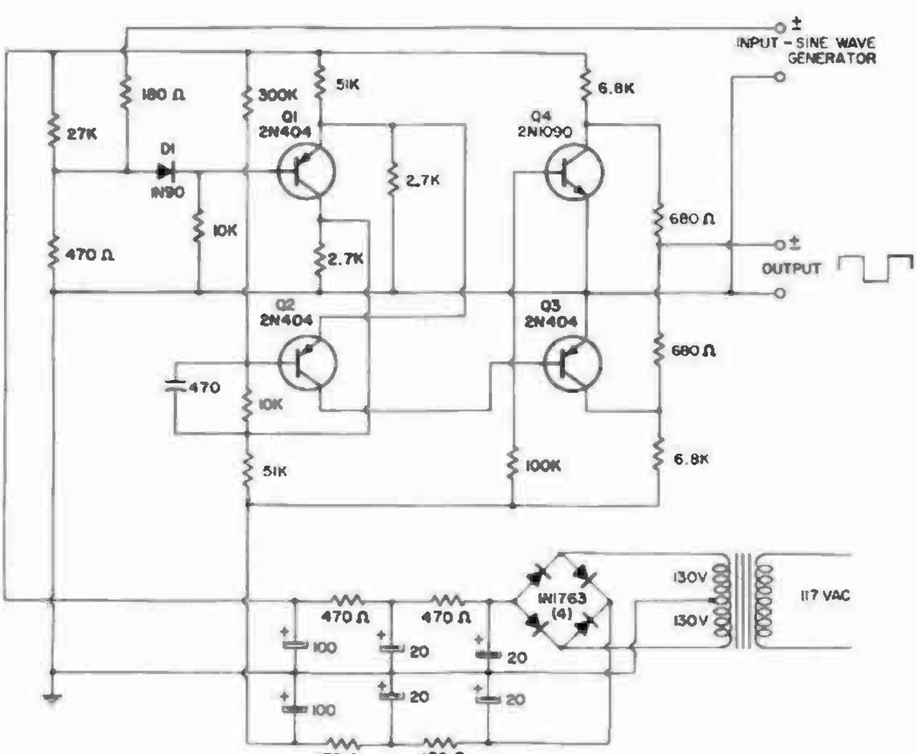


Fig. 22-54. Schematic diagram for transistor square-wave generator. External sine-wave generator is required. (Courtesy, Waveforms Inc.)

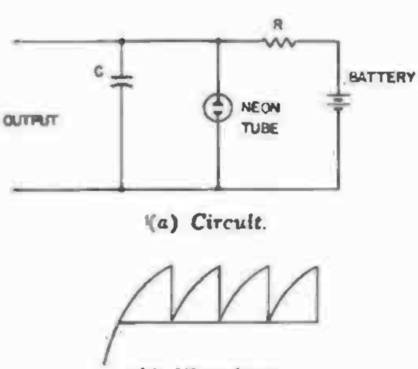
with an external audio oscillator is given in Fig. 22-54. Basically the circuit is a modified bistable multivibrator consisting of two regenerative direct-coupled stages and a switching circuit. It is designed to be connected at the output of a sine-wave oscillator, with the output of the square-wave generator going to the device under test.

The circuit functions as follows. The sine wave is applied to the base of transistor Q1 through a coupling capacitor and steering diode D1. The first stage of Q1 is collector loaded and dc coupled to the base of transistor Q2. Regenerative coupling from Q2 is through the emitter circuit. The output signal is developed by the collector of Q2 and passed through switching transistors Q3 and Q4, to a voltage divider network with a floating ground.

The power supply is designed to provide both negative and positive voltages and is of conventional design. If an attenuator network is used at the output of the generator, it must be capable of passing a wide band of frequencies since a square wave is theoretically of infinite bandwidth.

22.55 ***What is a neon oscillator?***—An oscillator consisting of a small neon light, a resistance, and a capacitor connected as shown in Fig. 22-55. The capacitor is charged and discharged by the neon light.

The values of the resistance R and capacitance C determine the rate of discharge and the frequency of oscillation. The waveform is of the sawtooth variety and rich in harmonics. The characteristics of neon tubes are given in Questions 25.99 and 25.100.



(a) Circuit.

(b) Waveform.

Fig. 22-55. Neon-tube oscillator. Frequency of oscillation is controlled by the value of R and C.

22.56 ***Describe a random- or white-noise generator.***—A random-noise generator is a device that produces a large amplitude of electrical noise at its output. This type of noise is useful in making the following tests: acoustical, speaker, microphone, psychoacoustic, filter, crosstalk, modulation of signal generators, and the comparison of effective bandwidths. Such a generator, Model 1390-B, manufactured by the General Radio Co., is shown in Fig. 22-56A.

Referring to the simplified block diagram of Fig. 22-56B, the generator portion consists of half of a 6D4 gas discharge tube placed in a transverse magnetic field which is supplied by a permanent magnet. The magnet eliminates oscillations usually associated with such tubes and increases the noise level at the higher frequencies. The noise output from the gas discharge tube is amplified by a two-stage amplifier, V3 and V4. The noise spectrum is shaped in one of three different ways between the first and second stage, depending on the setting of the range switch.

In the 20-kHz position of the range switch, a low-pass filter with a gradual rolloff above 30 kHz is inserted. The audio range is uniform to 20 kHz. In the 500-kHz position a low-pass filter with a rolloff above 500 kHz is used. In the 5-MHz position a peaking network compensates for the drop in noise output from the 6D4 tube at the high frequencies, and thus a reasonably good spectrum to 5 MHz is obtained. Leaving tube V4, the signal passes through an output control to an attenuator system. A rectifier-type averaging meter indicates the output voltage.

The maximum open-circuit output voltage in the 20-kHz band is about 3

Fig. 22-56A. General Radio Co. Model 1390-B random-noise generator.

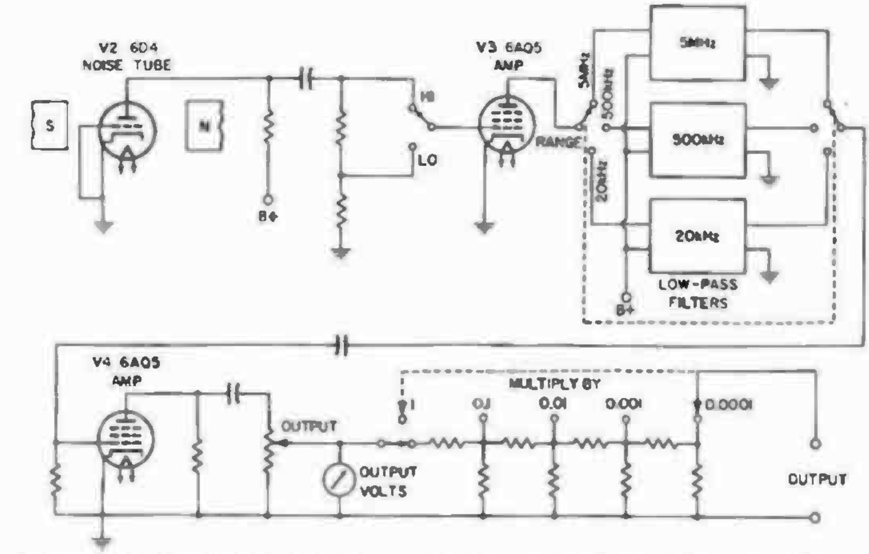


Fig. 22-56B. Simplified block diagram for General Radio Co., Model 1390-B random-noise generator.

volts; in the 500-kHz band, it is 2 volts; for the 5-MHz band, it is 1 volt. These voltages represent relatively high noise levels, since the output impedance of the generator is 900 ohms. The noise level may be expressed in terms of the resistance noise corresponding to 900 ohms at room temperature. The rms voltage in a one-cycle band, due to thermal agitation in a 900-ohm resistor at room temperature, is 3.8×10^{-9} volt. The level from this generator is about 5 millivolts for a one-cycle band when there is a total output voltage of 1 volt in the 20-kHz band. This level is then about 1,300,000 times the corresponding voltage for resistance noise, or about 122 dB above resistance noise at the same impedance level.

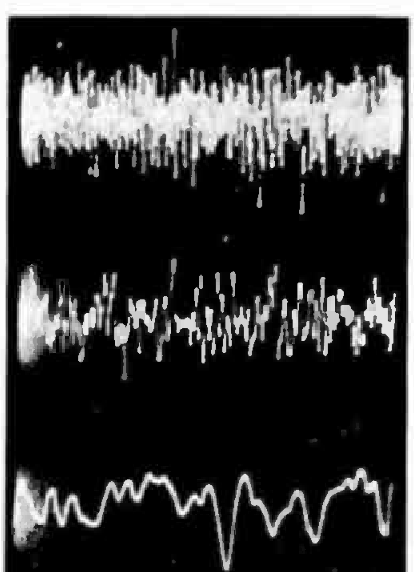


Fig. 22-56C. Oscillographs of three different samples of output voltage taken from the General Radio Co. Model 1390-B random-noise generator.

In random noise no regular pattern appears in the output waveform; it is characterized by randomness rather than by regularity. Noise is therefore described by statistical means, and it is characterized by its distribution of instantaneous amplitudes and by its frequency spectrum. Random noise is often defined as noise that has a normal or Gaussian distribution of amplitudes. White noise is discussed in Questions 1.140 and 1.141.

Oscillographs of three different samples of the output voltage, taken from the output of a noise generator are pictured in Fig. 22-56C. A typical spectrum-level characteristic for the described random-noise generator is shown in Fig. 22-56D.

Fig. 22-56E is a schematic diagram for a random-noise generator, using a Solitron Sounvistor. The heart of the generator is the deep, double-diffused silicon diode. These special diodes become noise generators by applying a suitable dc reverse voltage across a

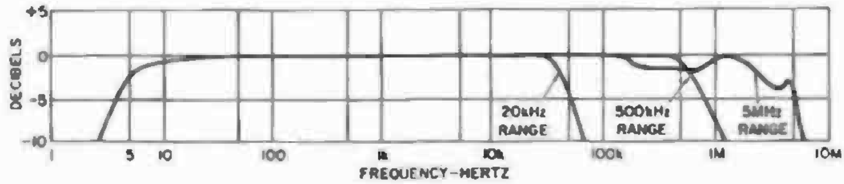


Fig. 22-56D. Frequency characteristics of General Radio Co. Model 1390-B random-noise generator, for its three bands of operation.

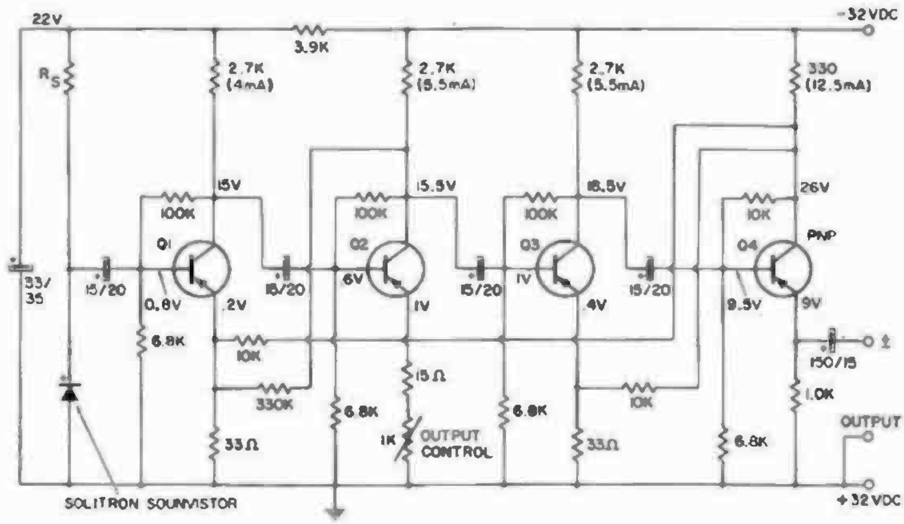


Fig. 22-56E. Random-noise generator using a Solitron Sounvistor. The noise-producing diode may be one of several different types. (Courtesy, Solitron Devices Inc.)

combination of a series feed-resistor in series with the Sounvistor. The random noise voltages are recovered from the diode by means of a conventional transistor amplifier. The noise is amplified to the desired level and bandwidth.

Two basic diodes which differ in noise frequency and also in fundamental operating characteristics are used. The most useful type is the type SD-1, a high-impedance generator that will generate frequencies from 1 Hz to about 100,000 Hz. This diode will develop a 200,000-Hz bandwidth of 500 to 3000 microvolts rms across a load resistance of 100,000 ohms or more. Each diode is studied at an ambient temperature of 25 degrees centigrade to determine its distribution of noise voltage and is then classified for a given service.

When listening to white noise over a wideband speaker system of fairly uniform frequency characteristics, the sound should be one of an unmusical tone swish, similar to that heard when tuning between fm channels. (See Question 7.114.)

22.57 What is a phase-shift oscillator?—An oscillator employing a single vacuum tube with a phase-shifting network rather than the conventional LC circuit. A typical phase-shift oscillator is shown in Fig. 22-57A.

Basically the circuit consists of a tube and a three-stage, phase-shifting network consisting of three capacitors, C1, C2, and C3, and three resistors, R1, R2, and R3, connected between the plate and grid circuits of the oscillator tube. The phase shift in the network is proportional to the current through the network. The circuit element values shown are for an oscillator of 1000 Hz. Each phase-shifting network has a phase shift of 60 degrees. The total phase shift for the three networks is 180 degrees. The circuit functioning may be explained as follows.

When an alternating voltage is applied to the first section of the network, C1 and R1, a current is caused, the magnitude being determined by the total impedance of the network. Since the impedance is capacitive, the current

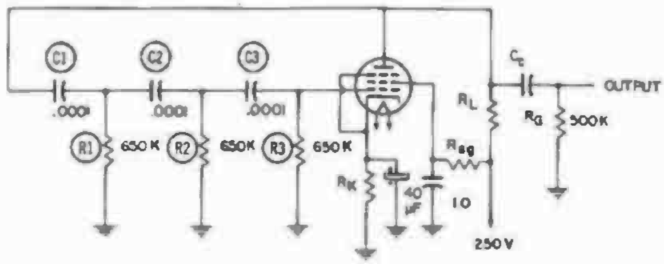


Fig. 22-57A. Phase-shift oscillator circuit.

will lead the voltage, which for one section of this network (C1 and R1) is 60 degrees. The voltage drop across R1 is in phase with the current through it; therefore, the voltage drop across R1 leads the impressed voltage by 60 degrees.

The output voltage of the first section is applied to a second section, C2 and R2, which shifts the impressed voltage another 60 degrees for a total of 120 degrees. The third network, C3 and R3, shifts the voltage another 60 degrees, making a total phase shift of 180 degrees. This brings the signal from the plate into phase with the control grid, which is 180 degrees out of phase with the plate.

Oscillation is started when a small disturbance occurs in the plate circuit, such as turning on the power and the starting of plate current. Any slight change from a static condition causes the disturbance to be amplified, inverted 180 degrees by the phase-shifting network, and applied to the control grid and again amplified. This build-up continues until a state of steady oscillation is reached.

The output waveform of a phase-shift oscillator is almost sinusoidal, if the bias is adjusted to a value where oscillation can just be sustained. The tube then operates on the linear portion of its characteristic. Decreasing the resistance or capacitance in the network will increase the frequency of oscillation and vice versa. The angle of phase shift is dependent on the ratio of the capacitive reactance to the resistance in any one section of the network:

$$\phi = \frac{X_c}{R} = \frac{1}{2\pi fC/R}$$

where,

X_c is the capacitive reactance,
R is the resistance for a single section of the phase-shifting network.

From a practical standpoint, the tube should have a transconductance above 5000 micromhos, preferably between 800 and 10,000 micromhos, to overcome the loss of the network and sustain oscillation. It has been shown by Ginzton and Hollingsworth that a circuit gain of 29 or more is required to maintain oscillation. The value of R must be

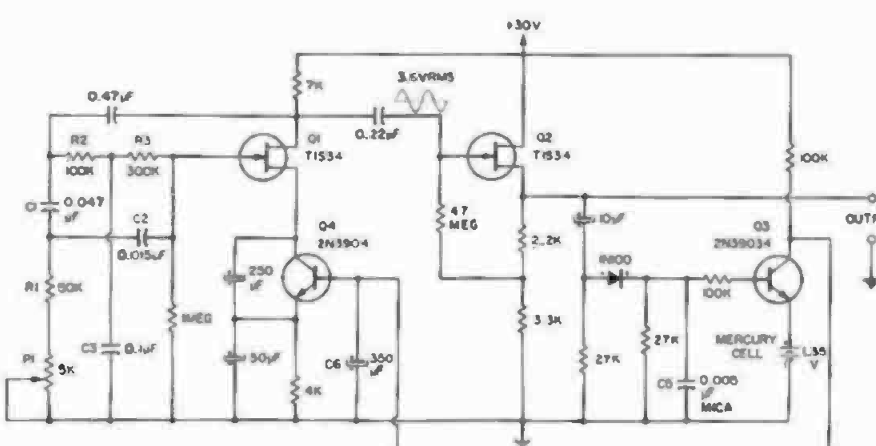


Fig. 22-57B. Transistor phase-shift oscillator with constant-amplitude control circuit.

several times the value of plate load R_L in parallel with grid resistor R_g and coupling capacitor C_c. A second stage is essential to isolate the oscillator section from the external circuit.

The frequency of oscillation for a given set of values may be approximated:

$$f = \frac{1}{2\pi RC\sqrt{6}}$$

where,

R and C are the values of a single network.

The frequency of oscillation may be adjusted to an exact value over a very limited range by varying the value of any one resistor in the network. A variable-frequency oscillator may be constructed by varying the values of C and R. This may be done by ganging and adjusting their values simultaneously. Typical values for several frequencies, using a plate load resistance (R_L) of 50,000 ohms, are given below:

Freq. (Hz)	*R1, R2, R3*	*C1, C2, C3*	
50	1.25 meg	0.001	μF
100	1.00 meg	0.0006	μF
500	510K	0.00025	μF
1000	650K	0.0001	μF
5000	100K	0.00011	μF
10,000	510K	0.0000125	μF

The actual phase angle for a single section using the values given in the foregoing will vary between 66 and 77 degrees, which is satisfactory. Thus standard values of resistance and capacitance may be used.

Fig. 22-57B shows the circuitry for a transistor phase-shift oscillator employing field-effect transistors (FET's) in combination with conventional transistors. This circuit, developed by Mourlam, also includes a stabilizing feedback loop. This circuit is particularly adaptable to frequencies below 100 Hz. The required 180-degree phase shift is obtained by the use of an unbalanced T network, consisting of resistors R1, R2, and R3, capacitors C1, C2, and C3, and potentiometer P1, which provides a small amount of frequency adjustment.

The output from transistor Q1 is fed to buffer FET Q2, rectified, and used to charge capacitor C5 proportional to the amplitude of the oscillation as compared to a reference voltage supplied by a 1.35-volt mercury cell in the emitter of Q3. The resulting amplitude variations are then applied to the base of Q4, which in turn varies the bias of Q1, thus compensating for the variations in amplitude.

It is claimed for this oscillator circuit that the stability is to within a few parts in 10^4. The circuit shown has an operating frequency of 33 Hz.

22.58 ***What is a resistance-stabilized oscillator?***—Basically the oscillator may be a Hartley or a tuned-plate type circuit except for a feedback resistance and capacitance connected between the plate and the tuned circuit. The resistance and capacitance are shown as C_{fb} and R_{fb} in the schematic diagram of Fig. 22-58.

The tube operates as a class-A amplifier stage with a self-biasing resistor in the cathode circuit. This resistor is adjusted for the lowest possible harmonic distortion. The value of the feedback resistor R_{fb} is selected for a value that will just permit the tube to oscillate, with stability.

Resistance-stabilized oscillators are often used where a high degree of stability and low distortion are required. Capacitor C_t across the tuned circuit is changed for each desired frequency. The feedback resistor, R_{fb}, is readjusted for each frequency, for the lowest distortion consistent with stable operation.

22.59 ***Describe a multivibrator oscillator.***—There are two types of multivibrators: the astable and the bistable. The first to be discussed will be a transistor version of the astable type. The astable multivibrator (or *free-running multivibrator* as it is sometimes called) develops a square-wave output that has a peak value equal to the dc voltage (V_{cc}) and a minimum value equal to the collector saturation voltage of the transistor. The circuit is basically a

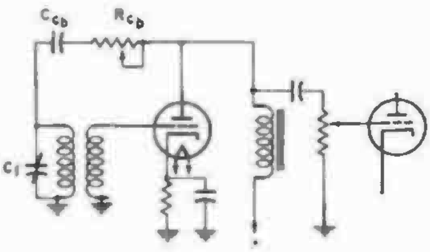


Fig. 22-58. Resistance-stabilized oscillator. Value of capacitor C_t is set for each frequency and resistor R_{fb} is set for lowest distortion.

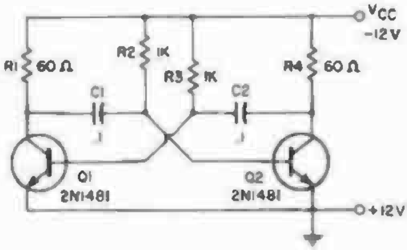


Fig. 22-59A. Astable multivibrator oscillator circuit.

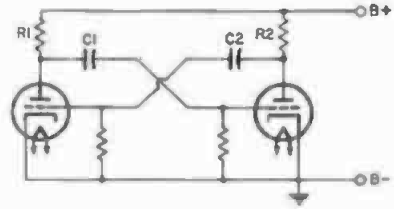


Fig. 22-59C. Astable multivibrator circuit using vacuum tubes.

two-stage nonsinusoidal oscillator in which one stage conducts at saturation, while the other is cut off until a point is reached at which the stages reverse their conditions. The circuit shown in Fig. 22-59A employs two 2N1481 transistors, operated in identical common-emitter amplifier stages, with regenerative feedback resistance-capacitance coupled from the collector of each transistor to the base of the other. The frequency of oscillation for the circuit shown is controlled by the value of the resistors and the capacitors and may be calculated:

$$f = \frac{1}{(0.7C1R2)\ (0.7C2R2)}$$

The frequency of the circuit shown is approximately 7000 Hz.

A bistable multivibrator, commonly called a *flip-flop oscillator*, is shown in Fig. 22-59B. This circuit finds wide usage in computer applications and electronic switches. The circuit is in a stable state when either transistor is conducting and the other is cut off. The state of the transistors is switched by the application of a properly applied trigger pulse. The 1N126 steering diodes ensure that the 2N404 pnp transistors are triggered to alternate states only when positive pulses are applied to the input.

The output voltage, which may be taken between collector and ground of either transistor (or both), is a unit step voltage, changing when the trigger is applied. A square-wave output is obtained by a continuous periodic pulsing of the input. For the circuit shown, the frequency division between input and output is 2:1.

The circuit given in Fig. 22-59C is an astable or free-running multivibrator employing vacuum tubes. Although the astable multivibrator is free-running, it may be triggered by pulses of a

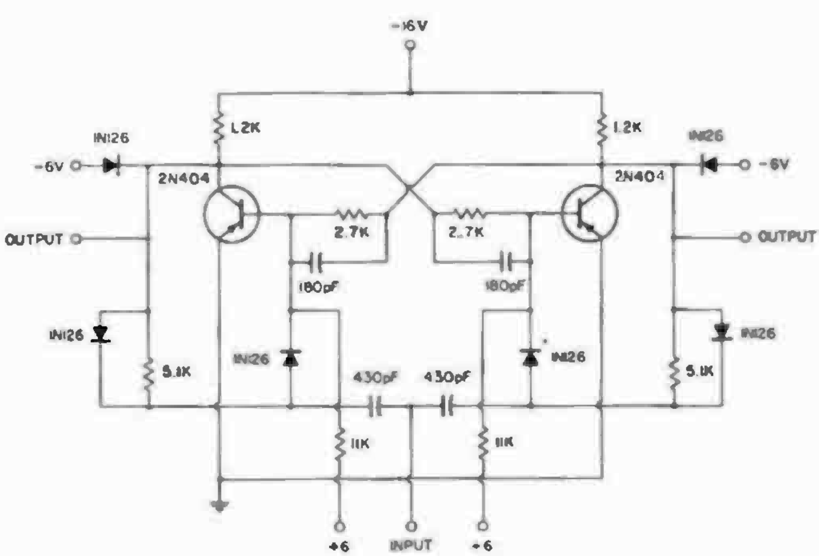


Fig. 22-59B. Bistable multivibrator or flip-flop oscillator circuit.

given amplitude and frequency to provide a frequency-stabilized output.

22.60 What is an LC-filter type of distortion analyzer?—An LC-filter type of distortion analyzer is an instrument employing a group of LC type band-elimination filters, designed for measuring the total harmonic distortion (THD) of amplifiers. This type of instrument does not contain any electronic equipment internally, but requires an external null detector, such as a vacuum-tube voltmeter or oscilloscope.

The instrument shown employs five band-elimination filters with fundamental frequencies of 50 Hz, 100 Hz, 400 Hz, 1 kHz, 5 kHz, and 7.5 kHz, which are selected by means of a switch at the lower left of the panel. A block diagram of this instrument appears in Fig. 22-60A.

Fundamentally the device consists of two variable attenuators, one having 60 dB of loss in steps of 10 dB, the other a total of 10 dB loss in steps of 1 dB, making a total of 70 dB of loss variable in steps of 1 dB. Following the attenuators is a group of six band-elimination filters, any one of which may be selected by switch S2. Connected externally to the output of the distortion analyzer is a vacuum-tube voltmeter or oscilloscope for indicating the amplitude of the fundamental frequency or the harmonics of the fundamental.

To operate the instrument, an oscillator of low harmonic distortion is connected to the input of the device to be tested. The distortion analyzer is connected to the output of the device under test and a sensitive indicating instrument is connected to the output terminals of the distortion analyzer. Switch S2 is set to the desired filter and switch S1 to position A, which connects the selected filter into the circuit. The oscillator is set to the filter frequency and slowly rocked above and below the filter frequency while balance control P1 on the analyzer is adjusted for a minimum deflection of the indicating instrument. At the point of minimum deflection or null, the fundamental frequency has been completely balanced out, leaving only the harmonics. The balance control (P1) varies the Q of the filter circuit, permitting an exact balance.

When a minimum deflection has been attained, switch S1 is thrown to position B, substituting the two variable attenuators for the filter network. The loss of the attenuators is adjusted for the same amplitude reading on the indicating meter as previously obtained

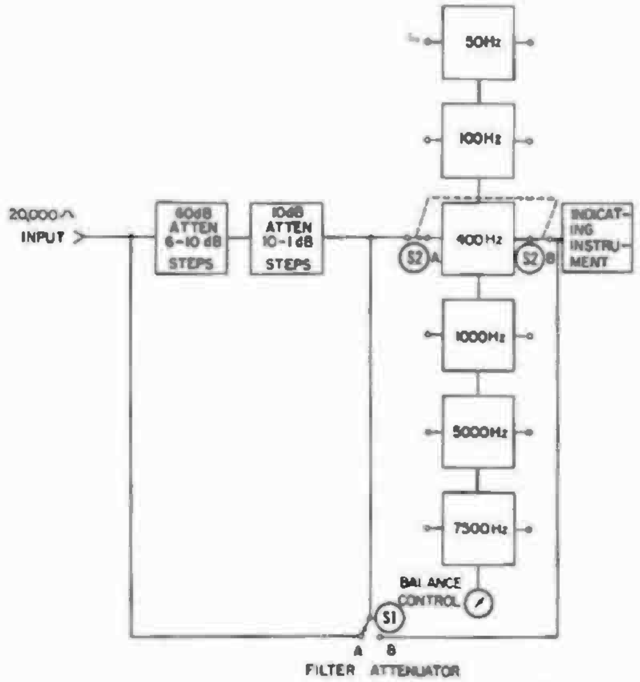


Fig. 22-60A. Block diagram of LC-filter type distortion analyzer.

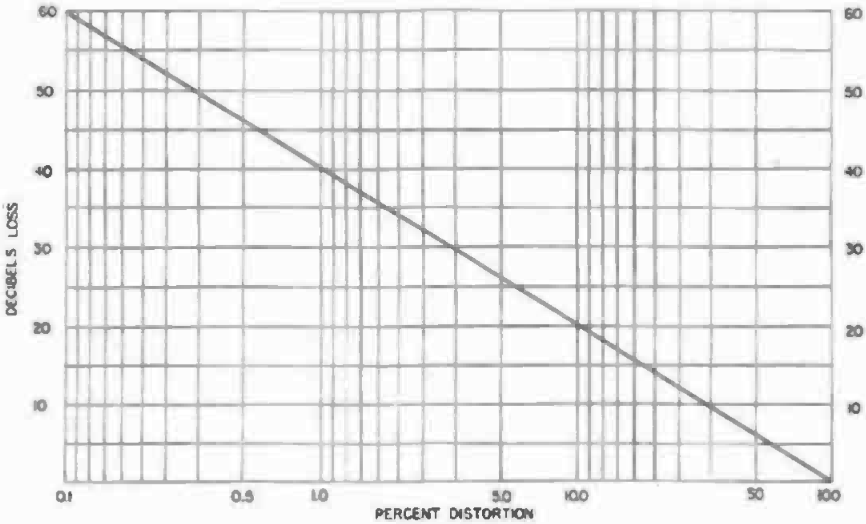


Fig. 22-60B. Attenuator loss versus percent harmonic distortion (THD).

with the filter in the circuit. When the two indications are the same, switch S1 may be thrown from position A to position B with only a fractional change in level on the meter (this is generally less than 1 dB).

When a condition of balance has been attained, the loss of the attenuators is totaled and the percent harmonic distortion read from the graph given in Fig. 22-60B. As an example, if the loss of the attenuators totals 40 dB, the harmonic distortion is 1.0 percent. The attenuators loss is also an indication that the harmonics are 40 dB below the amplitude of the fundamental frequency. It will be noted from the graph that when the harmonic distortion is 10 percent, the harmonics are only 20 dB below the fundamental, and for 0.10 percent they are 60 dB below the fundamental frequency. The schematic diagram of a single filter section is shown in Fig. 22-60C. The procedure for use of the foregoing instrument is similar to that of other distortion meters and analyzers described in Section 23.

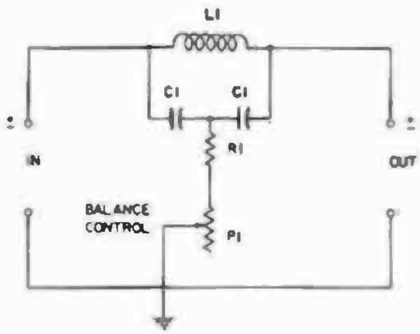


Fig. 22-60C. Schematic diagram of a single section of a multiple LC filter-type distortion analyzer.

22.61 Describe a distortion meter using a single 400-Hz high-pass filter.—In the early design of distortion-factor meters (distortion analyzer) the harmonic distortion was measured only at 400 Hz and not over a wide band of frequencies as is today's practice. These early analyzers employed a single 400-Hz high-pass filter with a steep cutoff characteristic, as shown in Fig. 22-61A. For the null detector, an amplifier and thermocouple current-squared meter were used. The schematic diagram for such a meter is shown in Fig. 22-61B. With the filter in the circuit, the fundamental frequency of 400 Hz is suppressed about 75 dB, and all harmonics up to the fifteenth are passed without attenuation. For historical interest, a picture of such an instrument, manufactured by the General Radio Co. in 1930, is shown in Fig. 22-61C.

Referring to the schematic diagram in Fig. 22-61B, included with the filter are three resistors, R1, R2, and R3, and potentiometer P1 for comparing the voltage at the output of the filter with respect to the voltage at the input of the filter. A dial calibrated 0 to 3 and 0 to 30 percent is mounted on the shaft of the potentiometer P1.

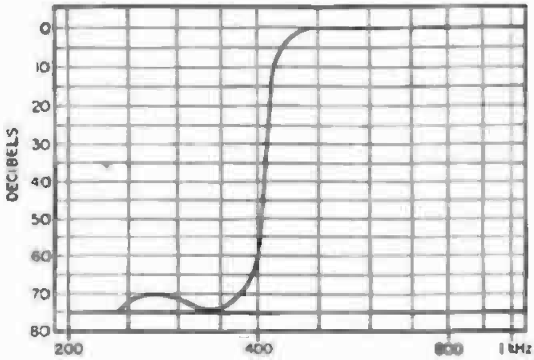


Fig. 22-61A. Frequency characteristic of 400-Hz high-pass filter used in distortion-factor meter.

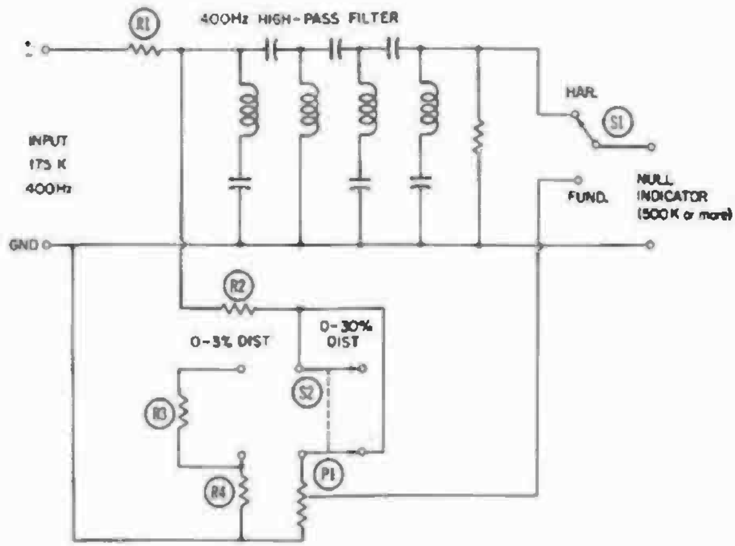


Fig. 22-61B. Schematic diagram for distortion factor meter.

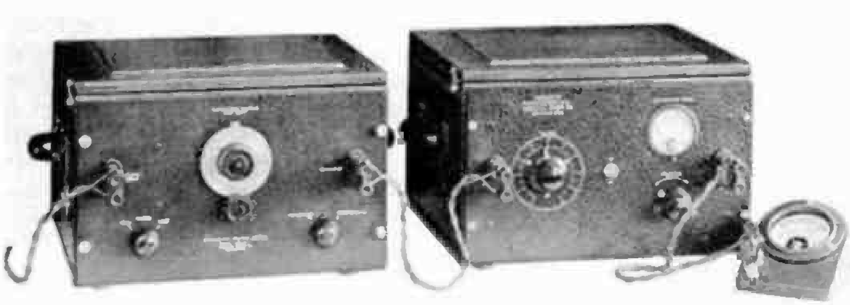


Fig. 22-61C. For historical interest, 400-Hz high-pass filter type distortion factor meter, amplifier, and current-squared thermocouple meter, manufactured by General Radio Co., in 1930.

When a measurement is made, the distortion-factor meter is connected across the load resistance of the amplifier under test. A 400-Hz signal is applied to the amplifier input, and output switch S1 is set to fundamental. S2 is set to 30 percent and a reference reading taken on the indicating meter. S1 is then put in the harmonic position and the potentiometer balanced for a null reading of the meter. S2 is set for the desired range, rebalanced, and the percent distortion factor read from the dial. Series resistances R1 and R2 form an

L-type network ahead of the filter, thus preventing the calibration of the analyzer from being affected by the impedance of the device being tested. Distortion factors as low as 0.2 percent may be read. As this device (and others of similar design) read percent distortion as the total rms voltage of the harmonics, no indication of the individual harmonic amplitudes is given.

In subsequent models of distortion meters, the amplifier and meter were combined into one complete unit. Modern distortion-measuring equipment generally uses a parallel-T resistive network which is continuously variable. (see Question 22.63.)

22.62 *Describe the circuitry of a distortion-factor meter (DFM)*—A distortion-factor meter (DFM) is an instrument designed for measuring the total harmonic distortion (THD) of amplifiers and similar devices. This is accomplished by applying the distorted signal to the input of the instrument, nulling-out the fundamental frequency by a highly selective RC network and then measuring the remaining harmonics in percent of the fundamental frequency, including the harmonics. It will be noted that the measurement is made of the total remaining harmonics; therefore, unless other means are used, the individual harmonic amplitude cannot be determined from the instrument reading. However, the distortion products can be observed with an oscilloscope or their individual amplitudes measured by the use of a wave analyzer, as discussed in Question 22.65. The discussion to follow will apply generally to any type DFM employing an RC null network of the Wien-bridge or parallel-T configuration. The variable

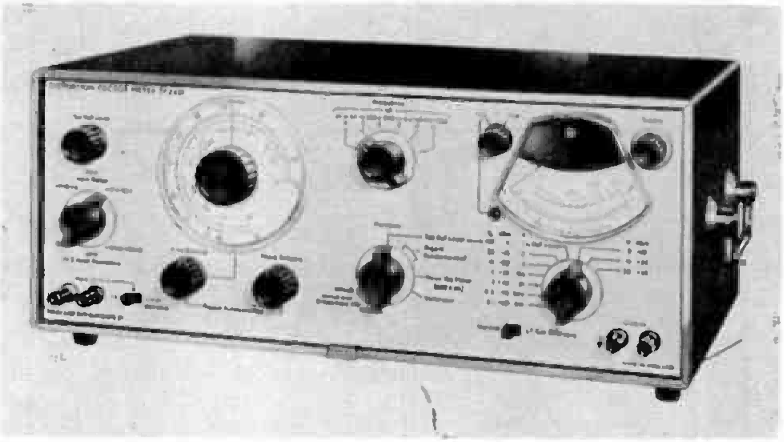


Fig. 22-62A. Marconi Instruments Ltd. (England) Model TF 2331 distortion factor meter. (Courtesy, Canadian Marconi Co.)

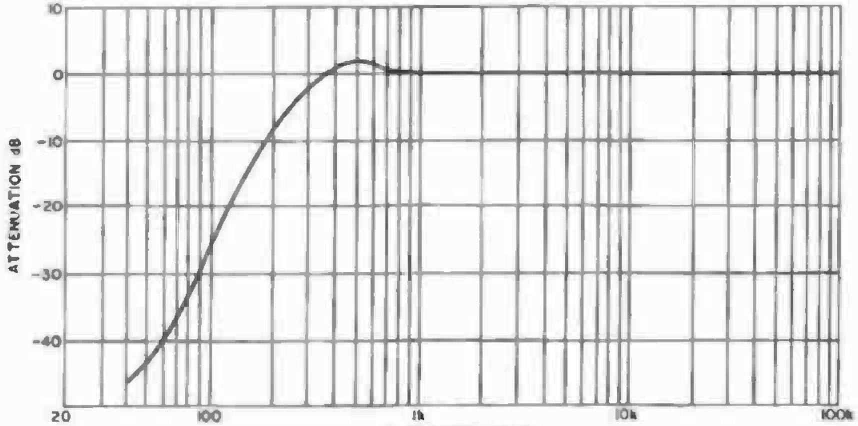


Fig. 22-62B. High-pass filter network for attenuating power-line frequencies.

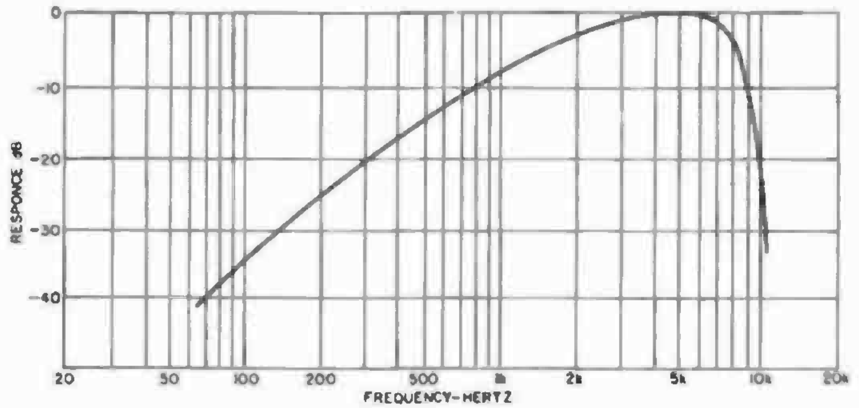


Fig. 22-62C. Weighting network used for noise measurements.

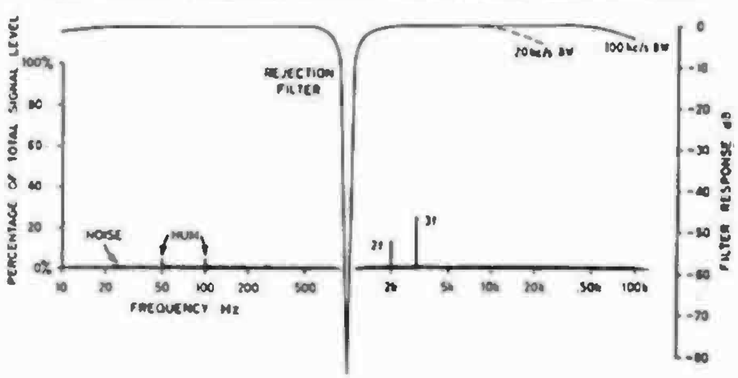


Fig. 22-62D. Normal frequency characteristics of Marconi DFM meter null network.

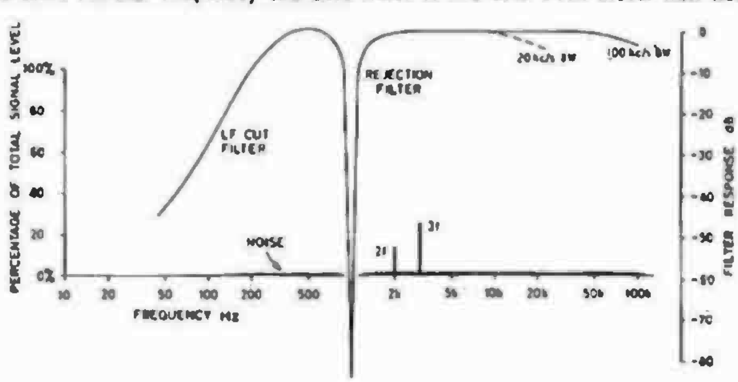


Fig. 22-62E. Frequency characteristics of Marconi DFM meter null network, with high-pass filter for reducing hum frequencies.

components for nulling out the fundamental frequency may be resistive or capacitive.

Fig. 22-62A shows a completely transistorized distortion-factor meter, Model TF 2331, manufactured by Marconi Instruments, Ltd., of England. This instrument measures total harmonic distortion (THD) in the fundamental range from 20 to 20,000 Hz. A distortion bandwidth of either 20 kHz or 100 kHz may be selected and in addition, a high-pass filter having a characteristic as given in Fig. 22-62B, may be switched in to eliminate power-line frequencies from the measurement. Noise can be measured using the same two bandwidths as for distortion measurements, or via a weighting network with a characteristic as shown in Fig. 22-62C.

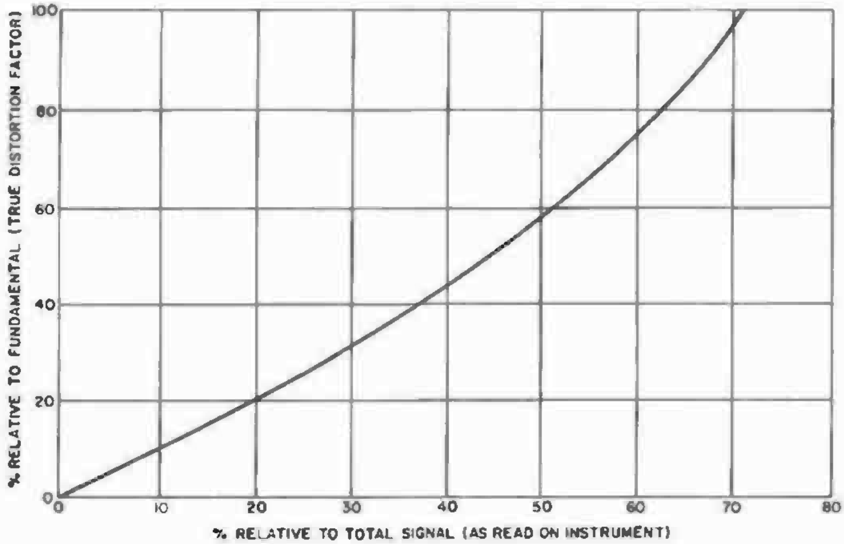


Fig. 22-62F. Conversion graph for indicated distortion and true distortion.

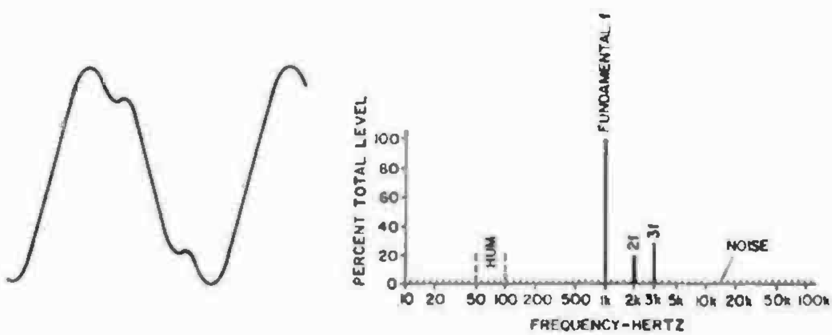


Fig. 22-62G. A 1000-Hz signal with second- and third-harmonic distortion.

This response was originally defined by the CIFF as standardized by the British Standards Institution.

The distortion-measuring rejection network consists of a selective amplifier tuned to reject the fundamental frequency component, and a wideband voltmeter to measure noise and distortion content. Terminals are provided in the metering circuit for connection of an external oscilloscope. The characteristic of the rejection network, tuned to reject 1000 Hz (lowest null point of the meter), is given in Fig. 22-62D. The network has a rejection (when properly tuned) of at least 80 dB, with a second-harmonic attenuation of less than 0.5 dB at 1000 Hz, 1 dB up to 6000 Hz, and 2 dB up to 20,000 Hz. With the high-pass filter (low-frequency cutoff) in the circuit, the rejection characteristic is that shown in Fig. 22-62E. The meter circuit is an average-reading meter and subject to error, as set forth in Question 22.103.

Distortion-factor meters measure distortion relative to the total signal (fundamental plus harmonics) rather than to the fundamental alone. Below 10-percent distortion, the discrepancy between the fundamental frequency alone and the fundamental plus the harmonics is negligible, but it can become quite large above 10-percent distortion. If a conversion from the indicated distortion to the true value of distortion is important, it can be made by the use of the chart in Fig. 22-62F and will hold true for any type of meter of those under discussion. The distortion as read on the instrument is shown at the bottom of the graph, with the correction shown at the left vertical margin. As will be noted, no correction is

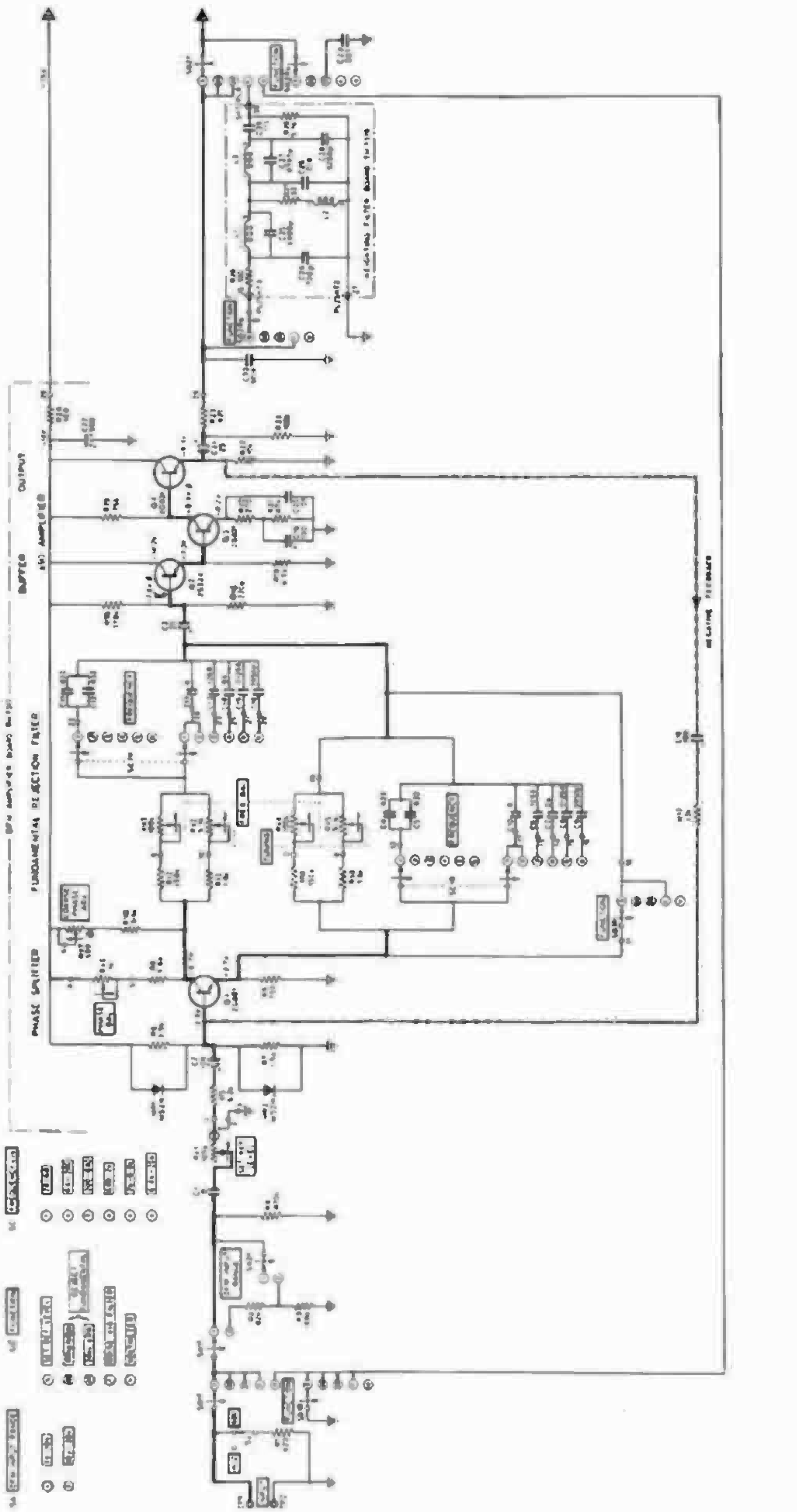


Fig. 22-62H. Distortion measuring section of Marconi Instruments Ltd. Model TF2331 distortion-factor meter.

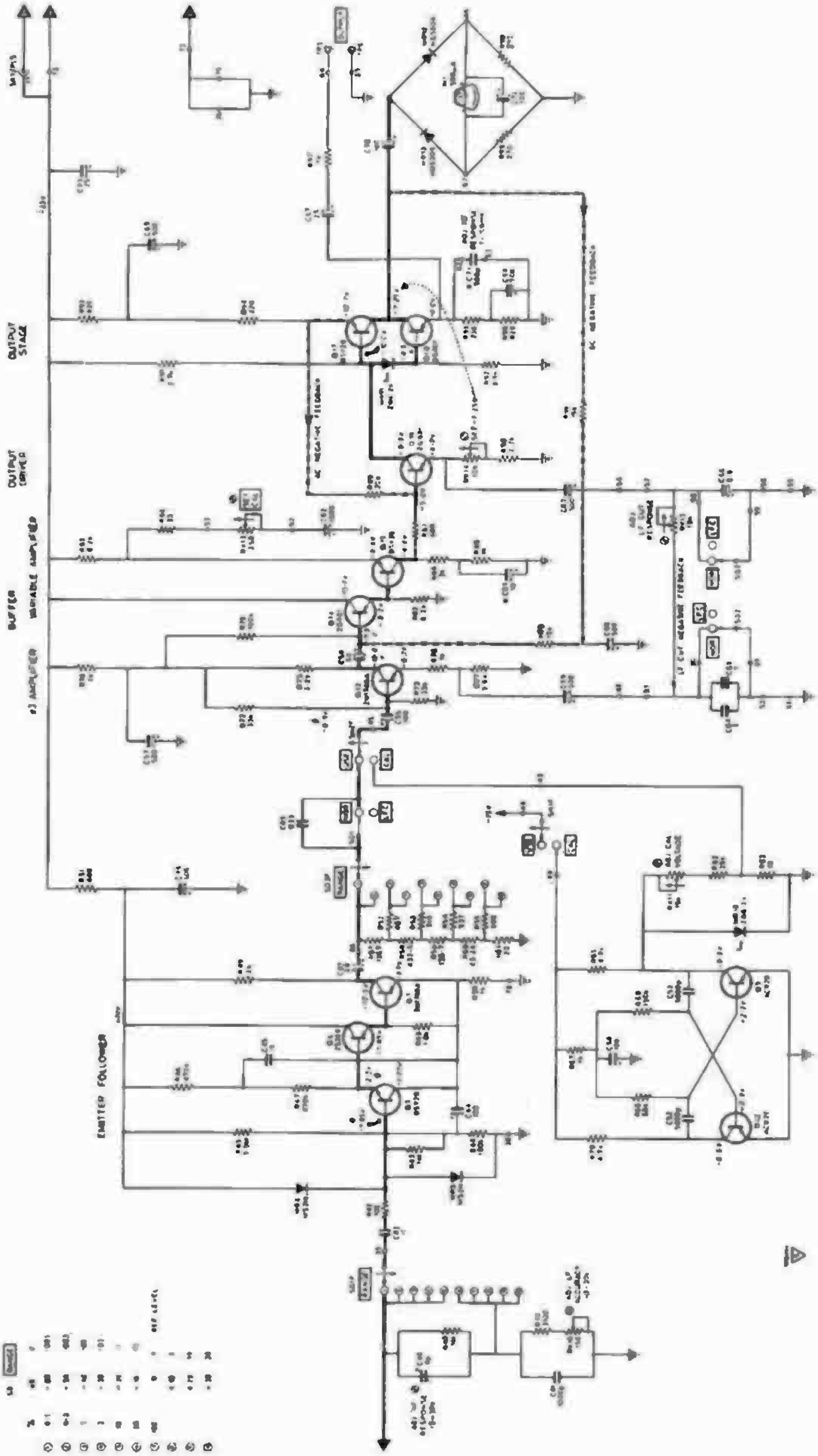


Fig. 22-621. Voltmeter section of Marconi Instruments Ltd. Model TF2331 distortion-factor meter.

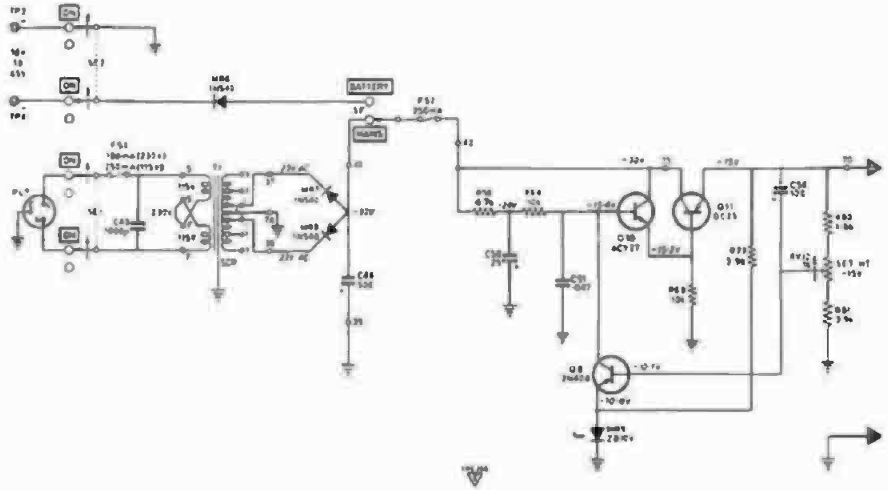


Fig. 22-62J. Power-supply section for Marconi Instruments Ltd. Model TF 2331 distortion-factor meter.

Fig. 22-62K. General Radio Co., Model 1932A distortion and noise meter.

required below 10 percent; however, at 30-percent distortion, the true distortion is 31 percent, and at 50 percent true distortion is 57 percent. It is true that such distortion values are not normally found in conventional equipment. However, in certain types of equipment the distortion may be intentionally high; therefore this factor must be taken into consideration when this type of DFM is used. The spectrum for a 1000-Hz signal having second- and third-harmonic distortion is shown in Fig. 22-62G, including hum and noise frequencies.

Referring to the schematic diagram of Fig. 22-62H, two input impedances are available—a 600-ohm impedance and a high impedance that varies from 6200 to 100,000 ohms, depending on the amplitude of the incoming signal. Because of negative feedback, the impedance at the base of transistor Q1 is quite low. Resistors RV1 and R5 in series with the input signal determine the input level for the 1- to 10-volt range. At maximum sensitivity, corresponding to an input voltage of 0.6 volt, the impedance is at its minimum (6200 ohms) and rises to 100,000 ohms at 10 volts.

When the DFM input range switch is set to the 10- to 30-volt range, attenuators R2 and R3 are introduced. The latter prevent large variations in input sensitivity from affecting the impedance, and it remains at a nominal 100,000 ohms from 10 to 30 volts. Transistor Q1 is a phase splitter and provides two outputs of opposite phase and amplitude (2:1) to drive a Wien-bridge type of network. This network is the fundamental-frequency rejection network. Overall negative feedback taken to the base of Q1 through R17 and C18

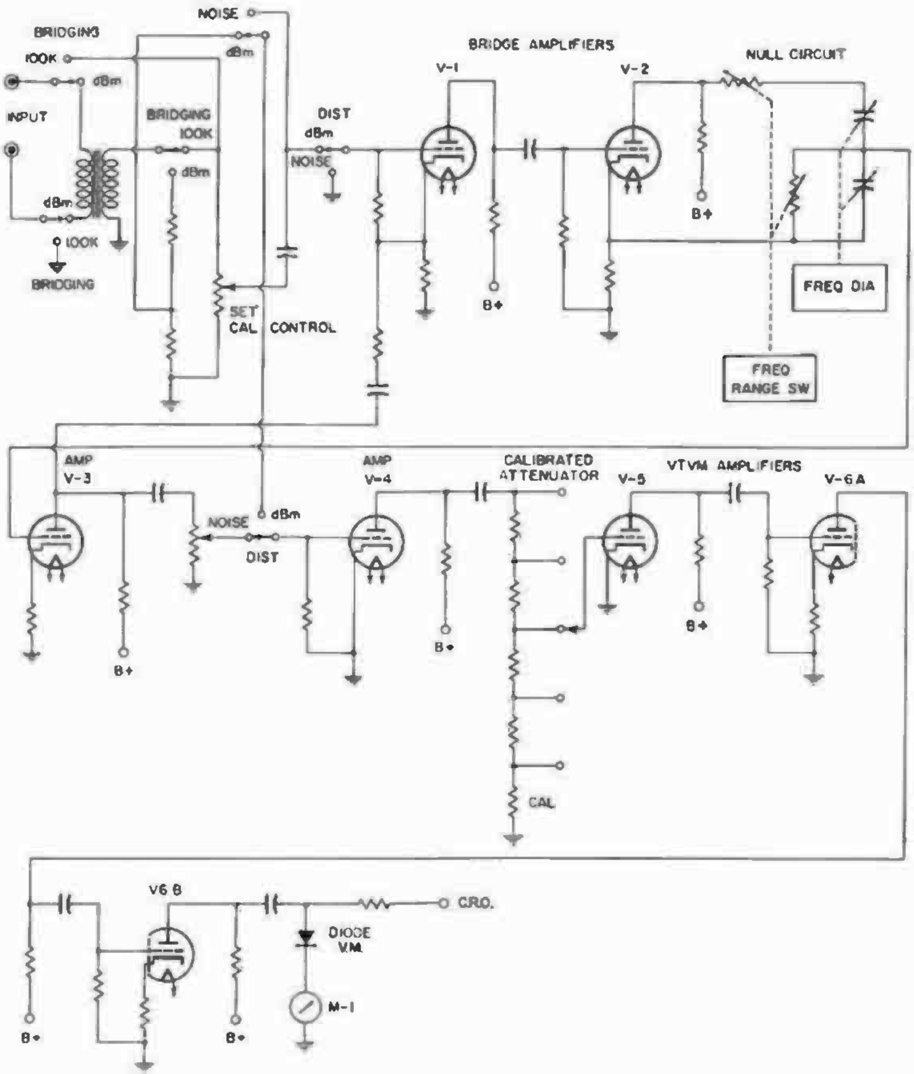


Fig. 22-62L. Elementary diagram for General Radio Co., Model 1932A distortion and noise meter.

from Q4 serves to flatten the skirts of the network response and to prevent a sharper null response to the selective amplifier.

Transistor Q1 is critical with respect to noise and distortion and its operating conditions provide a compromise between the two. The emitter load resistor is 750 ohms, with a collector load adjustable at a nominal value of 1500 ohms, by the phase-balance control. Resistor RV6 in the collector circuit is a 10-turn precision potentiometer adjustment. The collector load feeds the series RC branch of the rejection network, and the emitter load of the parallel RC branch. At least 80-dB rejection is provided by the network. Variable resistors RV2 and RV5 are ganged together to form the Main Tune control, and RV3 and RV4 (also ganged together) form the Frequency Balance control.

Buffer transistor Q2 is a silicon emitter follower, with amplifier Q3 providing the gain. The emitter of Q3 is bypassed by the capacitor C20 to maintain the high-frequency response. Output stage Q4 is also an emitter follower. The 100,000-Hz bandwidth is determined by C30, and the 20,000-Hz range is determined by C23 and C30 in parallel. The output signal may be passed through a weighting network which simulates the human ear characteristic (Fig. 22-62C).

Referring to Fig. 22-62I, input transistors Q5 to Q7 combine to form the equivalent of a cathode follower with an overall gain of approximately unity. Transistor Q5 operates at a collector current of 10 μA. It produces a current gain sufficient to drown the noise of Q6, but produces very little noise within itself, thus resulting in low noise plus a high input impedance.

Emitter follower Q7 has a 2000-ohm load resistor in its collector circuit to feed the following attenuator, which is controlled by the voltmeter range-selector switch. Since the attenuator itself presents a 2000-ohm load, the resultant ac collector load is 100 ohms.

Fine-adjustment attenuator (R52 to R61) is a simple potentiometer with four 10-dB steps, providing 10-, 20-, 30-, and 40-dB losses for the 10-, 3-, 1- and 0.3-volt ranges respectively. When the voltmeter range is set for 0.1 volt and the fine attenuator switch out, the input is attenuated 50 dB by R40, R41, and RV10 (frequency compensated by C40 and C41). When the voltmeter range is lowered, the four 10-dB steps are again switched in.

Transistor Q13 is the first transistor of the main voltmeter amplifier and has a gain on the order of 3. The base of buffer emitter-follower Q14 is biased by dc feedback from the output stage. Resistor R78 provides preliminary bias to protect Q15 from saturation when the instrument is turned on. Amplifier stage Q15 is a pnp-type transistor. Its gain is adjusted by the ac-coupled Calibrate preset potentiometer on the front panel. The collector circuit is composed of 3000-ohm resistor R84 and 1000-ohm R85 in series, the latter resistor being decoupled with 10-μF capacitor C61 to flatten out the low-frequency response.

Driver transistor Q16 has its dc conditions set by RV14, so that the output transistors have a centralized dc output to prevent limiting or low gain. The output stage consists of Q17 and Q18 connected in push-pull. Their bases are clamped 6 volts apart by diode MR11. Each emitter circuit has a 220-ohm resistor in series with an 820-ohm resistor bypassed by a 500-μF capacitor to stabilize the ac gain and dc operating conditions. Output terminals taken from the emitter of Q18 appear on the front panel for the connection of an oscilloscope to observe the waveform of the distortion products, or a wave analyzer for measuring the amplitude of the individual harmonics.

The metering circuit consists of two diodes connected in a full-wave bridge rectifier circuit. It will be noted the rectifier circuit is not returned to ground through the feedback loop, as is done in many wide-range sensitive voltmeter circuits. Total harmonic distortion may be measured from 0.10 percent full-scale to 50 percent, and noise level to 80 dB below a reference level of 1 milliwatt.

The schematic diagram for the power supply section is given in Fig. 22-62J and may be either battery operated or from the ac mains. As such power supplies have been discussed elsewhere, it will not be discussed here, except to call attention to diode MR6 in series with the negative terminal of the battery, to provide protection against accidental reversal of the battery polarities.

A second distortion-factor meter, Model 1932A, manufactured by General Radio Co., is shown in Fig. 22-62K. This instrument is completely vacuum-tube operated. An elementary diagram of its circuitry appears in Fig. 22-62L. The principal components are a high-gain amplifier system, with an RC interstage coupling unit that is balanced to a sharp null using a variable air capacitor (in contrast to the Marconi instrument using resistance) in a parallel-T network. This system is followed by a vacuum-tube voltmeter of high sensitivity. Degeneration is used to maintain a high degree of stability and ensure a wide frequency response. The null frequency is continuously variable and can be switched out of the circuit for making noise measurements of the equipment under test. When the network is balanced, the fundamental frequency is reduced 80 dB or more, leaving only the harmonics. These harmonics are measured as a whole and are indicated on the meter as total harmonic distortion (THD).

Two input circuits are provided, 600 ohms (internal impedance: 10,000 ohms) for balanced and unbalanced-line operation, and a direct connection into a 100,000-ohm potentiometer grounded on one side. Fundamental frequencies of 50 to 18,000 Hz and harmonics up to 55,000 Hz may be measured. Full-scale readings of 0.3, 1.0, 3, 10, and 30 percent are provided, with a 100-percent scale for setting the initial input level. The meter is calibrated to indicate both percent distortion and decibels, and it will indicate noise levels 80 dB below a reference calibration level, or 80 dB below zero dBm. The ballistics of the meter are similar to the standard VU and volume indicator meters. Terminals are provided at the rear for rack mounting. Push buttons are used for selecting the meter sensitivity and the frequency range of the null network.

To operate either of the previously described instruments, the procedure is the same except for slight differences in the terminology used with the controls. The proper input impedance is selected for the device from which the signal is to be obtained. Set the controls to Calibrate. Adjust the variable calibrate control for a 100-percent deflection of the meter. Set the controls to Distortion with the main dial set to the fundamental frequency of the signal under measurement. Select the proper frequency range for the null network. Tune the main dial and trimmers for a minimum (null) reading of the meter. Continue this procedure, increasing the meter sensitivity until a range is reached where the sensitivity cannot be further increased. Read the distortion, taking into consideration the meter range. Thus, if a minimum balance is on a 1-percent scale and the meter reads 3, the THD is 0.3 percent. The distortion measured is that of the remaining harmonics, which may be identified by displaying the waveform on an oscilloscope.

22.63 Describe an automatic-nulling distortion meter.—An automatic-nulling distortion analyzer consists fundamentally of a Wien-bridge circuit for rejecting the fundamental frequency, operating with two control loops for automatically tuning the two legs of the bridge circuit which rejects the fundamental frequency. Distortion is read directly on a meter calibrated to read percent distortion and decibels. The meter may also be used to measure the internal noise of the device under test. Instruments designed for use in the broadcasting industry may include an rf detector for measuring percent distortion of the modulated carrier. Such an instrument, the Model 334A, manufactured by the Hewlett-Packard Co., is pictured in Fig. 22-63A. The instrument to be discussed is completely solid state.

Referring to the block diagram in Fig. 22-63B, at the upper left is an a-m detector which is used to detect the modulating signal from an rf carrier. The rf components are filtered from the modulating signal, which is then applied to an impedance converter circuit.

For distortion measurements the impedance converter provides a low-noise input circuit, with a high impedance independent of the source impedance at the terminals. It also provides unity gain between the input of the instrument and the input of the rejection amplifier. The input signal is applied to the impedance converter through function-switch S1 and a 1-megohm attenuator S3. The attenuator network provides 50 dB of attenuation in 10-dB steps. The desired attenuation level is selected by sensitivity switch S2. The rejection amplifier consists of a preamplifier, a Wien bridge, and a bridge amplifier. A sensitivity vernier control, R1, at the input of the preamplifier provides a set-level signal to obtain a full-scale reading on the meter for any voltage level at the input of the instrument.

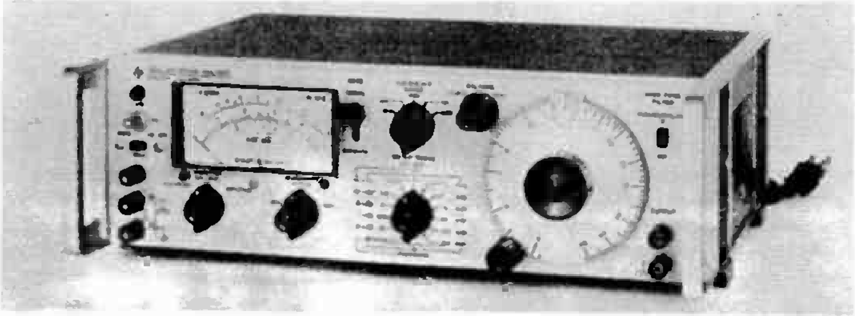


Fig. 22-63A. Hewlett-Packard Model 334A automatic null distortion analyzer.

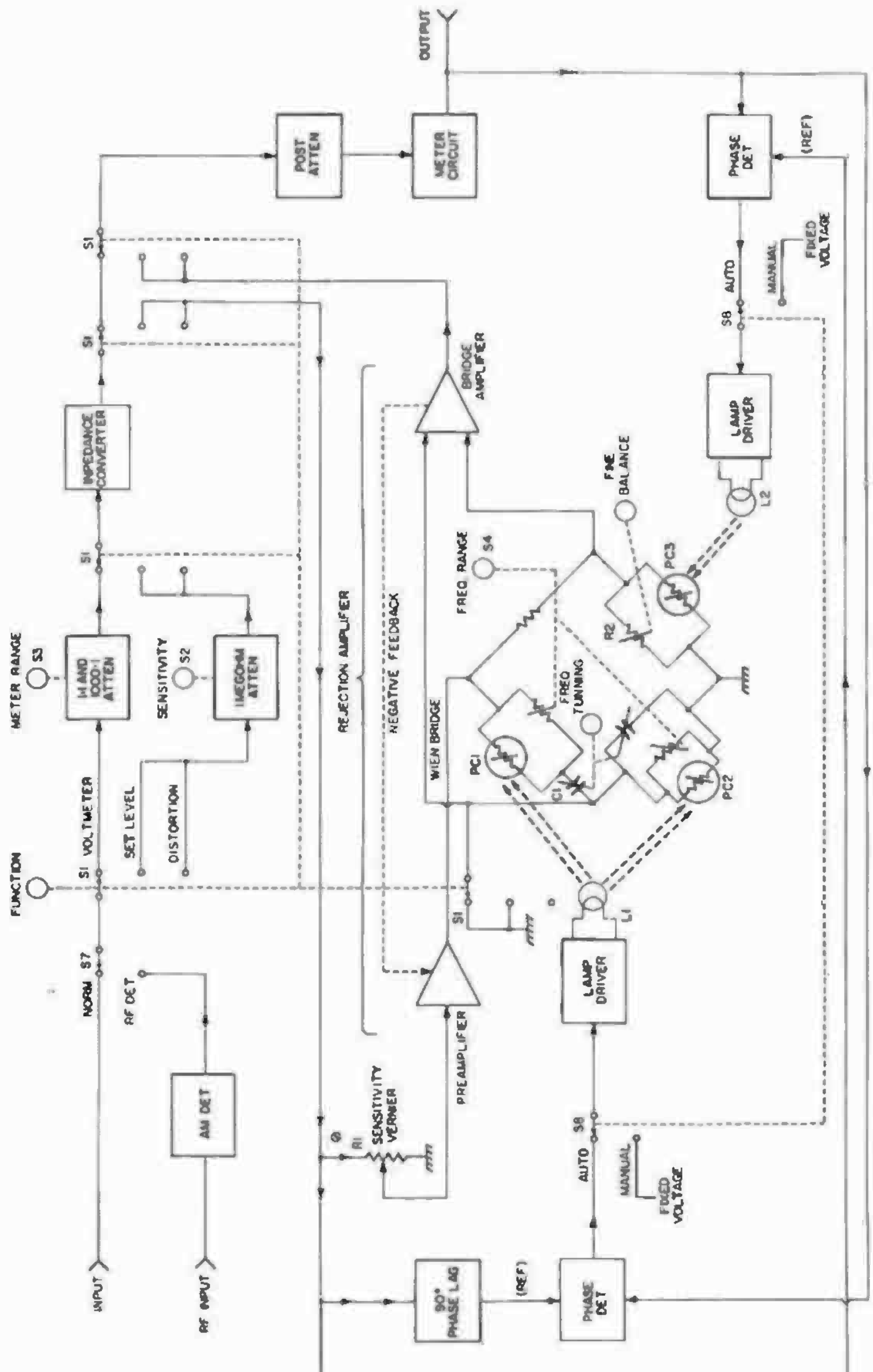


Fig. 22-63B. Block diagram for Hewlett-Packard Model 334A automatic null distortion analyzer.

With function switch S1 in the set-level position, a ground is applied to the Wien-bridge circuit to allow a signal reference level to be set on the meter. With the function switch in the distortion position, the Wien bridge is used as an interstage coupling network between the preamplifier and bridge amplifier. The Wien bridge is tuned and balanced to reject the fundamental frequency of the applied signal. Two automatic loops consisting of two phase detectors, lamp drivers, lamps, and photocells provide fine tuning and balance the automatic mode. The remaining frequency components are applied to the bridge amplifier and are measured as distortion by the metering circuit.

Negative feedback from the bridge amplifier to the preamplifier narrows the normal rejection response of the Wien bridge, similar to that shown in Fig. 22-62D. The output from the rejection amplifier is applied to the metering circuit through a post attenuator. This attenuator limits the input signal applied to the metering circuit to 1 millivolt for full-scale deflection. Output terminals in the metering circuit permit the connection of an oscilloscope for observing the character of the distortion components.

In the voltmeter mode of operation, the input signal is applied to the impedance converter through a 1:1 and 1000:1 attenuator. The 1:1 ratio is used in the 0.0003- to 0.3-volt position of meter range switch S3, and the 1000:1 ratio is used in the 1- to 300-volt range position. With the function switch in the voltmeter position, the output of the impedance converter bypasses the rejection amplifier and is applied to the metering circuit through the post-attenuator network.

The rejection circuit consists of the preamplifier, Wien-bridge resistive leg, the automatic control loop and its associated lamp and photocell, and the bridge amplifier. In the distortion position, the Wien-bridge circuit is used as a rejection filter for the fundamental frequency of the input signal. The bridge circuit is connected as a interstage coupling network between the preamplifier and the bridge amplifier. The bridge is tuned to the fundamental frequency of the incoming signal by setting frequency range switch S4 to the applicable range and tuning capacitors C1 and C2. The bridge is further balanced by fine balance control R2. In the automatic mode, fine tuning and balancing are accomplished by photocells PC1, PC2, and PC3 connected in the resistive and reactive legs of the bridge circuit. Error signals for driving the photocells are derived by detecting the bridge output, using the input signal as a reference.

When the Wien bridge is not tuned exactly to the frequency to be nulled, a portion of the fundamental frequency will appear at the output of the bridge. The phase of the signal will depend on which leg of the bridge is not tuned or on the relative error in tuning, if neither is set correctly. The magnitude of the signal is proportional to the magnitude of the tuning error of either or both legs of the bridge. The control loops derive their information from a common source and develop two independent control signals for nulling the two legs of the bridge. These control voltages are used to vary the brilliancy of lamps L1 and L2, which in turn cause a resistance change in the photocells which are a part of the bridge circuit. With the bridge in tune and balance, the voltage and phase of the fundamental frequency appearing at the junction of the series reactive and shunt reactive legs is the same as at the midpoint of the resistive leg. When these two voltages are equal and in phase, the fundamental frequency is balanced out. For frequencies other than the fundamental the reactive leg of the bridge offers various degrees of attenuation and phase shift, which cause a voltage at the output points of the bridge. This voltage is amplified by the bridge circuit and applied to the metering circuit.

Negative feedback from the output of the bridge amplifier is applied to the preamplifier to narrow the frequency rejection characteristic of the bridge. However, the normal rejection characteristic of a Wien bridge is not constant. Typically, the second harmonic is attenuated several dB more than the third, and the third more than the fourth. The use of negative feedback sharpens the rejection characteristic considerably, as shown in Fig. 22-62D. When the fundamental frequency is 1 kHz or higher, a T-filter section may be switched in the output of the bridge amplifier, which attenuates 50- or 60-Hz components 40 dB (100:1), but offers no attenuation to frequencies over 1 kHz (Fig. 22-62E).

The metering circuit consists of a bridge-type circuit with a diode in each upper branch and a dc milliameter connected across the midpoints of the bridge. Capacitors are also used in the lower legs of the bridge and are an essential part of the feedback loop. The mechanical inertia of the meter movement prevents the movement from responding to individual current pulses; therefore the meter pointer reading corresponds to the average value of current pulses rather than to the peak value. The meter is calibrated to read the rms value of a sine wave. The power supply is series regulated and has a positive-negative 25-Vdc output.

The general characteristics of the instrument are: input impedance, 1 megohm shunted by less than 60 pF; distortion measurement range, any frequency between 5 Hz and 600 kHz; distortion levels, 0.1 to 100 percent full-scale in seven ranges; elimination characterstic fundamental rejection 80 dB; input sensitivity 0.3 volt rms for 100 percent.

22.64 ***What is a phase-shifter cancellation-type distortion-factor meter?***—A distortion-factor meter which measures the distortion factor by suppressing the fundamental frequency by the use of a second signal which is of the same frequency and amplitude, but 180 degrees out of phase with reference to the signal being applied to the input of the device under test. A block diagram of a typical phase-shift of distortion-factor meter (DFM) with its external connection is given in Fig. 22-64A.

The meter indicates the distortion factor, which is the ratio of the rms total distortion to the amplitude of the fundamental frequency, including harmonics. This is accomplished by suppressing the fundamental frequency and measuring the rms total of the remaining harmonics. Phase shift is achieved by the use of a group of adjustable RC phase-shifting networks operated from the front panel.

Referring to the external connections in Fig. 22-64A, the signal from the oscillator is split into two branches. The first branch connects to a variable attenuator or gain set, to the iput of the amplifier under test, and from the amplifier output to the DFM. The second branch connects from oscillator to the DFM through a 20-dB pad in the cancellation signal loop. The cancellation signal is applied to a phase-shifting

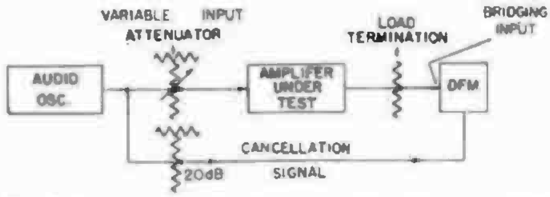


Fig. 22-64A. Block diagram of external connections for a cancellation-type distortion meter.

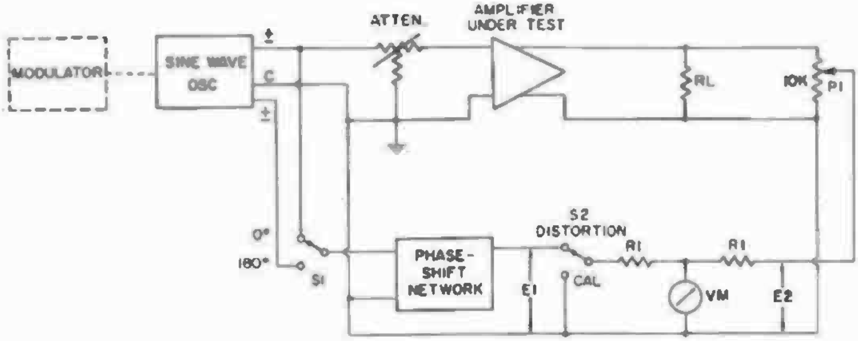


Fig. 22-64B. Method of measuring total harmonic distortion with a bridge that is frequency insensitive.

network in the DFM. The meter is adjusted in the calibrate position for a 100-percent deflection. The phasing controls are then adjusted for a minimum reading on the DFM, by increasing the sensitivity of the metering circuit. When a minimum reading has been obtained, the remaining indication is the total harmonic distortion (THD) in percent of the fundamental frequency. (See Question 22.62.)

The distortion factor measured by this meter may be evaluated:

$$D = \frac{\sqrt{E_2^2 + E_3^2 + E_4^2 + \ldots}}{E_1}$$

where,

E_2, E_3, and E_4 equal the voltages of the individual harmonic voltages,

E_1 is the voltage of the fundamental frequency, including the harmonic voltages.

A second type of phase-shift distortion-factor meter is given in Fig. 22-64B. This circuit is termed a "frequency-insensitive bridge." The same results are obtained as for the previous circuit, except in a slightly different manner. In this instance, after the amplifier output level has been established, the circuit is balanced by throwing phase switch S1 to one of two positions. S2 is placed in the calibrate position and potentiometer P1 is adjusted for a reference full-scale reading on the meter. Switch S2 is then thrown to the distortion position and the phase-shifting network is adjusted for a minimum reading on the meter. Under a condition where E_1 equals E_2, the fundamental frequency of the input signal is nulled out, leaving only the harmonics and noise. Phase switch S1 may have to be switched from zero to the 180-degree position to obtain the lowest reading. The ratio of the voltmeter reading in the calibrate and distortion positions (null) is the total harmonic distortion (THD) in percent of the fundamental frequency.

22.65 ***Describe the basic principles of a harmonic wave analyzer.***—A wave analyzer is an instrument designed for measuring the amplitudes of the individual components of a complex waveform such as might be encountered in making harmonic or intermodulation distortion measurements. Harmonic analysis of both acoustical and mechanical motion can be analyzed by the use of a proper transducer. Wave analyzers can also serve as a tunable narrow-band filter, so that any component of a complex waveform can be extracted and used to drive a graphic recorder or frequency counter. Such instruments operate on the superheterodyne principle, much like a superheterodyne radio receiver, only in this instance the frequency band is generally confined to a range of 20 Hz to 60 kHz.

The instrument consists of a local variable oscillator, a highly selective amplifier, and a wide-range metering circuit. The local oscillator modulates the incoming frequency to produce a constant difference frequency. This latter frequency is applied to a narrow-band intermediate-frequency (i-f) amplifier whose output voltage is proportional to the magnitude of the incoming voltage. The analyzer may also be considered to be a highly selective voltmeter.

The term "heterodyne" is defined as being a method of combining two frequencies in such a manner as to produce a third frequency which is the

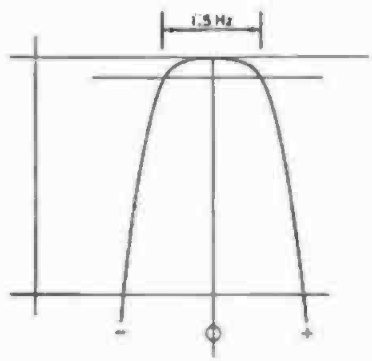


Fig. 22-65A. Frequency characteristic of a typical quartz crystal intermediate-frequency amplifier.

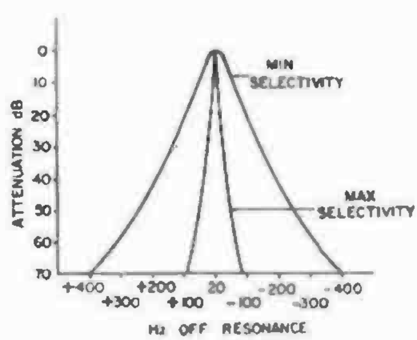


Fig. 22-65B. Selectivity curve for an i-f amplifier using positive and negative feedback.

result of beating the first two frequencies together. The beat frequency may be either the sum or difference of the two frequencies and is the basic principle used in wave analyzers.

The local oscillator of the wave analyzer is caused to beat with the frequency under observation, and the resulting beat frequency is applied to the selective amplifier. The amplitude of the beat frequency is measured at the amplifier output with a calibrated vacuum-tube voltmeter.

The selective or i-f amplifier may be one of two designs: quartz crystal or positive and negative feedback. In the older crystal designs the bandwidth of the i-f amplifier was generally about 4 Hz, while those using the feedback principle could be adjusted by the operator from a few hertz up to about 200 Hz. Typical selectivity response curves for a quartz crystal and negative feedback-type analyzers are given in Figs. 22-65A and B respectively. Modern wave analyzers are of transistor design, using i-f bandwidth amplifiers, variable in fixed steps ranging from 3 to 50 Hz.

The advantage of the variable selectivity analyzer over the fixed selectivity type is that if a harmonic analysis is being made of a device which has mechanical motion such as a disc, tape, or film recorder, a certain amount of flutter is encountered. If the selectivity of the intermediate amplifier is too great, the signal passes in and out of the selective amplifier passband, making the measurement of low amplitudes difficult. A selective amplifier with variable bandwidth may be adjusted for a bandwidth suitable for a given measurement. If the signal source is steady, such as would be obtained from an oscillator, either type of analyzer is satisfactory.

Fig. 22-65C. Marconi Instruments Ltd (England) Model TF2330 Wave Analyzer.

The local oscillator must be of stable design, and it may be either an RC or LC type. In both designs the shaft of the tuning capacitor is connected to a dial on the front panel. The dial is calibrated in terms of frequency, from 20 to 60,000 Hz, although the actual frequency range of the local oscillator is 100 to 160 kHz. The frequency of the local oscillator depends on the frequency of the signal under observation. The local-oscillator frequency is either the sum or difference of the i-f amplifier frequency and the input signal frequency. As an example, assume the analyzer uses a tuned amplifier of 50,000 Hz and the signal under measurement is 400 Hz. The local oscillator must be tuned to a frequency of 49,600 Hz (49.6 kHz). This frequency will produce a beat frequency between the incoming signal and the local oscillator of 50,000 Hz (400 plus 49,600). Another example is if the incoming frequency is 16,000 Hz, the local oscillator must be set to a frequency of 34,000 Hz. The beat frequency will also be 50,000 (16,000 plus 34,000).

The scale of the indicating meter is calibrated in volts; however, it may be read directly in percentage, if desired. The dial of the local oscillator is calibrated in audio frequencies so that the amplifier output is proportional to the amplitude of the frequency to which the local oscillator is set. An internal means of calibration is provided whereby a signal of known amplitude is applied for adjusting the gain of the intermediate amplifier for a standard output. Once this adjustment is made, the meter is reading directly in volts. Wave analyzers employing negative feedback in the intermediate amplifier are quite similar in their operation to the quartz-crystal type; therefore, it is unnecessary to go into the details of their design.

Wave analyzers are used to measure the distortion of a complex waveform by measuring the amplitude of the individual harmonic voltages. The total harmonic distortion may be computed:

$$\% \text{ Dist.} = \sqrt{\%f_2^2 + \%f_3^2 + \%f_4^2 + \ldots}$$

where,

f_2, f_3, aud f_4 equal the amplitude of the individual harmonic voltages in percent of the fundamental frequency.

A model TF2330 wave analyzer manufactured by Marconi Instruments, Ltd.,

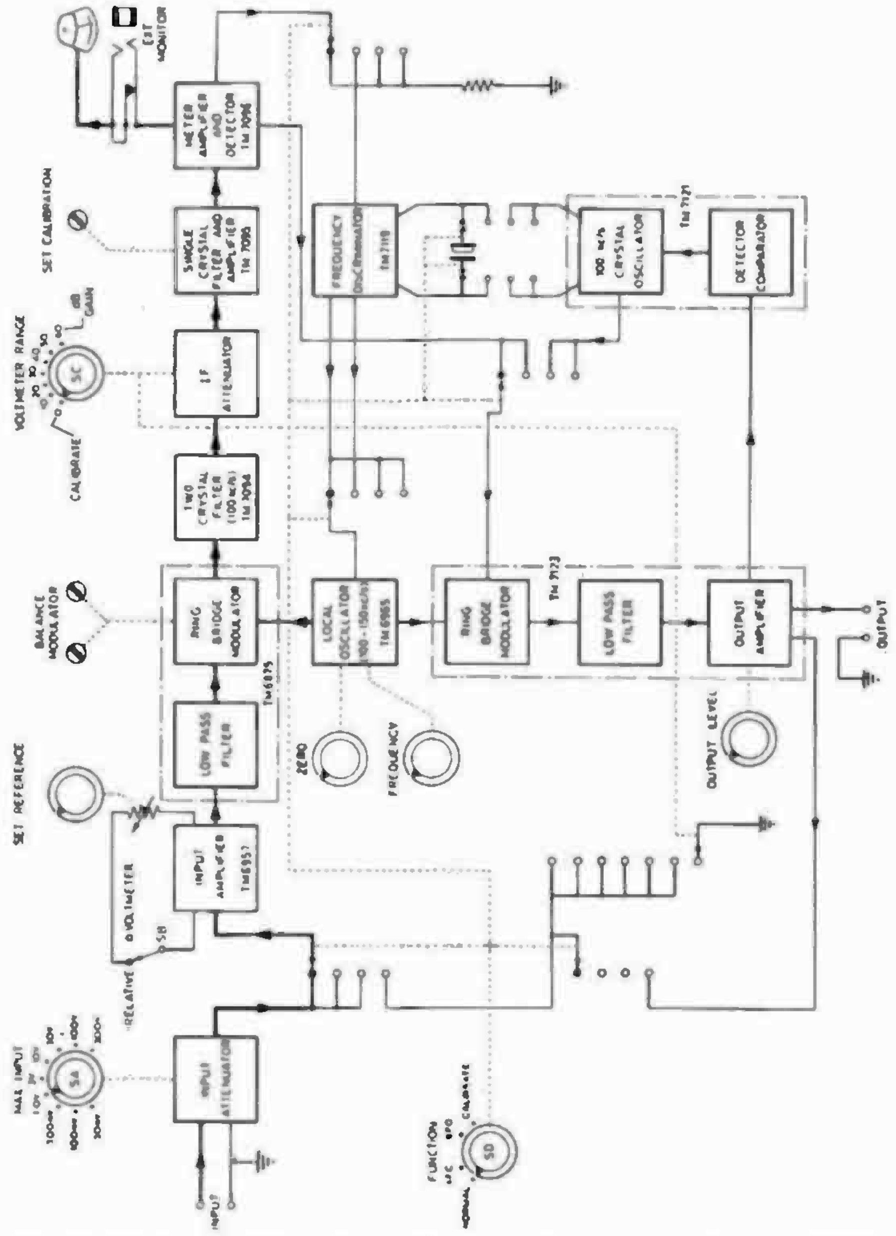


Fig. 22-65D. Simplified block diagram for Marconi Instruments Ltd. Model TF2330 wave analyzer.

of England, is shown in Fig. 22-65C. This instrument operates on the superheterodyne principle and covers a frequency range of 20 Hz to 50 kHz, using an intermediate frequency of 100 kHz. Referring to the block diagram in Fig. 22-65D, the incoming signal is fed via an input attenuator to an input amplifier. The signal then passes through a low-pass filter to a ring-bridge modulator. The low-pass filter ahead of the modulator prevents any 100-kHz component in the input signal from reaching the intermediate amplifier stages.

Also fed to the ring-bridge modulator is the output of a variable-frequency local oscillator, ranging from 100 to 150 kHz. The resulting difference signal from the ring-bridge modulator is fed to the first part of a crystal filter, having a 7-Hz passband. This 100-kHz difference signal is then amplified and

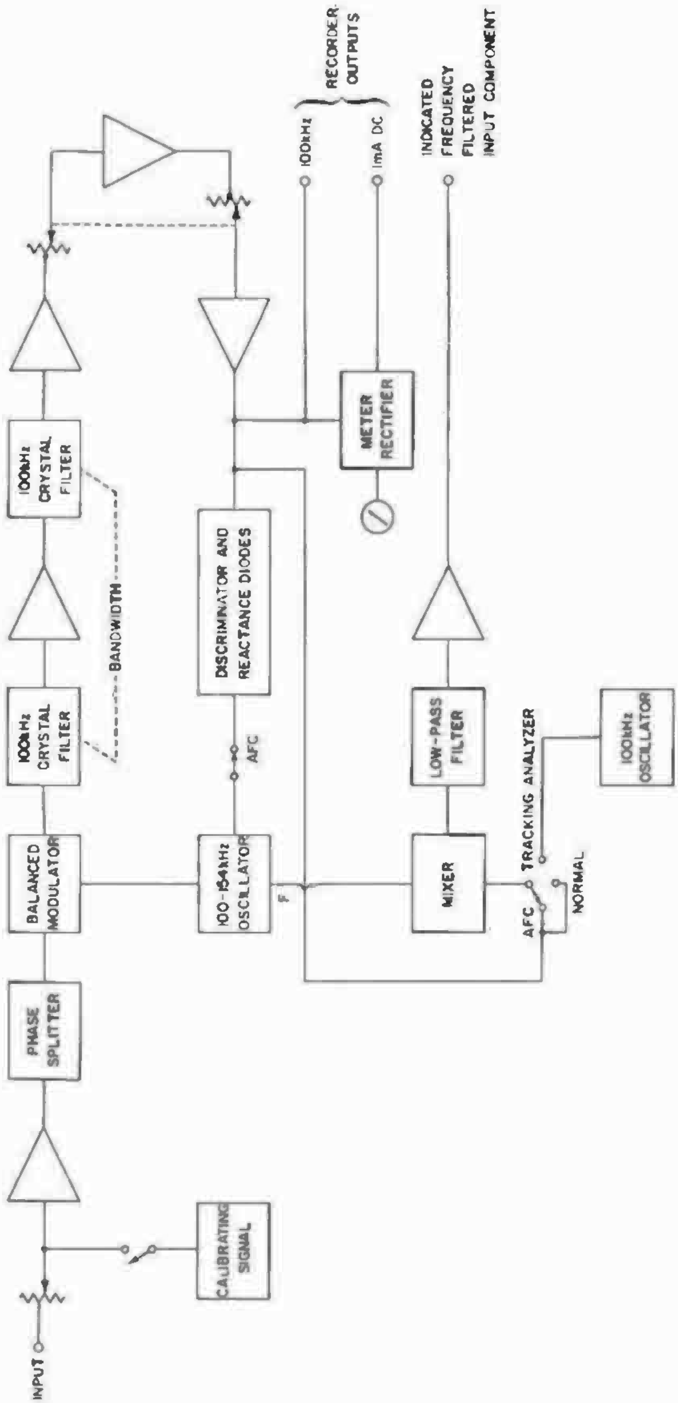


Fig. 22-65E. Simplified block diagram for General Radio Co. Model 1900-A wave analyzer.

passed to the i-f attenuator, a ladder network having six steps of 10 dB each.

From the attenuator the signal goes to the second part of the crystal filter and another amplifier stage. The latter is followed by the i-f amplifier which has adjustable gain set by the calibration control. The i-f output signal is fed to the meter amplifier and a rectifier circuit. A linear meter scale is obtained

Fig. 22-65F. General Radio Co., Model 1900-A wave analyzer.

by having the rectifier circuit in a negative feedback loop of the amplifier. A separate output from the meter amplifier is used for automatic frequency control (afc) purposes, and when mixed with the local oscillator output provides the restored frequency output.

Permeability tuning is used in the local-oscillator circuit with padder and trimmer inductances in the tuned circuit to enable the frequency to be adjusted for a precalibrated frequency scale. A crystal discriminator provides the afc signal which is applied to a voltage-variable capacitor included in the tuned circuit. A restored frequency output is derived by combining the i-f signal with the local-oscillator output in another ring-bridge modulator. The difference frequency, being that of the original input signal, is filtered and amplified before being fed to the output terminals via the Output Level control.

A beat-frequency oscillator (BFO) output is similarly derived by combining the output from a separate 100-kHz crystal oscillator with the local oscillator output. A signal taken from the output amplifier via a detector comparator provides an error signal which is added to the feedback of the 100-kHz oscillator to control its amplitude. This results in a signal of stable output level, which is utilized by also making it the calibration reference used when the gain of the voltmeter section is set up. A relay switches the reference signal to the input amplifier. The same 100-kHz crystal is used for both the discriminator and the oscillator, since the two circuits are not used simultaneously.

Power for the instrument is derived from a series stabilized power supply, with an output voltage of 15 volts. The instrument may be operated from the ac line or external batteries. The device is completely transistorized, with the components mounted on printed-circuit boards. The local oscillator is tuned by an iron-dust tuning core actuated by a precision lead screw of 10 turns per inch. The required tuning range of 100 to 150 kHz is achieved by a movement of 0.85 inch, or 8.5 turns of the dial. The appropriate portion of the scale is indicated by a colored marker behind the dial.

A simple mechanical computer is used to combine the fuuctions of the two attenuators. A dial mounted in the voltmeter range-switch spindle, but driven by the maximum input switch, is printed with full-scale values. The dial is viewed through a window on the front panel arranged so that only seven ranges are visible at one time, this being the coverage of the meter range switch.

Fig. 22-65E is a simplified block diagram for a General Radio Co. Model 1900-A wave analyzer. The incoming signal to be analyzed is applied to a calibrated attenuator, then passed through a phase splitter and into a balanced modulator circuit, where it is then heterodyned by a local oscillator. The frequency of this local oscillator is adjusted so that the difference frequency between it and the desired component (harmonic) is 100 kHz, the frequency of the intermediate amplifier. The i-f amplifier is highly selective, but it can be adjusted for a bandwidth of 3, 10, or 50 Hz by means of a front-panel control. The 100-kHz signal from the filter is amplified and adjusted by means of a second attenuator. The signal is then indicated on a meter. In one mode of operation the signal is heterodyned back to the original frequency and is indicated on the front panel as Filtered Input Component (Fig. 22-65F). In the second mode the local oscillator is caused to beat with a 100-kHz quartz crystal oscillator and the combination functions as a beat-frequency oscillator. This output is also available at the front panel and is indicated as Tracking Analyzer.

The frequency of the local oscillator is made adjustable over a frequency range of 100 to 150 kHz by means of two large coaxial frequency knobs. The difference between the actual oscillator frequency and the 100-kHz intermediate freqnency is indicated on a counter-dial combination. A capacitor in the oscillator circnit, with its dial on the panel indicating Δf (delta f), can be used to change the indicated frequency plus or minus 100 Hz of any indicated setting of the frequency controls. A front-panel adjnstment provides a three-speed response for the meter movement, slow, medium and fast. The slower speeds are recommended for noise measurements. The analyzer pictured may be mechanically coupled to a graph level recorder.

22.66 What Is a cross-modulation oscillator?—A special type of modulated oscillator used when recording variable-area sonnd tracks on photographic film for motion pictures. The cross-modulation oscillator permits the determination of the correct negative and print densities. Such oscillators are discussed in Questions 18.230, 18.235, and 23.185.

22.67 What is a cross-modulation readout panel?—A filter panel used for measuring the distortion components of a cross-modulation test for the determination of negative and print densities when recording on motion picture film using a variable-area recording system. This equipment is discussed in detail in Question 18.234. (See Question 22.66.)

22.68 What is a secondary phase-standard?—It is a device for shifting the phase of an electrical circuit a given number of electrical degrees, or determining the unknown phase shift of a device. Generally, such devices are referred to as *phase generators*. A Model PG-3 phase generator, manufactured by Theta Instrument Corp., is pictured in Fig. 22-68A. The internal circuitry of such a device is given in Fig. 22-68B. It consists of four coils attached to a dial calibrated 0 to 360 degrees, with a vernier scale for reading fractional parts of a degree.

To determine the phase shift of a device, the phase generator is connected as shown in Fig. 22-68C. Initial calibration is made by first bypassing the device with the unknown phase shift (shown by dotted line), and applying the test signal to the X and Y axis of the oscilloscope to determine the internal phase shift of the oscilloscope amplifiers. The phase-generator dial is rotated until a vertical straight line appears on the scope. The dial of the phase generator is then set to read zero and locked. The device with the nnknown phase shift is then returned to the circuit, and the phase-generator dial is again rotated for a straight vertical line. The reading of the phase-generator dial is the phase shift in electrical degrees of the device nnder test. If a considerable amount of distortion is present in the signal, the line will be looped and may be difficult to determine accurately.

A more accurate measurement may be made using the circuit of Fig. 22-68D. Here, the device of unknown phase and the phase generator are connected to a transformer T1, a bandpass filter, and a sensitive voltmeter. The filter has a passband suitable for the frequency of the test signal. The dial of the phase generator is rotated until the voltmeter indicates a null. At this point the device of unknown phase and the phase generator are in phase.

Fig. 22-68A. Phase generator Model PG-3, manufactured by Theta Instrument Corp.

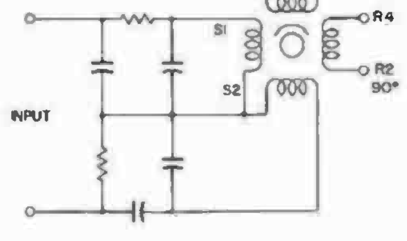


Fig. 22-68B. Internal connections of a typical phase generator.

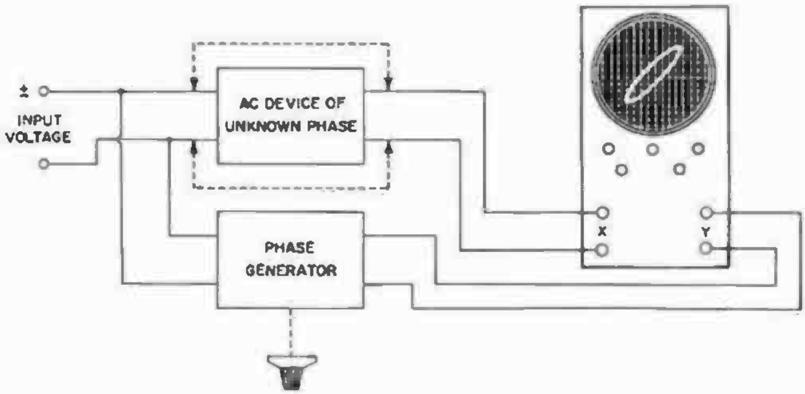


Fig. 22-68C. Phase generator connected for determining the phase shift of an unknown device.

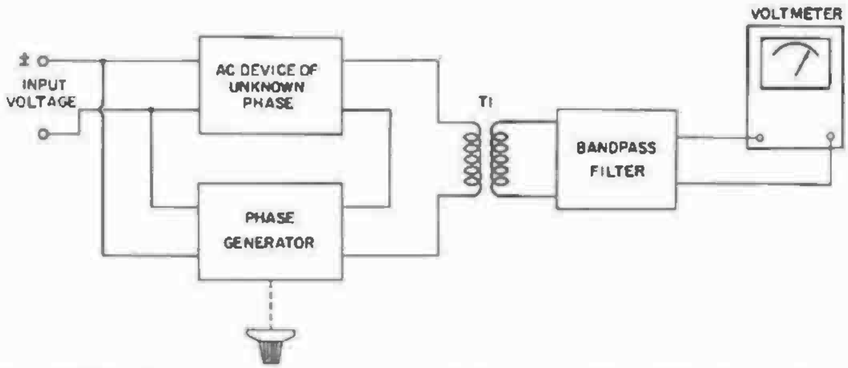


Fig. 22-68D. Phase generator connected to indicate the phase-shift of an unknown device using a transformer, bandpass filter, and sensitive voltmeter.

22.69 Describe the function and basic components of a cathode-ray oscilloscope.—A cathode-ray oscilloscope is an electronic instrument for displaying a visual curve of an electrical quantity as a function of a second electrical quantity or time. An oscilloscope is not a device within itself for accomplishing a purpose or for performing a designed operation on an electrical signal, but rather an instrument which will display the electrical characteristics of the circuit to which the oscilloscope is connected.

Many different types and designs of oscilloscopes are available. Oscilloscopes may be obtained with the horizontal and vertical amplifiers as an integral part of the instrument, or in plug-in forms which supply the circuitry for single-, dual-, or multiple-trace displays or other functions. The circuitry may be solid state or vacuum tube. In many models the instruments are of the hybrid type, utilizing the best features of vacuum tubes and solid-state components. Cathode-ray-tube display screens are generally 3 or 5 inches in diameter; however, in some instances they may be 8 inches or larger. The phosphor coating of the screen can be of long, medium, or short persistence, depending on the display required. Provision is generally made for mounting a camera over the graticule for photographing the display. The graticule may be mounted external to the tube face, or etched on the interior side of the display area. (See Questions 22.88 and 22.113.)

A basic block diagram of the components and controls for a typical oscilloscope is given in Fig. 22-69A. The components consist of a vertical-amplifier input attenuator A, vertical amplifier B, cathode-ray tube C, sync-amplifier attenuator D, sync circuit E, sweep generator F, horizontal-amplifier input attenuator G, and horizontal amplifier H. Two switches, S1 (a sync-drive selector switch) and S2 (a horizontal-amplifier input switch), along with the power

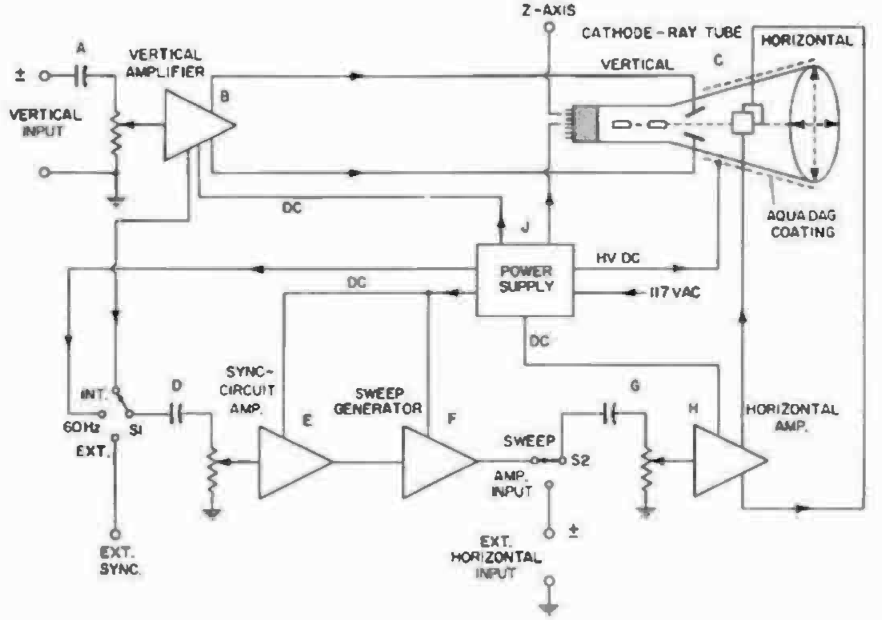


Fig. 22-69A. Simplified block diagram for a cathode-ray oscilloscope. The amplifiers and associated components may be either solid-state or vacuum tubes.

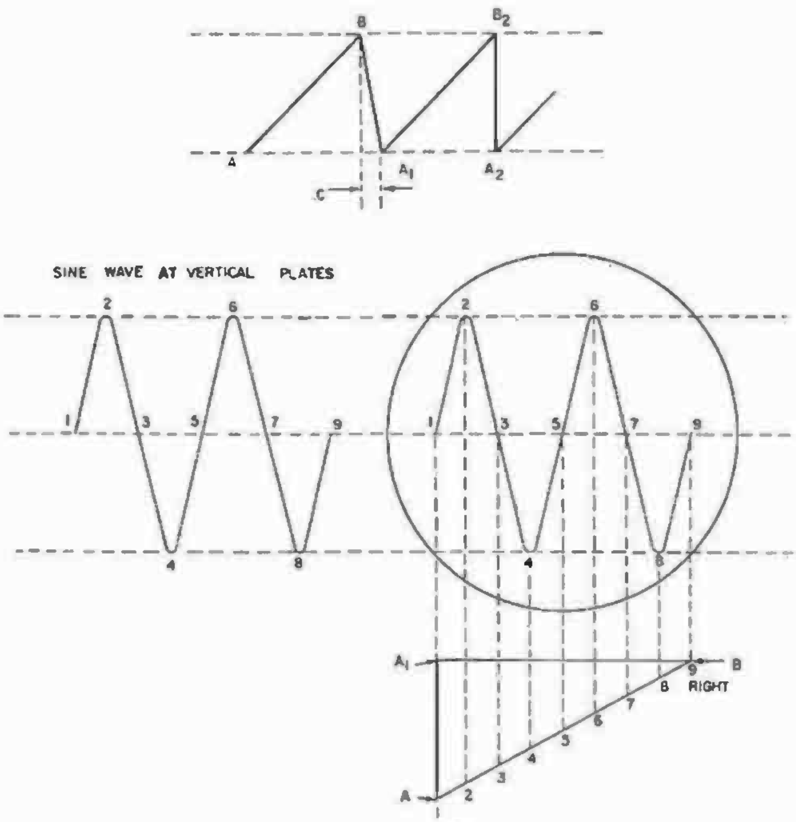


Fig. 22-69B. Basic principle of drawing out waveforms on a cathode-ray tube using a sawtooth oscillator.

supply J complete the major components.

When a waveform is to be observed, the signal voltage is applied to the input of the vertical amplifier, shown at the upper left in the block diagram. The vertical amplifier applies the signal to the vertical deflecting plates of the cathode-ray tube which causes the beam to be deflected in a vertical plane. In the meantime a sweep (or sawtooth) oscillator, working through the horizontal amplifier, deflects the beam in the cathode-ray tube from left to right in a horizontal plane.

If the linearity of the sweep oscillator is uniform and its frequency is eqnal to, or a multiple of, the waveform under observation, a stationary pattern or image will appear on the cathode-ray tube screen. The resulting image is the vertical voltage plotted against time in a horizontal plane. The ratio of the frequency of the vertical voltage to the frequency of the sweep voltage will determine the number of cycles that will be displayed on the screen.

To have a wide range of use, the amplifiers in a cathode-ray oscilloscope must have broad frequency response, low distortion, and negligible phase shift. A properly designed cathode-ray oscilloscope will display sawtooth, square-wave, sine-wave, triggered pulses, and complex waveforms without distortion.

Complex waveforms contain many harmonics which must be amplified proportionally to appear on the cathode-ray tube in the same phase relationship as when applied to the input of the vertical amplifier.

Theoretically, a square wave is composed of a fundamental frequency and an infinite number of odd harmonic frequencies all in phase. If a square wave is passed through an amplifier with a poor frequency characteristic, the waveform will be distorted and the flat portions will no longer be flat but irregular and slanted. Therefore it is highly important that the amplifiers in an oscilloscope have a wide-frequency bandwidth (to 1 MHz) and be capable of passing all frequencies that go to make up a square wave. The horizontal amplifier must also have a good frequency response because it must pass the sawtooth waveform of the sweep oscillator which also contains a large number of harmonics.

The waveform of the sweep generator is a ramp or sawtooth voltage and, when applied to the horizontal deflection plates of the cathode-ray tube, causes the beam to move from left to right at a constant rate of speed, and then rapidly return to the left to start over again.

While the sweep oscillator is moving the beam in a horizontal plane, the signal voltage applied to the vertical plates of the cathode-ray tube moves in a vertical plane. These two forces, one horizontal and the other vertical, trace out the shape of the waveform under observation on the fluorescent screen in multiples of the applied frequency, or as a single trace depending on the frequency of the horizontal oscillator. A sawtooth voltage as applied to the horizontal plates of an oscilloscope is shown in the upper portion of Fig. 22-69B. It will be noted the voltage rises from point A to point B at an even rate of speed, pulling the beam in the cathode-ray tube across the screen from left to right. This rise is called the *linear time rise*. At the end of the cycle the voltage returns to the base line at point A_1, which is the same as point A. In the short time interval indicated at C, the beam returns to the left to start over again. The change from right to left is extremely fast and blanked out by a blanking pulse which cuts off the beam, to ensure that no return trace will be seen. The drawing in the lower portion of Fig. 22-69B shows how the sawtooth voltage and the vertical signal trace out the waveform pattern under observation on the screen of the cathode-ray tube.

The simplest method of obtaining a gradual rise time followed by a rapid discharge time is by charging and discharging a capacitor in a relaxation oscillator or multivibrator. To obtain a stationary image on the screen, the period of the sweep oscillator must be exactly equal to, or a multiple of, the period of the waveform being displayed. For perfect synchronization of the sweep oscillator, a small portion of the input signal voltage is applied to the sweep oscillator circuit. This control voltage, called the *sync voltage*, triggers the sweep oscillator at exactly the correct interval of time. Synchronizing pulses may also be obtained from an external source. (See Question 22.59.)

To display the waveform patterns properly, two positioning controls are provided which apply a dc voltage to the horizontal and vertical deflection plates. Adjustment of these controls permits the image to be moved either in a vertical or horizontal direction, to center the image on the screen. The movement of the image in no way affects the display.

To accelerate the electron beam in the cathode-ray tube to a high velocity, a potential of several thousand volts is required between the cathode and the accelerating anode. As the elements require only a few milliamperes of current for their operation, the power supply may be made quite simple. The amplifiers and associated equipment require a source of well filtered dc. This is obtained from a full-wave regulated power supply with a heavily shielded power transformer. The cathode-ray tube employs a voltage of several thousand volts, supplied from a half-wave rectifier. A magnetic shield is placed around the cathode-ray tube to protect it from stray magnetic fields. Cathode-ray tubes are discussed in Question 11.91.

22.70 ***Describe an oscilloscope using a plug-in preamplifier.***—A basic oscilloscope made by Marconi Instruments, Ltd., and shown in Fig. 22-70A, is designed for plug-in preamplifier operation. Basically the instrument (without the plug-in preamplifier) consists of a cathode-ray tube, power supplies, horizontal amplifier, time base, and trigger circuits. At the lower left of the main frame is an opening for inserting a vertical preamplifier unit. The plug-in unit pictured is a dual-trace preamplifier. By loosening a knurled nut at the lower edge, the unit may be interchanged with a single-trace preamplifier or one with other characteristics.

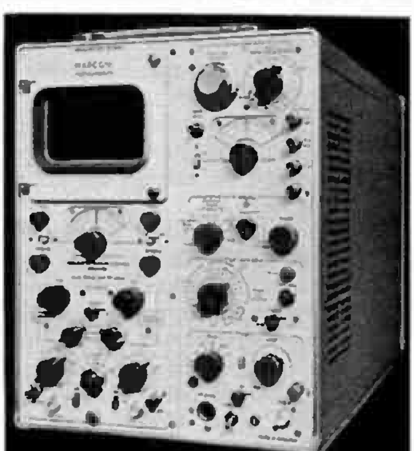


Fig. 22-70A. Marconi Instruments Ltd. (England) Model TF-2200 cathode-ray oscilloscope designed for plug-in preamplifier operation. The preamplifier is plugged-in at the lower left. (Courtesy, Canadian Marconi Instruments Ltd.)

An interesting feature of this instrument is that in addition to the use of the graticule for time and voltage measurements, an alternate system of measurements is provided which is independent of the amplifier gain settings and gives greater accuracy for both axis. This sytem uses the shift method, in which the trace is first positioned with respect to a convenient reference point on the graticule, and then repositioned by a second shift control to bring another point on the trace against the reference. The second shift control is a potentiometer which is supplied with a stabilized dc voltage, and introduces known voltages into the deflection amplifiers. Its scale is graduated in terms of the vertical (Y axis) input voltage or in terms of the time base according to the axis.

The right-hand portion of the instrument is occupied by two time-base generators and the function and trigger circuits. Direct access to the CRT deflection plates for certain measurements is made possible by removing a portion of the outer case.

A block diagram of the basic unit with the connections for a plug-in amplifier unit (upper left of diagram) is given in Fig. 22-70B. The bandwidth and sensitivity are dictated by the particular preamplifier in use. For a single-trace unit it is dc to 40 megahertz, and for a dual-trace unit it is dc to 35 megahertz at a sensitivity of 50 millivolts per centimeter.

Controls are provided on the dual-trace plug-in unit for operating either of the two inputs separately or simultaneously, or they may be selected to provide the following displays:

Dual trace, alternate sweep input A and B
Single trace, channel A only

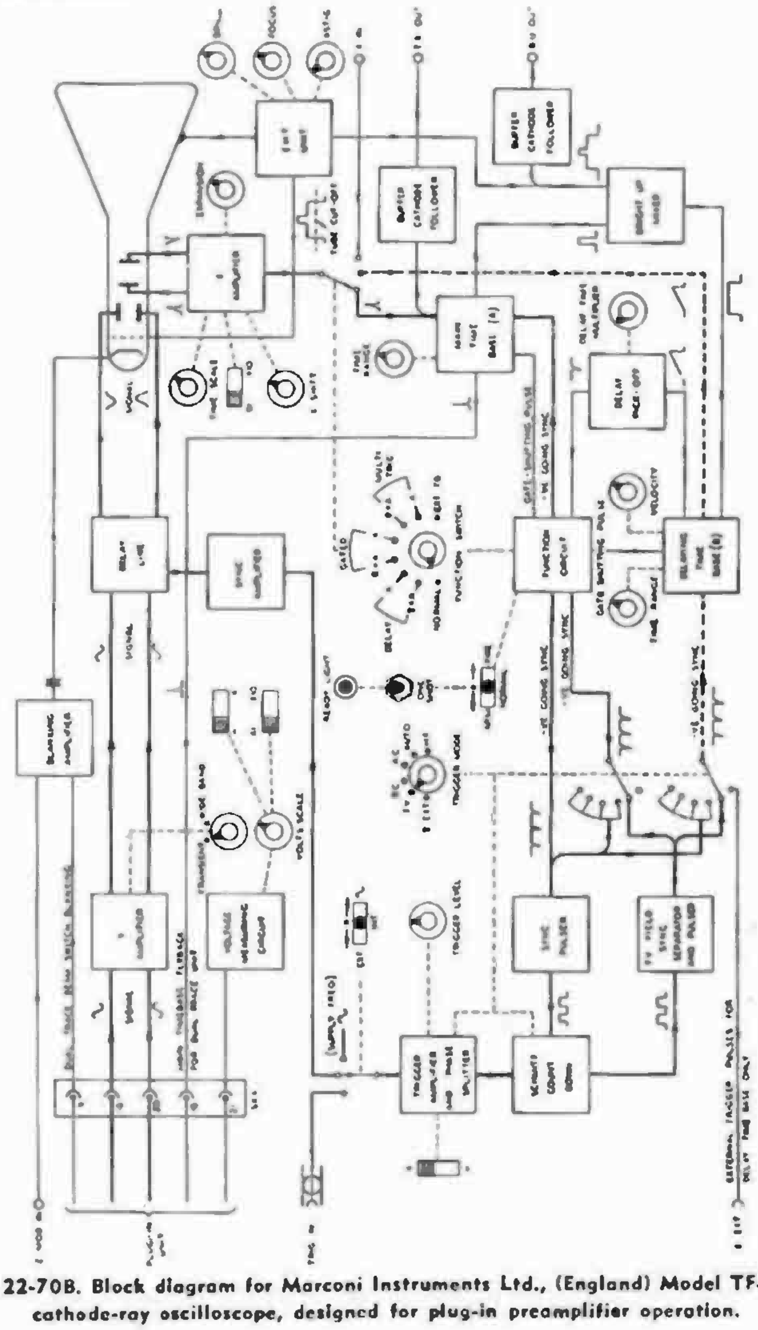


Fig. 22-70B. Block diagram for Marconi Instruments Ltd., (England) Model TF-2200 cathode-ray oscilloscope, designed for plug-in preamplifier operation.

Single trace, algebraic sum of A and B
Single trace, channel B only
Dual trace, chopped between A and B

Each input may be reversed separately and coupled for ac or dc operation. Isolation between inputs is 60 dB. Interior views of this instrument are given in Figs. 22-70C and D.

A Model 82 dual-trace plug-in preamplifier with test probes, manufactured by Tektronix, Inc., is pictured in Fig. 22-70E. A Model 564 oscilloscope using plug-in preamplifiers and timebase appears in Fig. 22-70F. This instrument is of the storage type in which waveforms may be superimposed on the screen and held up to 1 hour, permitting comparison of several waveforms without resorting to photography.

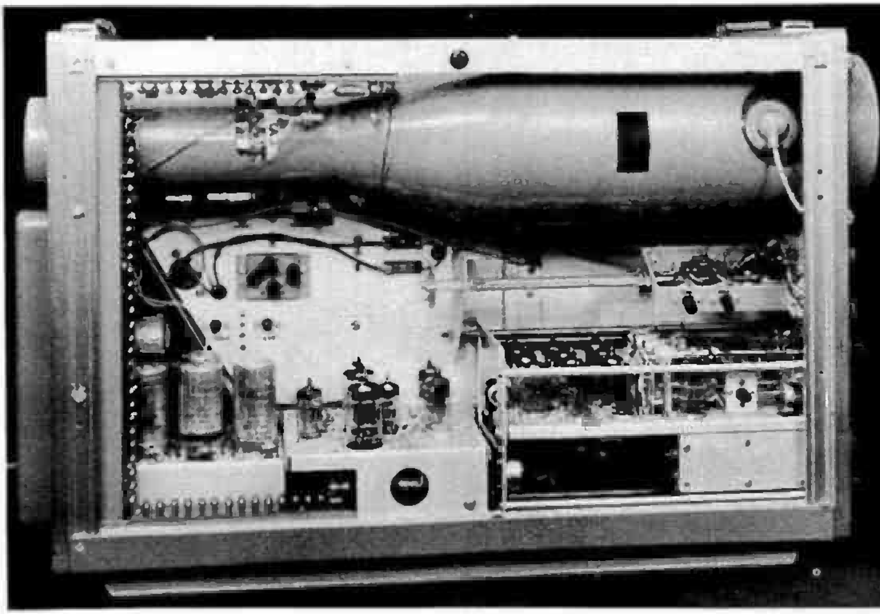


Fig. 22-70C. Left side interior view of Marconi Instruments Ltd. (England) Model TF-2200 oscilloscope. (Courtesy, Canadian Marconi Instrument Ltd.)

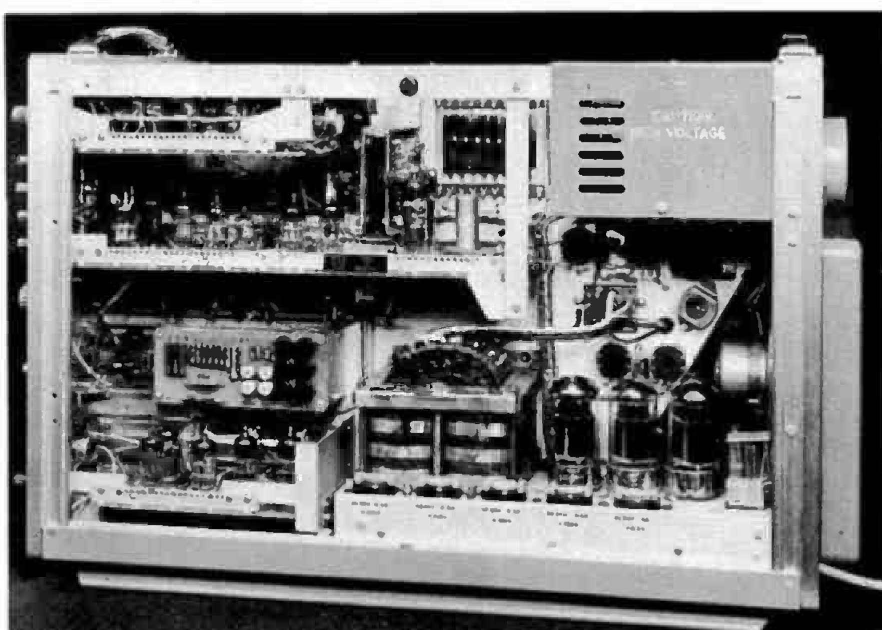


Fig. 22-70D. Right side interior view of Marconi Instruments Ltd. (England) Model TF-2200 oscilloscope. (Courtesy, Canadian Marconi Instrument Ltd.)

The display may be erased in 0.25 second. The instrument shown is using a differential amplifier and time base of given characteristics for observing waveforms generated by an impact machine.

22.71 ***Describe the circuitry for a 5-inch cathode-ray oscilloscope.***—A front-panel view of a Model KG-2000 Allied Radio Corp. 5-inch cathode-ray laboratory oscilloscope is given in Fig. 22-71A.

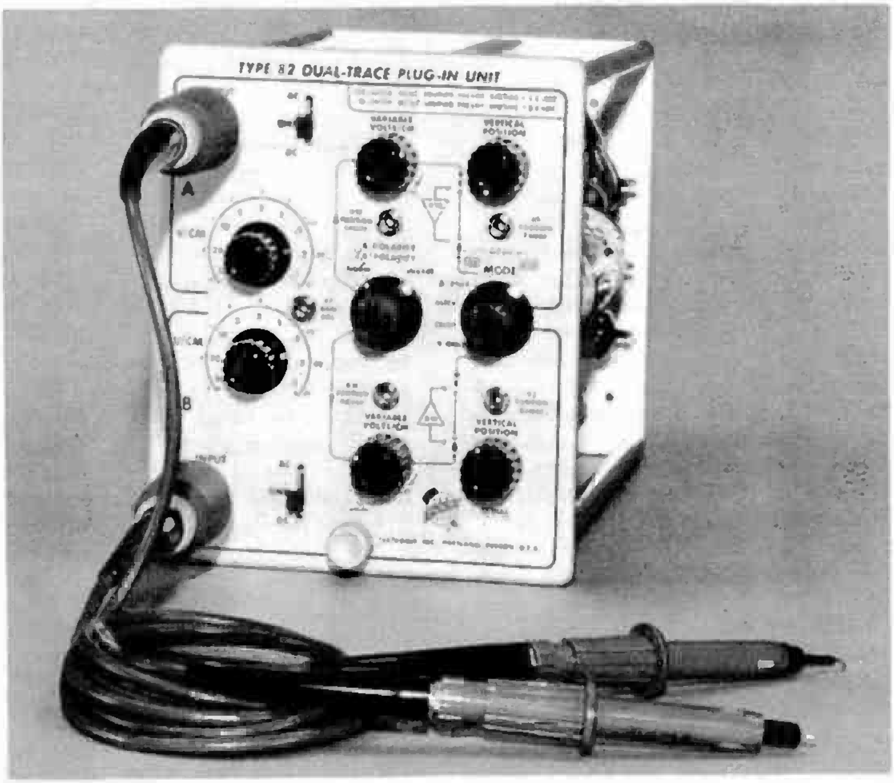


Fig. 22-70E. Tektronix Inc. Model 82 dual-trace plug-in preamplifier and test probes. With this unit both the input and output waveforms of a device under test may be displayed simultaneously.

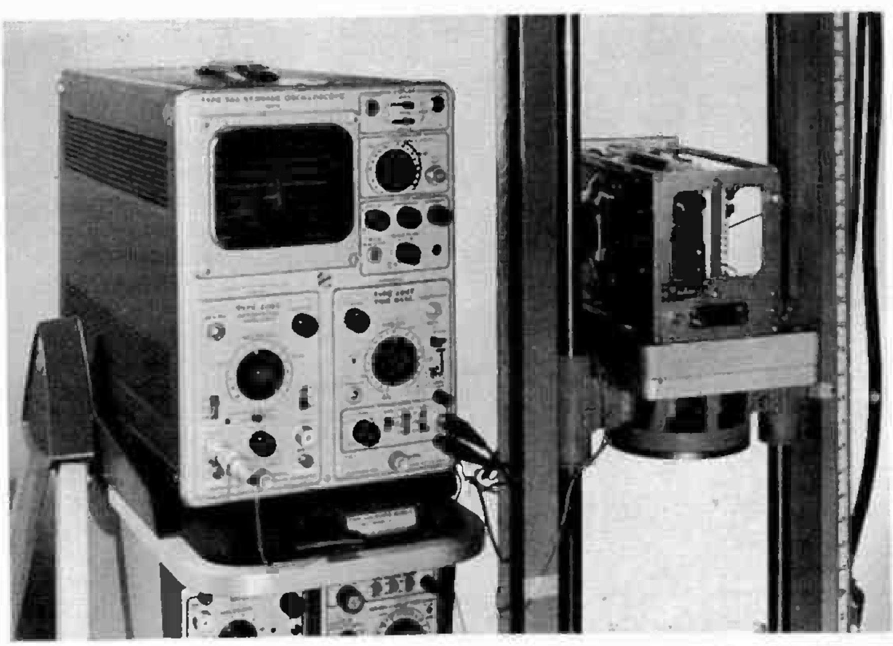


Fig. 22-70F. Tektronix Inc. Model 564 storage oscilloscope, using plug-in differential amplifier and time base.

Below the graticule are two positioning controls used for centering the cathode-ray tube (CRT) beam spot on the vertical and horizontal graticule reference lines. Below the vertical positioning control is a compensated vertical amplifier input attenuator, calibrated in volts per centimeter. This control is of coaxial construction (as are the others) with a variable gain control in the center. At the right, under the horizontal positioning control, is a time-per-centimeter sweep control, calibrated in milliseconds and microseconds per centimeter. The small knob of this assembly is a multiplier for the large knob. Below this control is a variable control operating in conjunction with the sweep control.

At the right of the panel are five coaxial-type controls. Starting at the top, the first control adjusts the beam of the CRT for astigmatism and turns on and off the power to the instrument. The second knob assembly adjusts the intensity and focus of the CRT beam. Below this is a trigger slope and preset control. The fourth assembly is a trigger input selector switch and trigger sensitivity control. The fifth and bottom knob assembly selects the horizontal input and provides gain control for an external signal.

Along the bottom edge of the panel is a male-type BNC coaxial fitting for applying the input signal to the vertical amplifier. To the right of the coax fitting is a switch for selecting either a dc or ac input circuit for the vertical amplifier. A terminal post supplies an external source of 0.10 volt pk-pk, 60 Hz for test purposes. The next three terminals are for the connection of an external capacitor in the sweep circuit for very slow sweep times, and an external signal input for the horizontal amplifier. The second coaxial fitting is for an external trigger input circuit.

Before starting a detailed description of the instrument, the general characteristics should be discussed. The wide frequency response to 5 MHz permits the display of pulses of fast rise time. Trigger and amplifier circuits are dc coupled throughout for application of very low frequencies or where dc levels must be displayed. Vertical displacement indicators above the CRT

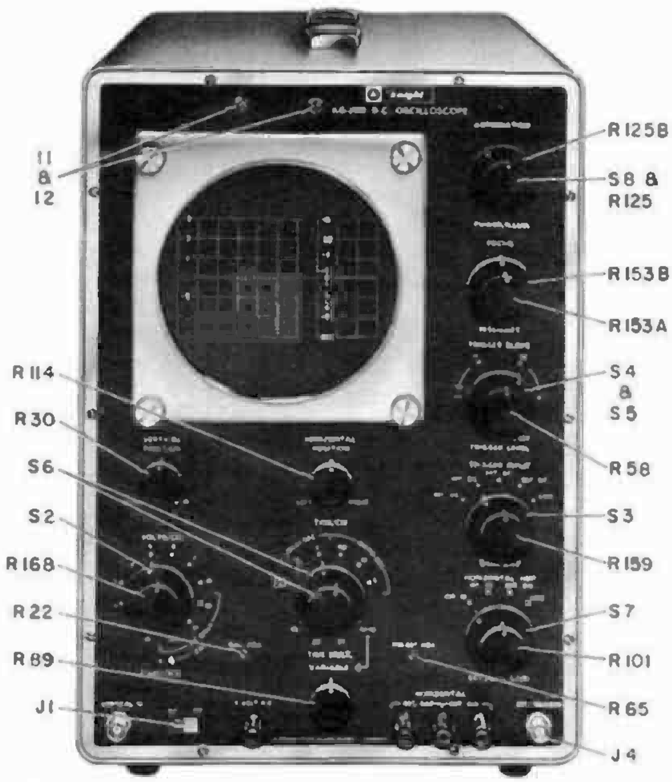


Fig. 22-71A. Allied Radio Corp. Model KG-2000 oscilloscope.

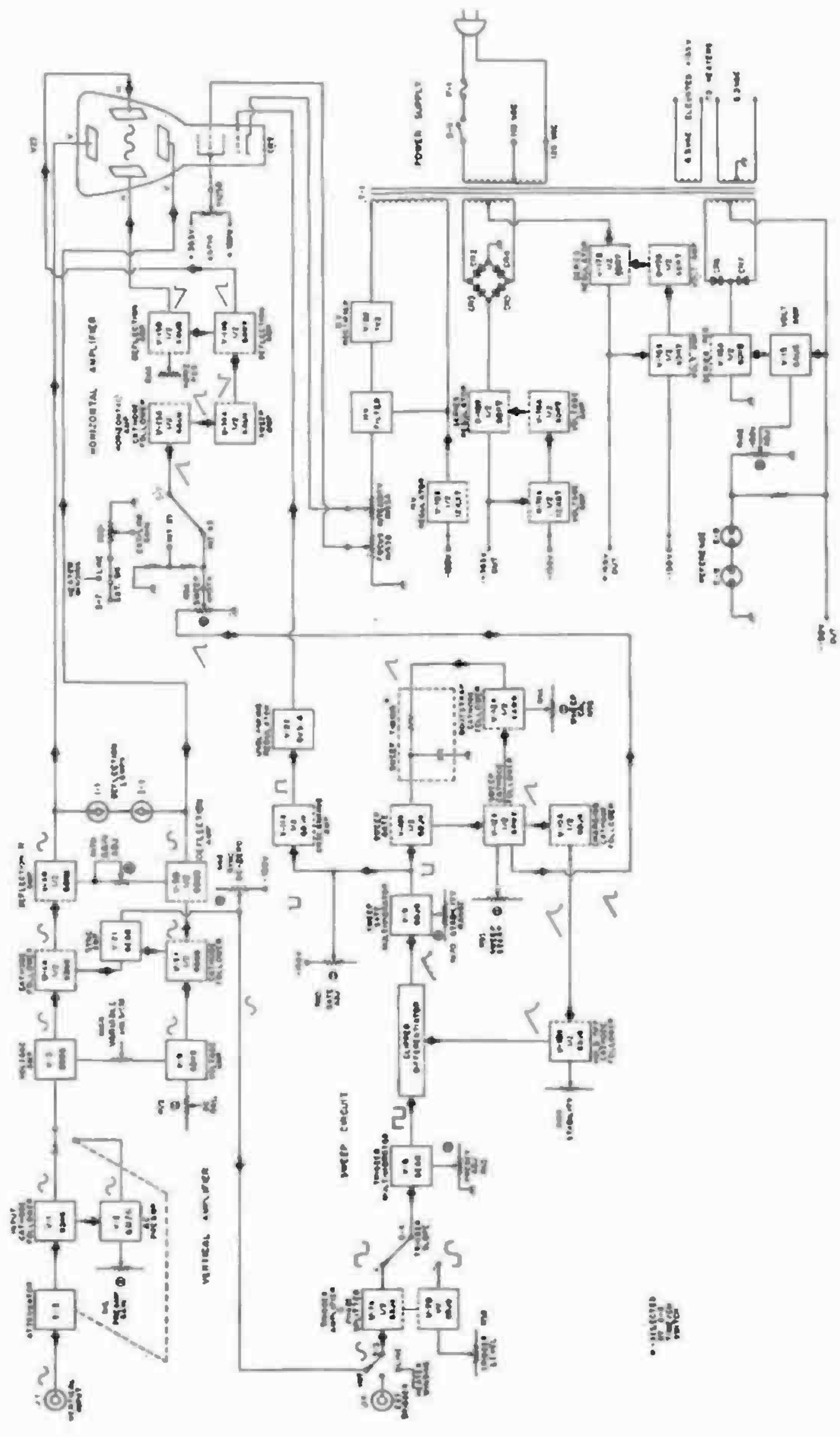


Fig. 22-71B. Block diagram for Allied Radio Corp. Model KG-2000 oscilloscope.

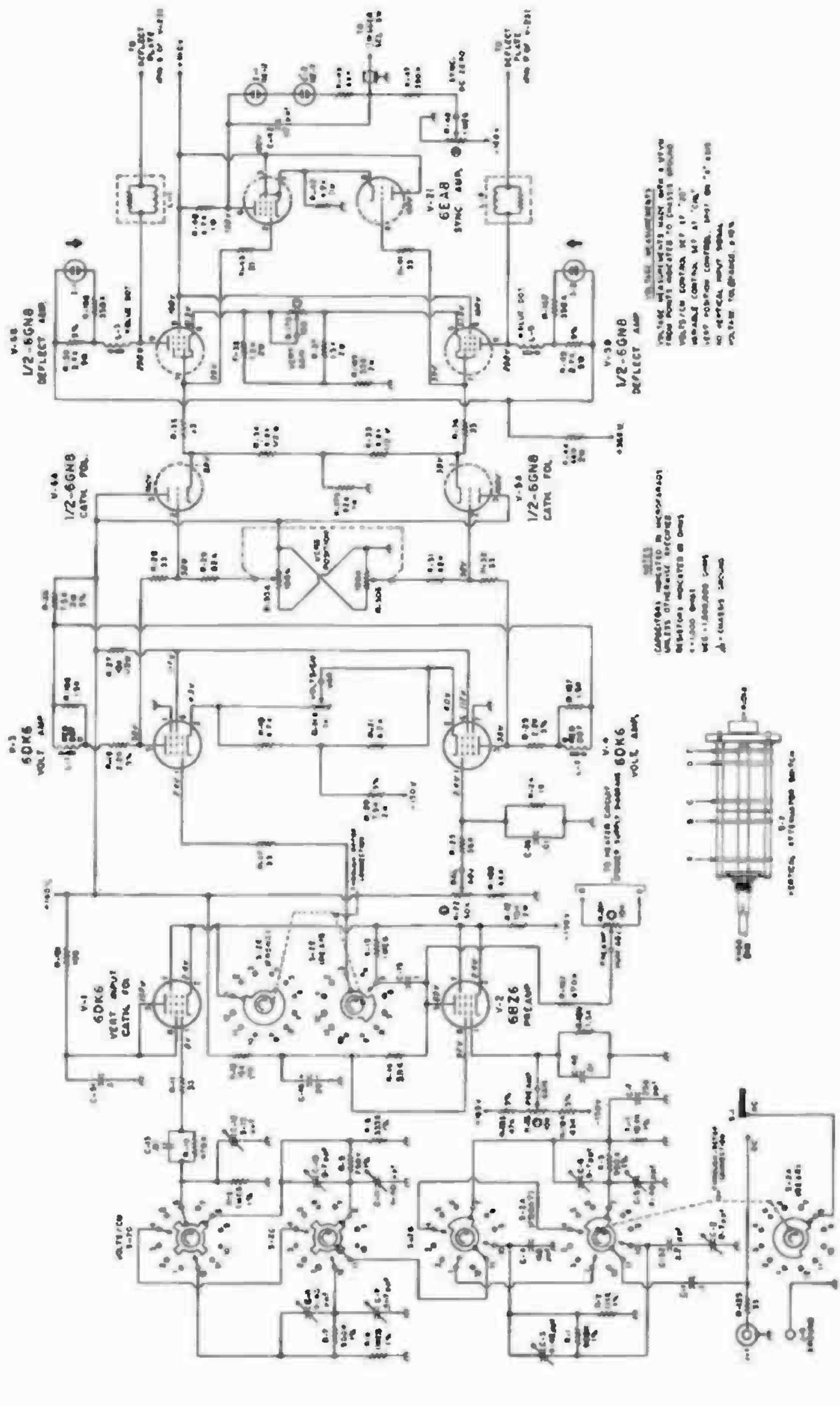


Fig. 22-71C. Schematic diagram for vertical amplifier section.

provide a quick check if the beam has been deflected off the scale. The horizontal displacement design is such that the image is never off scale to the left or right side of the graticule.

Preset lock-in of the vertical signal eliminates the need for synchronizing the sweep with input signals of varying amplitudes or frequencies. The preset feature may be disabled when required. A separate control permits locking-in of any portion of a signal, with a choice of either a positive or negative slope. The circuitry is that of a trigger-driven sweep. Only signal levels large enough to deflect the CRT beam ½ centimeter will permit triggering. The sweep speed is adjustable from 1 microsecond per centimeter to 50 milliseconds per centimeter, and can be increased beyond 100 milliseconds by the connection of an external capacitor greater than 0.10 μF. An expanding or magnifying circuit spreads the image five times to permit viewing minute portions of the display. Studs are provided at the graticule for mounting a standard oscilloscope camera.

To facilitate the description of this instrument, the front-panel view, schematic and block diagram are keyed with the same symbols. As the block diagram (Fig. 22-71B) is traced, the schematic diagrams should be referred to for clarification.

Referring to Fig. 22-71B, at the upper left the vertical amplifier system is direct coupled throughout for all vertical deflection sensitivities down to 0.05 volt/cm. The addition of an RC-coupled preamplifier stage further increases the sensitivity down to 0.005 volt per centimeter. The attenuator divides this sensitivity down in eight steps to 20 volts per centimeter. Preamplifier V2 adds a gain of 10 and three additional high-gain positions to the attenuator. Input attenuator S2 consists of four independent precision voltage dividers connected in various combinations to provide 12 steps of attenuation. A constant input impedance of 1 megohm shunted by 40 pF is maintained in all switch positions. This permits the use of a low-capacitance probe without sacrificing accuracy of calibration.

Tube V1, a 6DK6 pentode, functions as an input cathode follower. The low-impedance output of this stage is coupled directly to the grid of V3 in the 20 to 0.05 volt/cm positions of S2. In the three most sensitive positions of S2, V2 acts as a grounded-grid cathode-drive amplifier. The output is RC coupled from the plate of V2 to the grid of V3. The use of a grounded-grid preamplifier stage prevents an undesirable phase-inversion, when switching from 0.05- to 0.02-volt/cm sensitivity. To prevent accidental overload by a dc level, in the high-sensitivity position an input coupling capacitor is automatically switched into the circuit to override the ac/dc input switch S1.

Tubes V3 and V4 function as voltage amplifiers and phase inverters for driving the push-pull stages which follow. These tubes are cathode coupled, with V4 being operated as a grounded-grid stage. The cathode circuit employs a novel gain control. This continuously variable control is the volts/cm variable control (R168) on the front panel. The gain of V3 and V4 depends on the signal amplitude developed across their cathode resistors. With control R168 set to its minimum resistance, the calibrate position, the cathodes are connected directly together. This is the condition of maximum gain since all of the common cathode resistor signal is being applied to V4. R19 and R21 are in parallel and are simply adding to the value of R20. However, as the effective resistance of R168 is increased, R19 and R21 start to function as independent cathode resistors. This induces negative feedback into their respective stages. With resistor R22, the balance adjustment control properly set, the variable control will not shift the trace. The oppositely phased signals appearing at the plates of V3 and V4 are direct coupled to V6A and V5A respectively. Shunt peaking is used to preserve the wide bandpass characteristic. Positioning is provided by dual control R30. Tubes V6A and V5A functioning as cathode followers are directly coupled to the grids of the deflection amplifiers. Deflection amplifiers V5B and V6B employ low-resistance plate loads and shunt and series peaking, to extend the response well beyond 5 MHz.

Neon tubes I1 and I2, the vertical position indicators above the graticule, are connected across deflection-amplifier plate-load resistors R39 and R45. V21, the sync amplifier, functions as a differential amplifier. Signals from V6A and V5A, applied to the two control grids, yield an output at the plate of V21 only when they are out of phase. Common-mode signals such as hum or noise from the power supply are not amplified by V21. As a result, false triggering of the sweep current is eliminated. This stage is also direct coupled and adjustable for dc balance by R48. The output of this stage operates the trigger circuit in the internal ac and internal dc trigger modes. Neon lamps E1 and E2 are used as voltage translators to couple changes in voltage from one dc voltage level to another.

The horizontal amplifier (Fig. 22-71D) consists of two dual-purpose tubes, V13 and V14. Switch S7, the horizontal-input amplifier switch, selects the desired input to the amplifier. In the ×1 position, 20 percent of the sweep circuit sawtooth voltage is tapped off by a precision resistor voltage divider and applied to the amplifier input. With S7 in the internal ×5 position, all of the sawtooth amplitude is applied to the amplifier. Since this expands the entire sweep five times, the part seen on the screen appears to be magnified. Any portion of the sweep can be brought into the screen with the horizontal positioning control R89.

When S7 is placed in the external signal position, a signal is applied to coaxial fitting J4 and fed to the amplifier. In the line position a 60-Hz sinusoidal voltage from a heater winding is applied to the amplifier. In the external signal or line positions, the signal can be continuously attenuated by R101, the external line gain control. The horizontal amplifier is direct coupled throughout. Tube V13A is an input cathode follower and provides a high input impedance. Tube V14A amplifies the output of V13A and apples it to deflection amplifiers V14B and V13B. The deflection amplifiers also serve as a cathode-coupled phase inverter to provide the necessary phase reversal at the plate of V13B, with respect to the plate of V14B, to operate the CRT deflection plates push-pull.

The sweep circuits (Fig. 22-71D) provide accurately calibrated, linear, sweep voltages for the generation of a linear time base. Linear sawtooth generation is obtained by adding an incremental voltage to the charging voltage applied to the sweep-timing capacitor. This has the effect of extending the linear portion of the voltage-vs-time curve of the RC network. The charging voltage is supplied by the cathode of V12B, whose voltage rises with the sweep to maintain a constant voltage across the charging resistor, thereby supplying constant current to the timing capacitor.

For each sweep range the correct timing resistor and capacitor are selected by sections of S6. Provisions for the connection of an external capacitor greater than 0.1 μF make sweep speeds lower than 100 ms/cm possible. Resistors R89A and R89B (a two-gang variable control) provide sweep speeds other than those available at the step positions of the time/cm switch.

The sweep is controlled by a gate voltage generated by V9, a bistable cathode-coupled multivibrator. Negative pulses from the trigger multivibrator, V8, are applied to the grid of V9A. A negative pnlse cuts off V9A while turning on V9B to apply a gate of negative voltage to the grids of V11A (unblanking amplifier) and V11B (sweep gate tube). V11A and V11B cut off simultaneously so that the sweep begins at the same instant that a step of positive voltage from the plate of V11A unblanks the CRT.

Two neon-tube voltage translators E3 and E4 couple the multivibrator gate to V11A. The gate adjust control R80 sets the gate signal with reference to ground, but has little effect on the height of the gate. Resistor R80 is adjusted for proper gate operation, complete cutoff alternated with full saturation of V11. When V11B is cut off, the timing capacitor begins to charge. The charging current flows through V12A back to the power supply. V12A is driven by V12B, which in turn is driven by the sawtooth voltage developing at the plate of V11B. This completes the bootstrap feedback loop. Sweep voltage to drive the horizontal amplifier is taken off the cathode of V12B.

Potentiometer R96, the sweep calibrate control, varies the slope of the sweep sawtooth voltage by setting the charging voltage. Two neon tubes E5 and E6 provide a constant dc voltage, tapped by R96, to supply the desired charging voltage. Sweep start control R93 sets the dc level at which the sweep

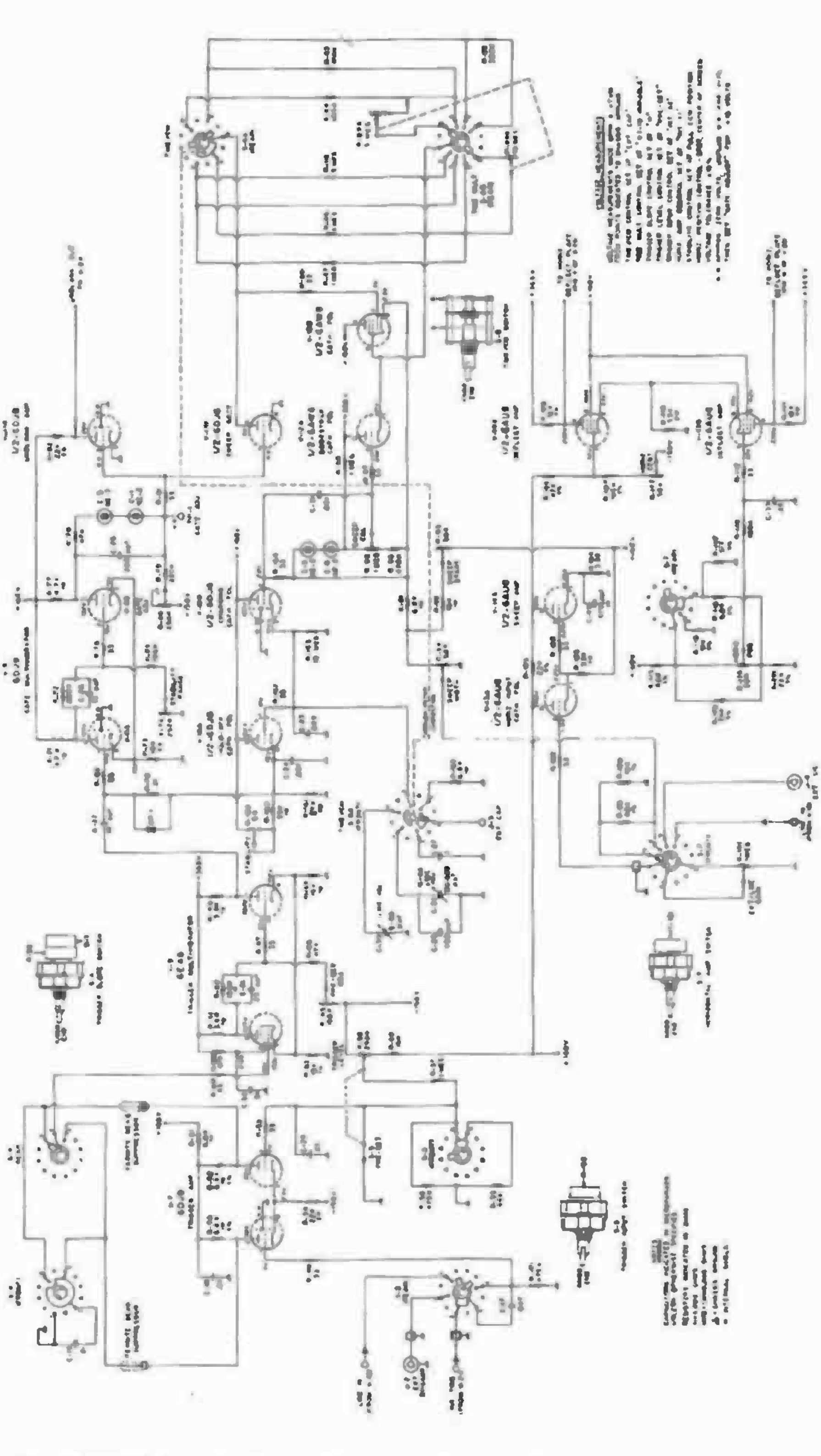


Fig. 22-71D. Schematic diagram for trigger circuits and horizontal amplifier section.

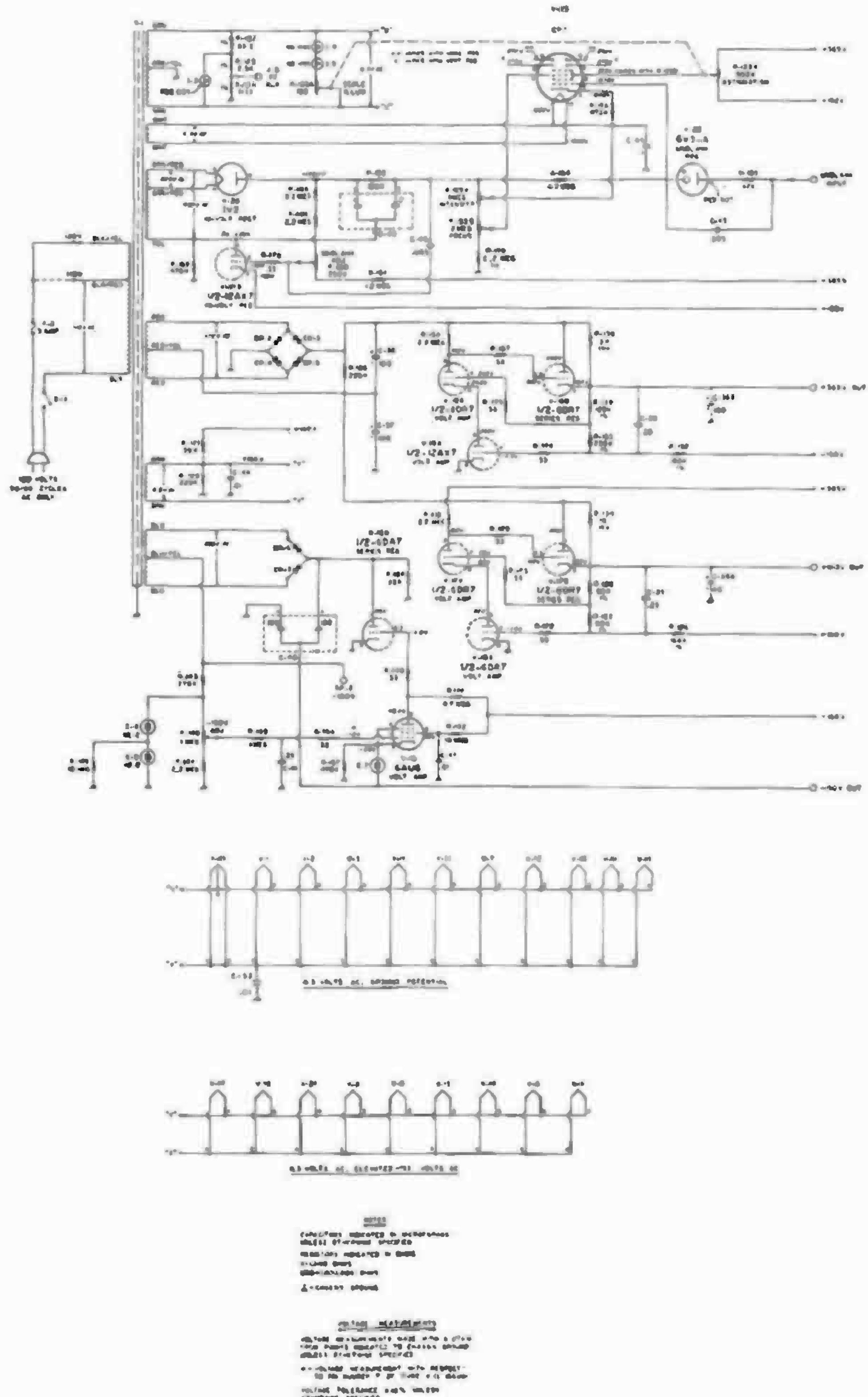


Fig. 22-71E. Schematic diagram for power supply section.

begins, fixing the undeflected spot at the left of the screen.

Part of the sweep sawtooth voltage is applied to V10B, the charging cathode follower. C25 and R163 form a fixed, parallel RC circuit in the cathode of this tube. The sawtooth voltage developed at V10B is coupled to the grid of the hold-off cathode follower, V10A. The voltage at the cathode of this tube is coupled to the grid of V9A. When this positive voltage reaches sufficient amplitude, V9 flips back to its original stable state with V9A conducting. The voltage at the plate of V9B instantly rises to B+ potential, saturating V11A and V11B. This blanks the CRT and causes the beam to almost instantly retrace. The timing capacitor quickly discharges through V11B. If it were not for the action of the hold-off circuit, another sweep would be initiated by the negative trigger pulses at the grid of V9A before the timing capacitor was completely discharged.

The hold-off circuit, V10A and V10B, prevents immediate decay of the positive voltage being applied to the grid of V9A at the end of the sweep. Instead, a parabolic decay shape is introduced at the end of the sawtooth by an RC network in the cathode of V10B. This gives sufficient time for complete discharge to the timing capacitor before the V9A grid voltage falls to the level below which a negative trigger pulse will flip the gate multivibrator back to its other stable state.

Potentiometers R159 and R74, the stability and stability range adjustments, set the operating levels of V9 for stable synchronization. If the stability control is advanced too far clockwise, V9 becomes astable and a free-running unsynchronized sweep occurs.

A synchronizing source for the sweep circuits is selected by switch S3, the trigger input switch. In the internal ac and internal dc positions a portion of the signal under study is used as a trigger source. This sync output is taken from sync amplifier V21 (Fig. 22-71C). In the external ac and external dc positions, the triggering source is introduced at external trigger coaxial jack J4. In the line position the line frequency feeding the instrument is used to trigger the sweep circuits.

The trigger signal is introduced directly at the grid of V7A in the two dc positions and through capacitor C17 in the two ac and line positions of S3. V7 is a cathode-coupled amplifier and phase inverter which provides a choice of trigger slope. In the plus (+) position of the trigger slope switch, the output is selected from the plate of V7A while the V7B plate is grounded through C19. Since only a negative-going output signal can trigger the next stage, V8, the sweep will start on the positive-going portion of the incoming signal. There is a 180-degree shift in the V7A. In V7B, however, there is no phase inversion because this stage is cathode driven. Therefore, in the minus (−) position of the trigger slope control with the output taken from V7B, the sweep starts on the negative slope of the incoming signal. Trigger level control R58 sets the dc operating levels of V7, thereby controlling the gain of this circuit.

Since the output of V7 is direct coupled to the trigger multivibrator tube V8, trigger level control R58 determines the exact point of the displayed waveform at which the sweep triggers. In the preset position, the trigger level has no effect, because switch S5 (on switch S4) closes, connecting the control grid of V7B directly to ground. Any signal above a minimum level will trigger the sweep when the trigger level control is in preset position.

The trigger multivibrator, V8, is a Schmitt trigger which employs a pentode for the input section for fast switching. This circuit produces square waves at the frequency of the incoming signal. The square wave is differentiated by C22 and R70, producing short-duration negative pulses (positive pulses are suppressed by diode CR1) for the purpose of triggering the gate multivibrator.

In the quiescent state, tube V8A is held in conduction by a positive control-grid bias, while V8B is cut off. When a negative signal is applied to the grid of V8A this section immediately cuts off and the output section conducts heavily, generating a negative step. V8A remains cut off until the grid of this tube is driven positive again by the incoming signal. The circuit immediately switches back to its original state, awaiting the next negative-going trigger signal.

The power supply (Fig. 22-71E) provides four basic potentials—minus 150 volts, plus 160 volts, plus 365 volts, minus 1900 volts. All of these potentials are regulated. The minus 150-volt supply is extra well regulated and is used as a reference voltage for the other three supplies. Heater power is supplied by two windings on power transformer T1. One winding is center-tapped and grounded, while the other winding is biased at plus 135 volts to prevent excessive heater-to-cathode potentials on the tubes.

An accurate 0.1-volt peak-to-peak reference voltage is also supplied at the front panel. A half-bridge circuit employing a selected No. 1490 bulb and a precision resistor voltage divider provides good regulation of this reference voltage.

The minus 150-volt supply consists of a full-wave silicon-diode rectifier circuit and a series regulator plus a pentode amplifier circuit. Series regulation takes place between the positive side of the supply and ground. Tube V16A, the power triode, is a series regulator driven by a "starved-current" pentode V15. Reference voltage for this error amplifier is created by two neon lamps E8 and E9. A portion of this constant voltage is tapped off by R146, the minus 150-volt adjust control. The amplifier is cathode driven through a constant-voltage-dropping neon lamp E7. Thus any tendency for the minus 150-volt potential to change is coupled through E7, without attenuation, to the cathode of V15. Since the grid of this tube is held constant, the change in cathode voltage constitutes an input voltage. This change is amplified and applied to the grid of V16A, which in turn increases or decreases the conduction of the tube to return the potential to minus 150 volts.

The plus 160-volt supply consists of a full-wave silicon-diode rectifier circuit and a series regulator with a cascade error-amplifier circuit. The cascade amplifier V16B and V19A provides gain and stability. Reference voltage from the minus 150-volt line is applied to the grid of V16B through R126 and R172. Error voltage is applied to the grid of V17A from R127 and R128, a precision voltage divider. The amplified error voltage appearing at the plate of V17A is coupled through R129, a parasitic-suppressor resistor, to the grid of the series regulator V17B. To increase the current capacity of the supply, R130 is in parallel with V17B from cathode to plate.

The plus 365-volt supply consists of a full-wave bridge rectifier circuit and a series regulator with a cascade amplifier. Two of the silicon diodes in the bridge circuit, CR2 and CR4, are shared with the plus 160-volt supply. Operation of this regulator is the same as in the 160-volt circuit. Resistor B139 reduces the regulator-tube current requirements by shunting current around V18B.

As in most oscilloscopes, the high voltage required to operate the CRT is negative with respect to ground, because the vertical and horizontal deflection plates are 200 volts above gronnd. This is the result of the direct conpling employed. This negative potential is regulated against both load and line voltage changes. Unregulated dc voltage is developed by a half-wave rectifier V20 and an RC filter network. The bottom of ground end of the transformer winding that supplies the high-voltage ac is connected to the plate of V19B. The cathode of V19B is connected to the minus 150-volt reference voltage.

A voltage divider composed of R148, R149, unblanking adjustment control R150, and R151 feeds an error voltage to the grid of V19B. Thus, regulation takes place by moving the whole high-voltage winding of T1 up or down, with respect to ground.

Focus and CRT grid-to-cathode bias voltage is obtained from a second high-voltage divider. This divider is made up of R153A (the intensity control), R153B (the focus control), and R158. With the proper setting of intensity control R153A, the CRT beam is just cut off during the absence of sweep. When sweep occurs, a positive voltage taken from the plate of unblanking amplifier (V11A) is applied to the grid of the CRT through V22. Tube V22, a 1750-volt corona regulator, couples the change in voltage at the plate of V11A (at about 160 volts) to the grid of the CRT (at about 1600 volts) without attenuation. Thus V22 functions like the neon lamps used elsewhere, to translate a voltage change at one voltage level to the same voltage change at another voltage level.

22.72 *Give the circuit for a high-gain differential plug-in oscilloscope amplifier.* — The primary purpose in using a differential amplifier with an oscilloscope is to eliminate undesirable common-mode signals. The term "common-mode signal" is defined as that signal which is common to both inputs of the differential amplifier. Usually, but not necessarily, one signal is an unwanted hum voltage. This mode of operation can be used, for example, to observe the signal across one circuit element while effectively eliminating the remainder of the circuit from the observation. This is accomplished by connecting one signal to input A of the differential amplifier and the second signal to input B. Differential or common-mode rejection ratio is a function of frequency in practical amplifiers. Depending on the design of the amplifier, this ratio may be 5000:1 or even as great as 20,000:1 for dc common-mode signals, and it will remain near these values throughout the audio-frequency band, decreasing as the frequency increases.

As an example, consider two signals, one of 10,000 Hz and one of 60 Hz, are intermixed, as shown by the upper trace of Fig. 22-74A. It can be seen that such a signal becomes quite confusing when displayed. A differential amplifier may be used to an advantage to reject the unwanted 60-Hz signal and permit a satisfactory display of the 10,000-Hz signal, as evidenced by the lower trace of Fig. 22-72A. Using the amplifier to be described, this is accomplished by the use of an external filter trap to block the desired signal from one input while the mixed signal is fed to the other input (Fig. 22-72B). Since the unwanted 60 Hz is common to both inputs, it is greatly attenuated by the difference amplifier, while the 10,000-Hz signal is relatively unaffected since it is fed to one channel only. If probes are being used, they must be matched quite closely; if their internal resistance is not the same, the common-

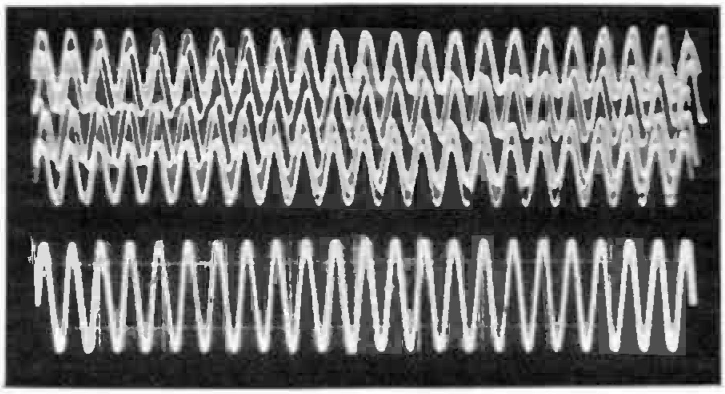


Fig. 22-72A. Top trace shows the appearance of a 10,000-Hz signal intermixed with 60-Hz hum. The lower trace shows the 10,000-Hz signal after passing through a differential amplifier.

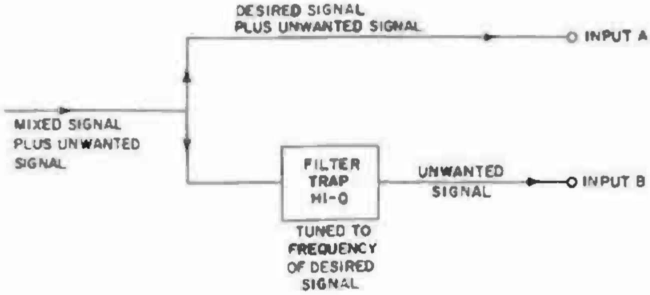


Fig. 22-72B. Tuned-filter trap in the external circuitry of a differential amplifier to block the desired signal from being received by one input while the mixed signal is fed to the other input.

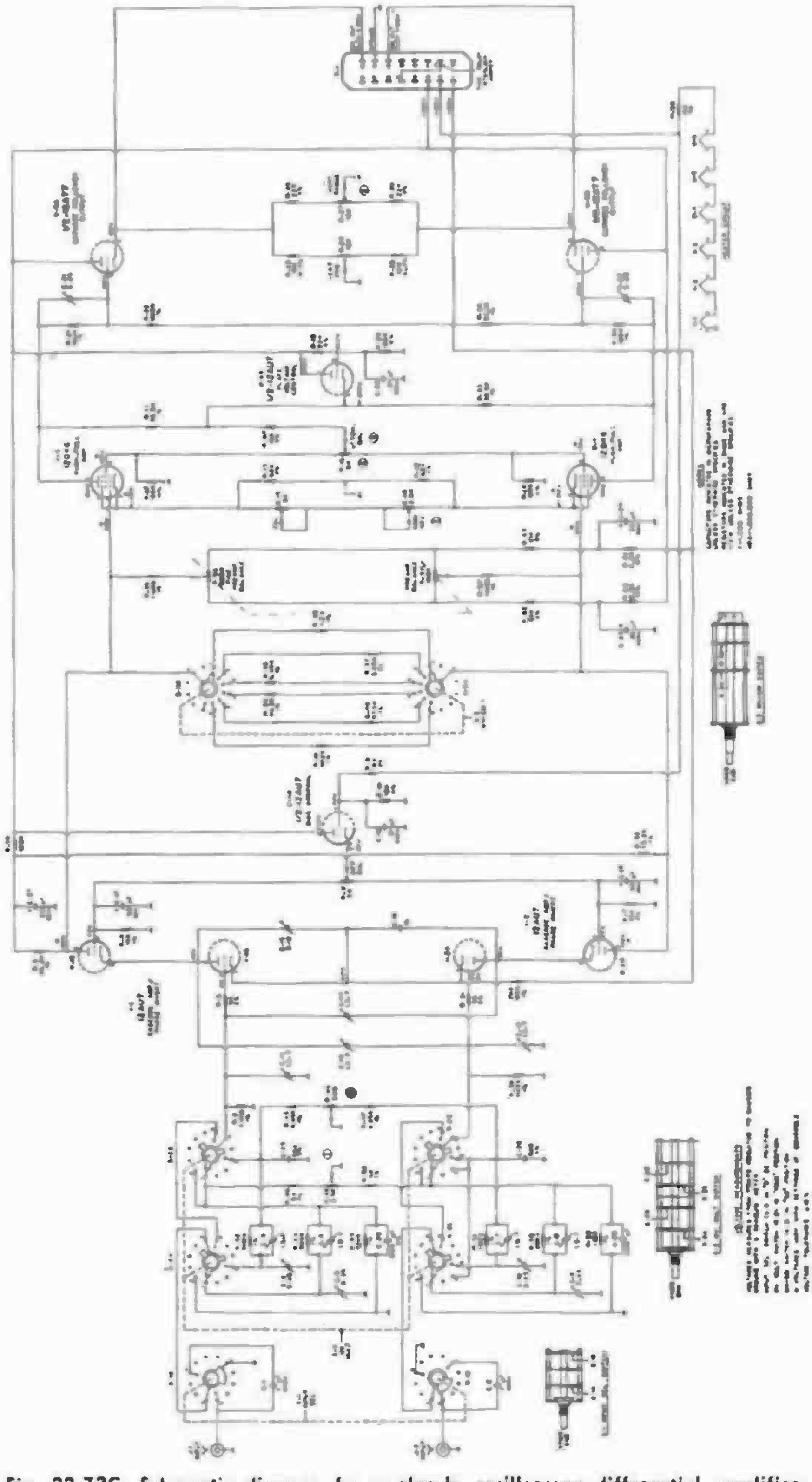


Fig. 22-72C. Schematic diagram for a plug-in oscilloscope differential amplifier. (Courtesy, Allied Radio Corp.)

mode rejection ratio will suffer because of the greater attenuation of one signal over the other.

The circuitry for a high-gain differential plug-in preamplifier is given in Fig. 22-72C. This amplifier is designed to be used with a basic oscilloscope designed for plug-in amplifier use. The gain of the circuit is adjustable over a range of 1 millivolt to 300 volts in 15 fixed steps, operating in conjunction with a continuously variable gain control. Two modes of operation are available, either straight single-channel amplification or as a differential amplifier with a two-channel input and a single-channel differential output. A six-position selector switch provides ac or dc input coupling, input A or input B separately, or both inputs, for differential output. The selector switch positions are:

Dc coupling	input A only
Dc coupling	input B only
Dc coupling differential mode	
Ac coupling	input A only
Ac coupling	input B only
Ac coupling differential mode.	

Dc coupling is used between all stages of the amplifier, with a choice of ac or dc coupling at the inputs. In the ac positions of the input selector switch S1, C1 and C2 block the dc components of any signal at inputs A and B. In the dc position of S1, these capacitors are shorted out, allowing dc information and very low-frequency signals to pass to input tube V1. When input A is selected (either ac or dc), input B is grounded, and vice versa. In the differential positions both inputs are open and signals at input A and input B are fed into the amplifier.

Since channels A and B are identical in operation, only channel B will be described. As shown in the schematic diagram, when input selector switch S1 is in the B-dc position, input A is grounded by switch S1A grounding the grid of V1A. A signal introduced at input B is brought directly through S1B to S2, the millivolt/multiplier switch. In the $\times 1$ position, the signal goes through to R31 and to the grid of V2A, a 12AU7 connected as a cascode amplifier. In the $\times 100$ position, before the signal is applied to the grid of V2A, it is first attenuated by a voltage divider, R51, R47, R46. Capacitors C7 and C10 provide high-frequency compensation. Similarly, in the $\times 1000$ position, the signal is attenuated by R53, R50, R49, and R52, and compensated by C8, C9, and C28.

Channel B input is applied to V2, a 12AU7 with its two halves connected in cascode to provide high gain with low internal noise. Input signals at the grid of V2A appear at the plate of V2A, amplified and 180 degrees out of phase with the input. Since theV2A plate is tied to the cathode of V2B, the signal is further amplified in V2B. However, no phase change takes place in V2B which functions as a grounded-grid amplifier (grid at ac ground.) Therefore, at the plate of V2B the signal is 180 degrees out of phase with the input of V2A.

Tube V1, another 12AU7, is operated in a manner similar to V2, but the grid of V1A is grounded because S1, the input selector, is in the B-dc position. A signal is introduced to the cathode of V1A from the cathode of V2A, in phase with the input to V2A. No phase inversion takes place in V1A or V1B, since both function as grounded-grid amplifiers. Consequently, the output at the plate of V1B is in phase with the input signal, whereas the output as the plate of V2B is out of phase with the input signal. The gain of V1 and V2 is balanced within a few percent because of the relatively high value of the common cathode resistor, R4.

Tubes V1 and V2 are direct coupled to V3 (12DK6) and V4 (12DK6) push-pull amplifiers. Since the output of the amplifier must be at a dc level of plus 100 volts to supply the correct bias for the input tubes of the main vertical amplifier, V3 and V4 outputs are dropped across frequency-compensated attenuators R21, R22, C21, R34, R35, and C22 and applied to the control grids of the output stage. Output stage V5 (12AU7) is a dual triode connected as cathode followers to deliver the signal to the vertical amplifiers in the main frame. V6, another 12AU7, functions as a voltage control to supply the 50 volts required to bias the grids of V1B and V2B, and the 200 volts for the V3 and V4 plates.

The mV/cm switch, S3B, provides a choice of six precision resistors which change the plate-to-plate feedback and consequently the gain of V1B and V2B.

S3B when used together with S2 provides 15 different calibrated sensitivities.

Since the balance of the two channels is extremely important, especially for the differential function, controls are provided to balance each stage. Resistor R46 balances the attenuation in the 100 position of mV/mult; in the 1000 position R49 provides the balance. Resistor R8, the front-panel differential balance control, equalizes the gain of V1 and V2 by controlling the dc grid voltage of V1B and V2B. Tubes V1 and V2 are also balanced by adjusting balance controls R58 (coarse) and R56 (fine.) These controls set the dc voltage for their respective tubes.

Front-panel calibrate control R15 and gain adjust control R14 are in series between the cathodes of V3 and V4, and vary the gain of this stage because the two cathodes are out of phase and coupling between then is degenerative. The gain adjust control sets the proper range for the calibration control. In the fully clockwise position of this calibration control, the calibrated sensitivities are accurate. Rotating the calibration control counterclockwise provides continuously variable sensitivities between the calibrated values. If the cathodes of the V3 and V4 are not in exact dc balance, the trace will shift vertically when the calibration control is turned. To prevent this effect R16, the attenuation balance control is used to vary the relative screen voltages of V3 and V4, thus controlling the dc currents through these tubes to bring the cathode voltages into balance. Since there is no dc drop across R14 or R15 the trace will not shift when the calibration control is used.

Vertical position control R24 (front panel) and vertical range R27, a screwdriver control, vary the dc output voltage at V5A and V5B to produce a vertical shift trace. Resistor R27 sets the range of R24 to permit vertical centering when R24 is in the center of its rotation.

In the differential mode position of the input selector switch, the signal from input A is applied to the control grid of V1A, while the signal from input B is applied to the control grid of V2A. Since the cathodes of these two tubes are tied together, the signals from inputs A and B appear at both cathodes. Therefore, V1A and V2A act as difference amplifiers, amplifying the difference between the grid and cathode, which in this instance is the difference between A and B input signals. In practice, since there is a slight difference in amplitude between the signal applied to the control grid of one tube and the cathode of the other tube, a small amount of the common signal will get through to the second stage.

Tubes V3 and V4 act as a push-pull amplifier for the desired difference signal. This provides further elimiuation of any common signal present because the common signal is applied to both grids in phase and therefore it cancels out in push-pull operation. The push-pull amplifiers in the main frame provide further elimination of the common signal.

Plug P1 at the extreme right of the diagram is attached to the rear of the plug-in amplifier chassis and mates with a female receptacle for supplying power and picking up the output circuits in the main frame. Although the amplifiers discussed may not apply to all oscilloscope main frames, the basic principles of operation are the same.

22.73 Describe the circuitry for a dual-trace plug-in oscilloscope amplifier. —Dual-trace oscilloscope preamplifiers permit the simultaneous observation and display of two waveforms. The waveforms may be the input and output of a device under test or two separate displays of waveforms in a system. The waveforms may be viewed separately or superimposed for comparison. Multiple-trace preamplifiers are not limited entirely to the viewing of two waveforms, but may be obtained to display up to four (or more) separate waveforms simultaneously.

The dual-trace unit to be discussed has a bandwidth of dc to 10 MHz (down 3 dB at 10 MHz. The input impedance is 1 megohm shunted by 40 pF and is constant; the rise time is 40 nanoseconds. The voltage range is 0.05 volt to 20 volts per centimeter in nine steps and operated in conjunction with a continuously variable gain control.

Referring to the schematic diagram of Fig. 22-73, a signal connected to input A is coupled to attenuator switch S1, the polarity switch. The position of this switch determines whether the signal is ac or dc coupled to the input

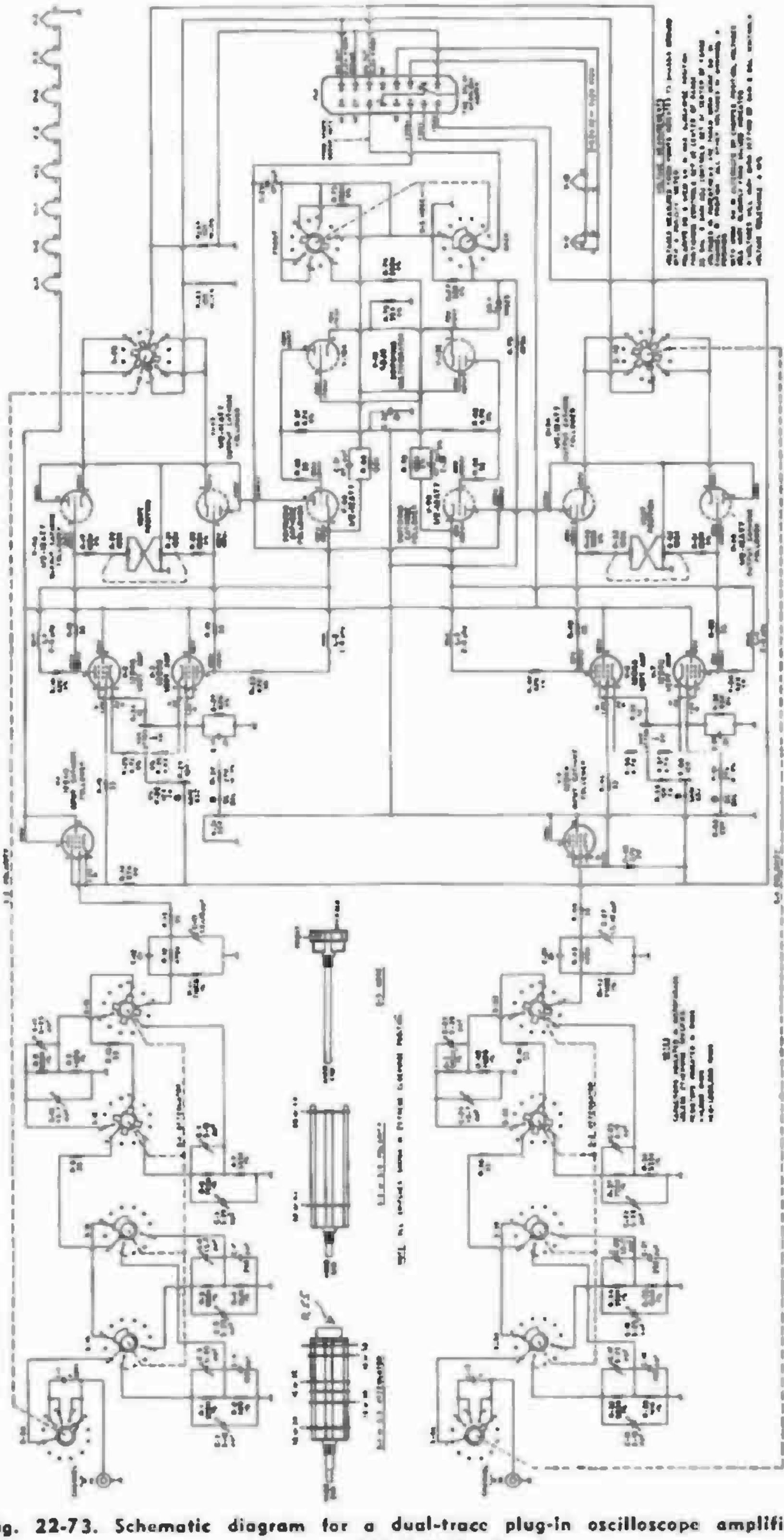


Fig. 22-73. Schematic diagram for a dual-trace plug-in oscilloscope amplifier. (Courtesy, Allied Radio Corp.)

attenuator network. In the ac position tne signal is passed through a 0.10-μF capacitor to hlock any dc component in the input signal. Switch S1 is also used to attenuate the input signal.

From the attenuator switch the signal passes to the grid of the first amplifier stage, V1, connected as a cathode follower to take advantage of the low input capacitances inherent with low-gain amplifier stages.

Tubes V2 and V3 form a cathode-conpled paraphase amplifier. The signal from V1 is impressed across the grid of V2 and the cathode of V3. At V2 the signal is amplified in the conventional manner, accompanied by the usual phase shift. Tube V3, however, being a grounded-grid amplifier, does not introduce a phase shift in the amplified signal. Consequently, the outputs of V2 and V3 are 180 degrees out of phase, or in push-pull. From here the signal is fed to the grids of V4A and V4B. Tube V4 is a dual triode, with cathode follower outputs to polarity switch S2. From this point the signal travels to the main vertical amplifier in the oscilloscope main frame, through connector P1.

The dual-trace feature is accomplished by the combined action of V9 and V10. With the mode switch in the alternate position, V10 acts as a bistable plate-coupled multivibrator. Due to the inherent imbalance in multisection tubes, one half of V10 will start to draw current (V10A for example), causing its plate voltage to drop. This lower plate voltage is reflected at the grid of V9A as an increase in negative bias, causing V9A to conduct less. Because of the reduced voltage at the cathode of V9A, the plate voltage of V2 and V3 is decreased to the point where these tubes will not pass the signal from channel A.

The negative-going cathode of V9A also causes an increase of negative bias at the grid of V10B, causing this section to conduct less and thus increasing its plate voltage. As the plate of V10B becomes more positive, it reduces the bias at the grid of V9B, causing this section to conduct more with a consequent increase of cathode voltage.

As the cathode voltage of V9B increases, the plate voltage at V6 and V7 will be increased proportionally, causing these tubes to pass the signal from channel B. The positive-going cathode of V9B reduces the negative bias at the grid of V10A, causing V10A to conduct more. This action tends to reinforce itself and remains in a stable condition until a positive pulse from the sweep gate of the scope reverses the action.

The positive pulse from the sweep gate (main frame) arrives at the cathodes of V10A and V10B through diode CR1. The additional positive charge on the cathode of V10A is reflected as an increase in negative bias on the grid of this section, causing it to become less conductive, with a consequent increase in plate voltage. This increase in plate voltage reduces the negative bias on V9A, causing it to condnct, which raises the plate voltage of V2 and V3. These tubes then pass the signal from channel A.

The positive-going cathode of V9A reduces the bias of V10B, causing this section to conduct. As the plate voltage at V10B is reduced, the negative bias at V9B is increased, causing V9B to conduct less. Because of the reduced cathode voltage at V9B, V6 and V7 will not pass the signal from channel B. The multivibrator will remain in this second state of stability until another pulse from the sweep gate upsets the condition, causing the switching cycle to repeat itself. The rate at which the multivibrator switches from one channel to the other is governed by the sweep rate of the scope. With the mode switch in the chopped position, V9 and V10 form a free-running multivibrator with a switching frequency of 100 kHz. The switching frequency is determined by the time constant of RC combinations C31-R69 and C30-R70.

Because grid-return resistors R73 and R74 have been switched ont of the circuit, the RC combinations become active. Capacitors C30 and C31 alternately charge and discharge at their respective sections of V10, alternating the voltages at the different tnbe sections in a manner identical to that explained for the alternate position. With the mode switch S3 in the alternate position, the two waveforms are displayed simultaneously from both channels and at a high sweep rate. In the chopped position the waveforms are also presented simultaneously, except at a lower sweep rate. By setting the switch to the A or B positions, only the waveform from the indicated position is presented. Position controls R2 and R53 separate the waveforms or superimpose them for comparison. Their positions on the CRT screen may also be interchanged vertically by these controls. The polarity of either the ac or dc inputs may be reversed 180 degrees by switches S2 or S4. The necessary operating voltages and output circuits are transferred through plug P1, which mates with the similar plug in the oscilloscope main frame. The input circuits may be directly connected to the device under test, or by use of the probes. A dual-trace plug-in amplifier and oscilloscope appears in Fig. 22-70E and 22-70F.

Although the preceding dual-trace amplifier may not be suitable for all oscilloscope main frames, the basic principles of operation are the same.

22.74 ***What is the rise time of an oscilloscope?***—It is the time it takes for a square-wave impulse to rise from 10 percent to 90 percent of its final amplitude, as displayed on an oscilloscope screen. In theory, a square wave is composed of an infinite number of harmonics and is of infinite bandwidth. In practice, neither an infinite number of harmonics nor infinite bandwidth is achieved. As a result, there is always some rounding of the so-called square corners of the waveform. To display the true waveform of a square wave generator, the bandwidth of the oscilloscope used for the measurement must be considerably greater than the bandwidth of the generator and the circuit under test; otherwise, only the bandwidth of the oscilloscope is indicated. A perfect square wave and one with rounded corners (dashed lines) are shown in Fig. 22-74A, with Fig.22-74B showing how rise time is measured.

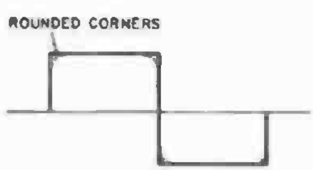


Fig. 22-74A. Perfect square wave with the rounded corners of a practical square wave as shown.

To measure rise time, the sweep timing controls on the oscilloscope are set for 0.5-μs/cm division, with the image spread 5 centimeters vertically. After the image is stabilized, a line is drawn upward from the 10-percent point, and another drawn downward from the 90-percent point. The rise time in microseconds is then the difference between these two lines. The example shown indicates a rise time of 0.03 microseconds, or 30 nanoseconds.

Both the bandwidth of an oscilloscope and its high-frequency cutoff point have a direct bearing on the rise time. It is standard practice to rate the bandwidth of an oscilloscope at a point where the frequency response is down

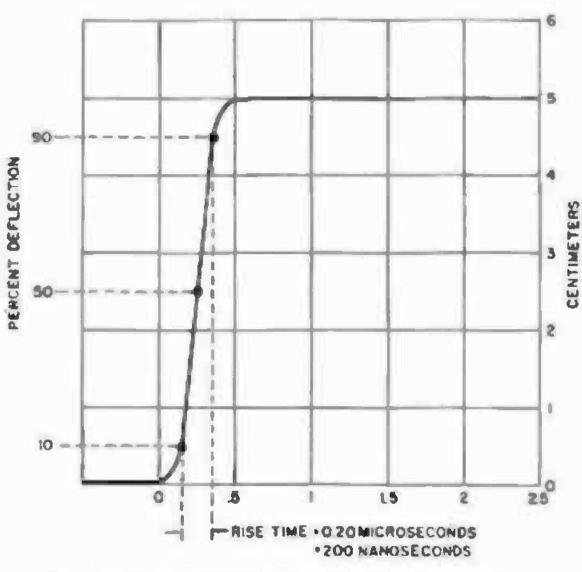


Fig. 22-74B. Typical rise-time measurement.

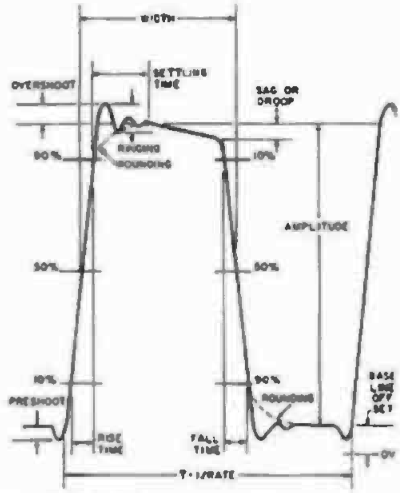


Fig. 22-74C. Square waveform with overshoot, ringing, and droop, and the terminology used in measuring rise time.

3 dB with respect to a reference frequency, or about 30 percent. If a square-wave generator is not available, the rise time may be approximated fairly easy, as rise time is the reciprocal of frequency, or 1/f. Therefore, if an oscilloscope is rated 3 dB down at 5 MHz, the rise time is approximately

$$t = \frac{1}{5{,}000{,}000} = 0.00000002 \text{ second,}$$

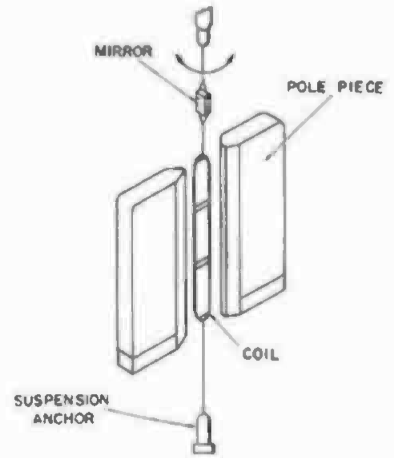
$$\text{or } 0.20\ \mu s$$

$$\frac{0.20}{3} = 0.065\ \mu s = 65 \text{ ns.}$$

Many times a square wave, after passing through a piece of equipment, will indicate ringing, overshooting, or rounding. The terminology associated with waveforms of this nature is given in Fig. 22-74C. The waveform appearance of the square wave should be monitored each time the frequency is changed. (See Question 25.106.)

22.75 ***What is an oscillograph?***—It is an electromechanical transducer employing a small mirror (0.10-inch square) mounted in a loop of wire. This assembly is placed in a strong, permanent magnetic field (Fig. 22-75A). The signal current flowing through the wires causes them to react in the magnetic field, resulting in a rotating motion proportional to the current through the wires. A pinpoint of light is projected on the surfaces of the mirror and reflected to a scale or photosensitive paper.

Recording galvanometers used in the original RCA variable-area sound-on-film motion-picture recording systems (1928) used the principle of the Dudell string oscillograph with certain modifications (Fig. 22-75B). Here, the mirror is mounted directly on a pair of wires in the form of a hairpin. The assembly was placed in a magnetic field and a light projected onto the surface of the mirror and reflected back through an



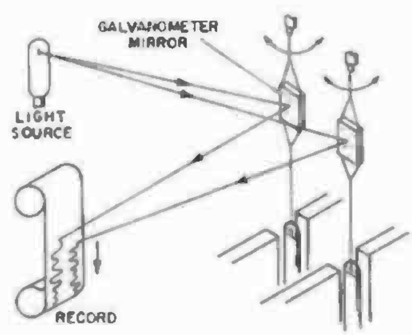


Fig. 22-75A. Galvanometer movements used in an oscillograph.

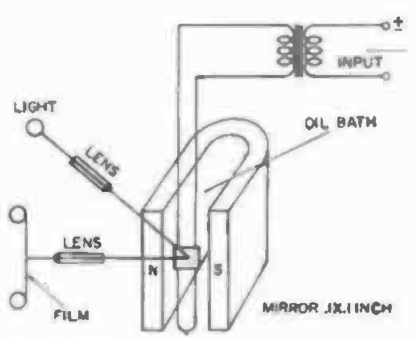


Fig. 22-75B. Basic principles of an early model recording galvanometer.

optical system to the surface of a moving film. These devices had a frequency response out to about 5500 Hz, but with a strong peak at that frequency. To provide damping and reduce the peak somewhat, the whole assembly was immersed in an oil bath.

Modern galvanometers permit the user to select units with a limited frequency range, or one of several thousand hertz. Similar galvanometers are available for use with direct current for balancing bridge circuits. Light is projected on the mirror surfaces and reflected to a scale several feet in length and mounted at a distance from the galvanometer, thus expanding the reflection of the movement for precise measurements. Galvanometers employing a small mirror are also called light-reflecting galvanometers and are used for scientific research and in graphic recorders using light-sensitive paper.

22.76 Describe an external oscilloscope calibrator and its use.—Many of the older oscilloscopes did not employ an internal voltage calibrator as do most of the later designs. Voltage calibration was accomplished by applying a known voltage from an external calibrating unit which developed a square wave, then passed to a voltage divider network calibrated in terms of peak-to-peak voltage. A typical circuit for such an external calibrator unit is given in Fig. 22-76A.

The circuitry consists of 6AL5 diodes, an internal calibrating resistor, and a precision output attenuator. The power supply furnishes both positive and negative bias voltage to the diodes at a voltage level of 75 volts. A sine-wave voltage of approximately 300 volts is taken from one side of the power transformer and applied to the clipper diodes. Both sides of the sine wave are clipped to form a square wave, which is applied to a series calibrating resistor R1, and also to a precision voltage divider.

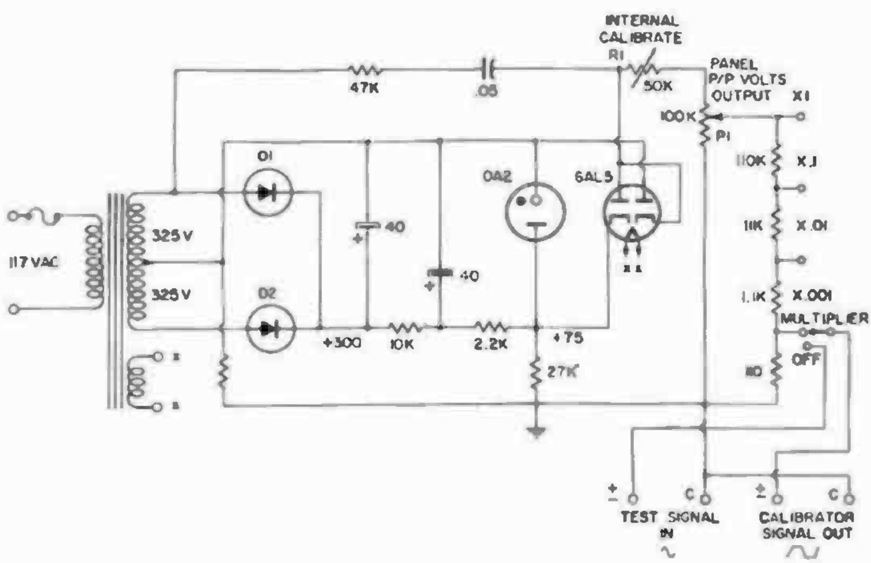


Fig. 22-76A. Schematic diagram for oscilloscope external voltage calibrator.

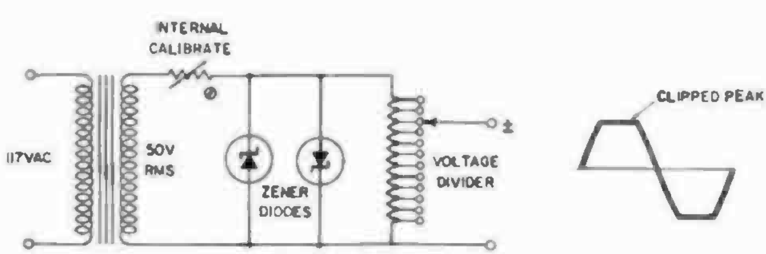


Fig. 22-76B. Oscilloscope voltage calibrator using two zener diodes.

When initially calibrated, series calibrating resistor R1 is adjusted for exactly 100 volts peak-to-peak (70.7 rms) at the top of the voltage dividing network. The divider network has five positions: 0.001, 0.01. 0.1, 1.0, and an off position. In this latter position, the signal under measurement bypasses the internal circuitry of the calibrator unit, permitting it to be applied directly to the input of the oscilloscope.

To measure the amplitude of a signal displayed on an oscilloscope, the signal under measurement is first referenced on the graticule, and the signal from the calibrator unit applied to the input. Potentiometer P1 and the multiplier of the calibrator are adjusted for an equal deflection on the CRT. When the two signals are equal, the voltage is read from the calibrator potentiometer and multiplied by the multiplier dial.

A somewhat simplified calibrator is shown in Fig. 22-76B. Here two zener diodes connected back to back develop a square wave by clipping the peaks of the sine wave. The signal is then fed to a voltage divider. As the voltage divider in Fig. 22-76A and this unit operate at 60 Hz, no compensation is required aross the resistors.

It should be kept in mind that an oscilloscope is a peak-reading device. The calibrator may read in terms of peak-to-peak or rms voltage; this must be taken into consideration when the oscilloscope is calibrated. As a rule, calibrating devices are designed to read peak-to-peak voltage. The line voltage for either type must be fairly well regulated to maintain the accuracy of calibration.

Oscilloscopes with built-in calibrators generally employ a multivibrator circuit, followed by a cathode follower tube, with a precision voltage divider in the cathode circuit. A square wave provides a flat-top signal, which is much easier and more accurately adjusted to a reference line than the peak of a sine-wave signal.

22.77 What is the purpose of a long-persistence screen in a cathode-ray tube?—To better observe low-frequency waveforms at low repetition rates; also, to preserve the image over a period of seconds or minutes for superimposing other images over the original for comparison or photography.

However, since the development of the storage oscilloscope (Fig. 22-70F), the long-persistence tube is not used to any great extent, as it requires the CRT regularly used to be removed physically from the oscilloscope and replaced by the long-persistence tube.

Another interesting and useful oscilloscope is the split-screen type, which is capable of displaying a moving image on one half the screen, while storing an image on the other half; or, the entire screen may be used in the conventional manner. Independent control of both halves of the screen permits an image to be stored for reference on one half the other half for displaying waveforms for comparison. Either half of the screen may be erased or the storage facilities reversed.

22.78 What is Z-axis or intensity modulation of a cathode-ray tube?—It is a circuit for varying the intensity of the cathode-ray-tube beam, by applying an external signal between the cathode and ground. This modulation method is used for precise measurements of time. Either a pulse or square wave may be used. The application of a positive pulse increases the intensity of the beam, causing a series of dots, while a negative pulse blanks out the beam, causing a series of dashed lines (Fig. 22-78).

22.79 Why are electrostatic deflection circuits used almost exclusively in oscilloscope deflection circuits? — Because an electrostatic-type deflection circuit requires no appreciable current for its operation while magnetic deflection circuits require a considerable amount of current. Also, magnetic deflection circuits induce a frequency characteristic which is undesirable in a measuring device as it may induce

(a) A blanking pulse.

(b) An intensifying pulse.

Fig. 22-78. "Z" axis modulation of the beam in a cathode-ray tube.

distortion of the waveforms under observation. Magnetic deflection is quite satisfactory in television circuits, however, because they operate at a fixed frequency.

22.80 ***How are frequency-response characteristics of an oscilloscope stated?***—When the frequency response is rated dc to 10 MHz (10 MHz down 3 dB) it means the frequency response is flat to some reference frequency, then down 3 dB at 10 MHz. Three dB down is about 30 percent, or the oscilloscope has 70 percent deflection at 10 MHz, compared with 100-percent deflection at the reference frequency.

22.81 ***What effect does an input attenuator have on the frequency characteristics of an oscilloscope?***—The input attenuator, unless especially designed, will affect the frequency response because of its internal capacity and the design of the input potentiometers.

In elaborate oscilloscopes the input attenuator is adjustable in several fixed steps, in association with a continuously adjustable gain control. Each step of the attenuator is corrected for frequency discrimination (See Question 22.70.)

22.82 ***Describe an oscilloscope expander circuit.***—It is a circuit included in the horizontal amplifier for increasing the gain and expanding or magnifying the display, five, ten, or twenty times normal. This permits the user to expand a portion of a waveform and examine minute areas not visible in the normal-size display. This circuit is sometimes referred to as a *notching* circuit.

22.83 ***Define the term "writing speed" as applied to an oscilloscope.***—Writing speed is the total distance of travel the cathode-ray beam (spot) must take in both the vertical and horizontal directions in a given unit of time.

It is essential that the writing speed be known for photographing waveforms as the intensity of tha beam is greater in the horizontal direction than in the vertical; this is particularly true for transient phenomena. The easiest method of determining the writing speed is by use of a nomograph as given in Fig. 22-83. The amplitude of the trace (at the left) is connected by a straightedge to the frequency (at the right) and the writing speed read on the log-scale in the center. The term "writing speed" is analogous to the ASA "exposure index" used for rating the sensitivity of photographic film. For high-speed transients the exposure is very short (short-persistence phosphors). For extremely short exposures the reciprocal relationship of brightness and exposure no longer holds true, and the ASA index loses its significance.

22.84 ***What is a blanking pulse?***—A pulse used in a cathode-ray oscilloscope to blank out the trace of the linear time-base oscillator on its return from the right to the left side. This is accomplished by momentarily reducing the intensity of the trace.

22.85 ***What does the term "packing" mean?***—Crowding of the image on one side. This is generally caused by nonlinearity in the sweep oscillator circuits.

22.86 ***What are the various axes of an oscilloscope called?***—The horizontal axis is called the X axis; the vertical, the Y axis; and the modulating element or control grid, the X axis.

22.87 ***What are Lissajous figures or patterns?***—The patterns that appear on the face of a cathode-ray tube when voltages of different frequency, amplitude, and phase relationship are applied to its deflection plates. They were named for the French scientist, Lissajous. This subject is discused in detail in Question 23.176.

22.88 ***Describe a graticule used with a cathode-ray oscilloscope.***—Graticules for cathode-ray oscilloscopes are manufactured in several different patterns; however, the two most commonly used are illustrated in Fig. 22-88A and Fig. 22-88B. In certain oscilloscopes the graticule is made of plastic and is removable. Modern trends are to etch the graticule on the interior surface of the CRT display screen as this reduces the effects of parallax which otherwise exist because of the separation of the graticule by the thickness of the glass of the CRT.

The major lines on the graticule generally represent 1 centimeter each, and the small divisions 2 millimeters each. When properly calibrated in conjunction with the oscilloscope time-base controls, timing measurements on the order of 1 microsecond or less are possible. Graticules used on early-type oscilloscopes were generally calibrated

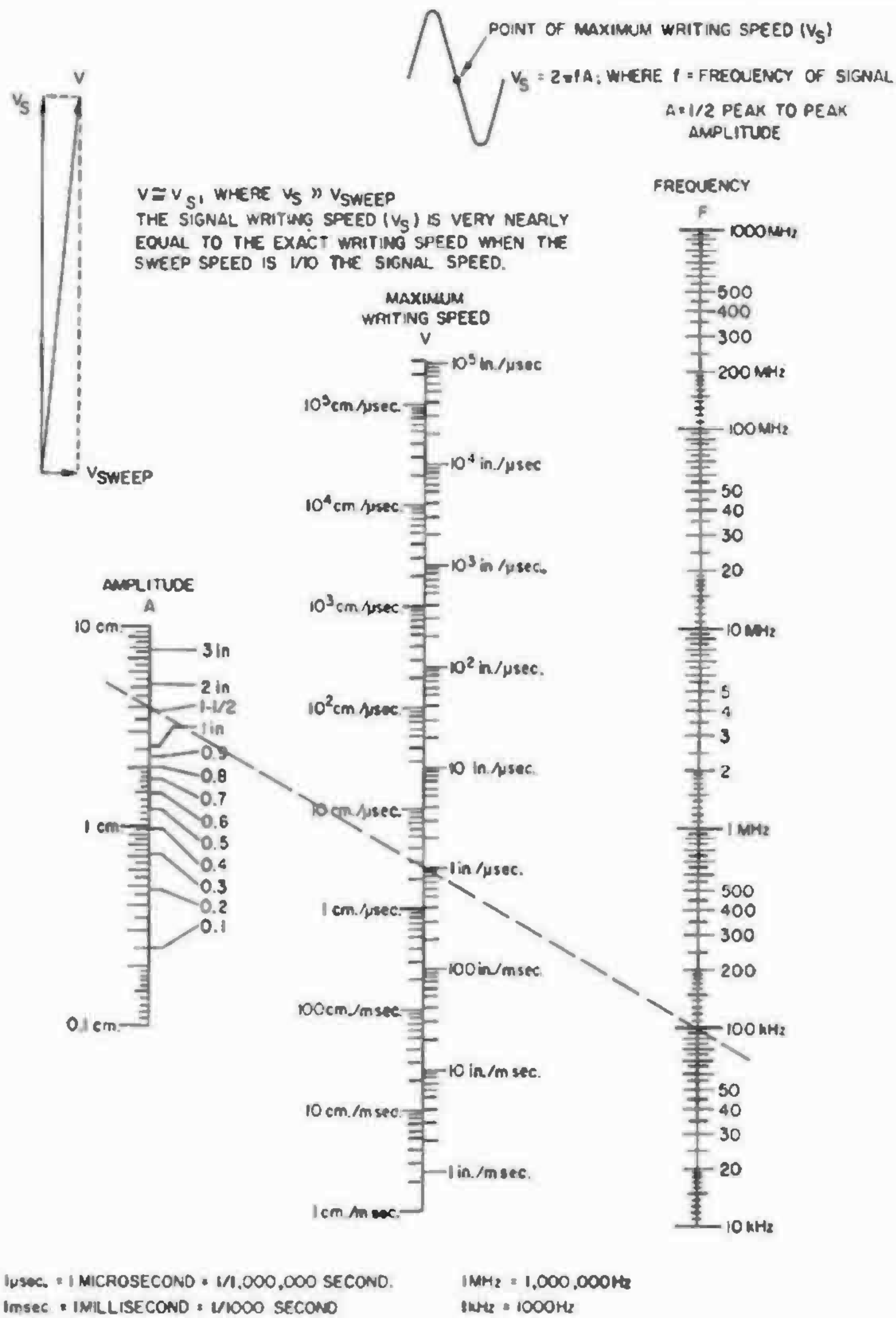


Fig. 22-83. Nomograph for relating amplitude, frequency, and maximum writing speeds for sinusoidal waveform traces. The frequency range may be extended below 10 kHz by the application of a suitable factor.

in 0.10 inch per division. Modern oscilloscope graticules are calibrated to read in centimeters and millimeters.

22.89 ***What is an oscilloscope current probe?***—Current probes are of the clip-on type and are designed for use in measuring very small alternating and direct currents in electronic equipment. Current ranges are generally from less than 1 milliampere up to several amperes, over a wide range of frequencies.

The probe and the wire carrying the current form a one-turn transformer. The wire carrying the current is the

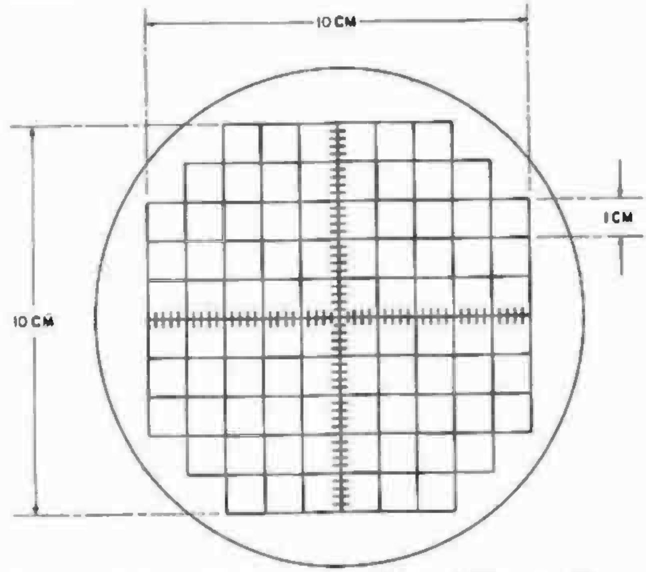


Fig. 22-88A. Graticule for cathode-ray tube calibrated 10 cm × 10 cm. Each small division represents 2 mm.

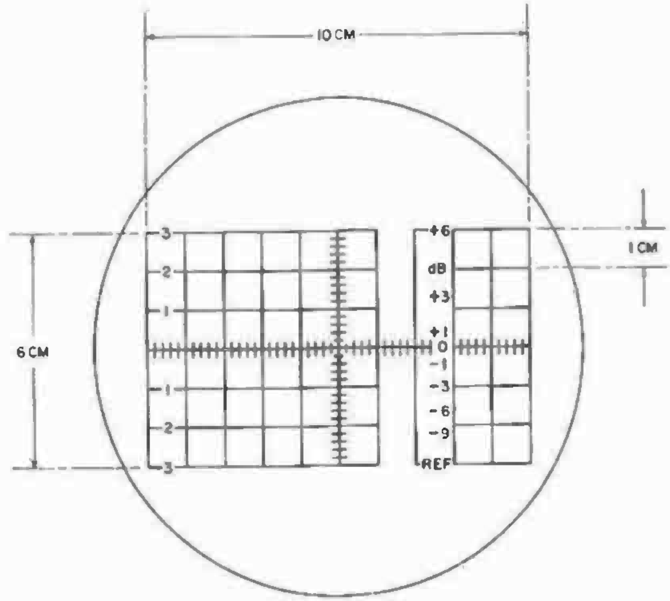


Fig. 22-88B. Graticule for cathode-ray tube calibrated 10 cm × 6 cm, with decibel scales. Each small division represents 2 mm.

secondary and the probe is the primary. Probes used for ac measurements employ a special amplifier which is connected to a conventional vacuum-tube voltmeter or oscilloscope. The dc-type probe is used with a special meter which is similar to a vacuum-tube voltmeter. The schematic diagram for a typical current probe with the appearance of the probe head is shown in Figs. 22-89A and B.

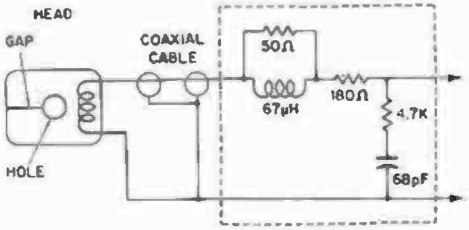


Fig. 22-89A. Schematic diagram for a typical current probe.

22.90 Describe a low-capacitance oscilloscope probe. — Because the input capacitance of the average cathode-ray oscilloscope will load down a circuit of high impedance and thus affect the frequency response and waveforms, a probe of very low capacitance is connected between the input of the oscilloscope and the circuit under observation. However, low-capacitance probes cannot be built without inducing loss; therefore they are generally designed to have an attenuation ratio of 10 or 100, although other ratios are sometimes employed.

The schematic diagram for the internal circuitry of a 10:1 oscilloscope probe is shown in Fig. 22-90A. Only two components are required: a precision 9.5-megohm resistor and a small variable capacitor, ranging from 2.5 pF to 12 pF, to compensate for the capacitance of the oscilloscope inpnt and cable. If the oscilloscope has an internal square-wave calibrator, the tip of the probe can be connected to the output of the calibrator and the waveform observed on the oscilloscope. The capacitor in the probe is then adjusted for a square waveform as shown in the top illustration of Fig. 22-90B. If an external square-wave generator is used, it should be monitored with and without the probe to ensure that the waveform seen with the probe is similar to that seen without the probe.

The internal circuitry for a 100:1 probe is given in Fig. 22-90C. Here three resistors are used, with a small compensating capacitor connected across the 9.9-megohm resistor. This probe is also adjusted nsing a square-wave generator.

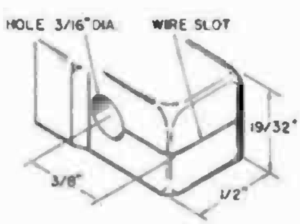


Fig. 22-89B. Appearance of the head for a typical current probe.

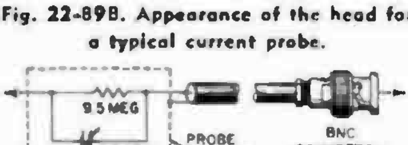


Fig. 22-90A. Internal connections for a 10:1 ratio oscilloscope probe.

22.91 What is a demodulator probe? —A probe designed for use with an oscilloscope for displaying modulated high-frequency signals. If a high-frequency signal is demodulated (detected) by a demodulator probe connected to the input of an oscilloscope, the modulation envelope shape may be seen on

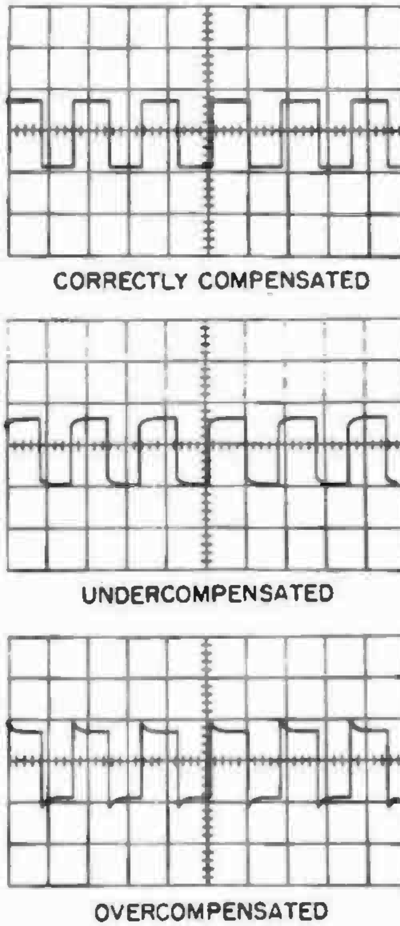


Fig. 22-90B. Correctly and incorrectly compensated oscilloscope probe. The internal capacitor is adjusted for the upper waveform with a square wave applied to the probe.

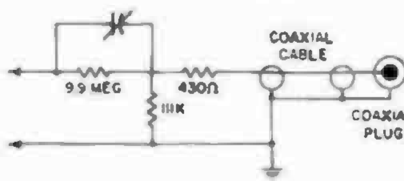


Fig. 22-90C. Circuit for a 100:1 oscilloscope probe. The small capacitor is to compensate for the cable and input capacitance of the oscilloscope.

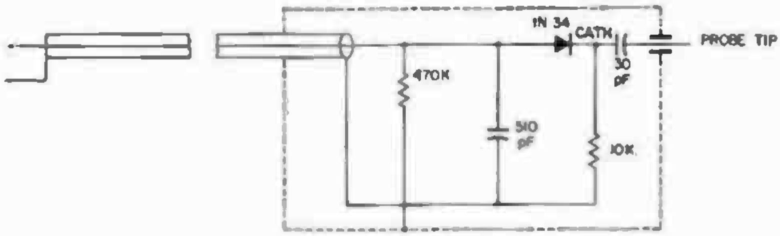


Fig. 22-91A. Schematic diagram of a demodulator probe.

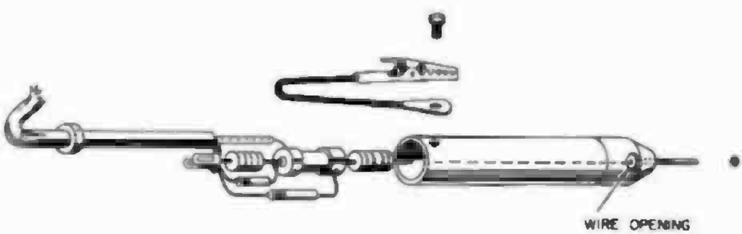


Fig. 22-91B. Internal construction of a typical demodulator probe.

an oscilloscope of limited frequency response because only the modulated envelope is displayed. The schematic diagram of a typical demodulator is shown in Fig. 22-91A.

A demodulator probe may also be used with a dc or ac vacuum-tube voltmeter. When used with a dc vacuum-tube voltmeter, the indication is proportional to the average amplitude of the signal. Used with an ac meter, the indication is proportional to the peak voltage of the modulation envelope waveform. Most demodulator probes induce a loss of signal voltage on the order of 5:1.

22.92 What is a high-voltage probe? —A probe with a high internal resistance for measuring extremely high voltages with a voltmeter having an internal resistance of 20,000 ohms per volt or greater. Fig. 22-92 shows a typical high-voltage probe designed to operate in conjunction with a 20,000-ohm-per-volt meter. The internal resistance of the probe is 600 megohms, permitting 30,000 volts to be measured using a 300-volt scale meter. Other scales may also be used by applying the correct multiplying factor.

High-voltage probes are fitted with a guard to protect the user and reduce corona discharge. Because of the large distributed capacitance of a high-voltage probe, it cannot be used for observing voltages containing high-frequency waveforms. The internal resistance of the average high-voltage probe is 100 to 1200 megohms.

22.93 Describe a panoramic spectrum analyzer. — Panoramic spectrum analyzers are used for the visual display

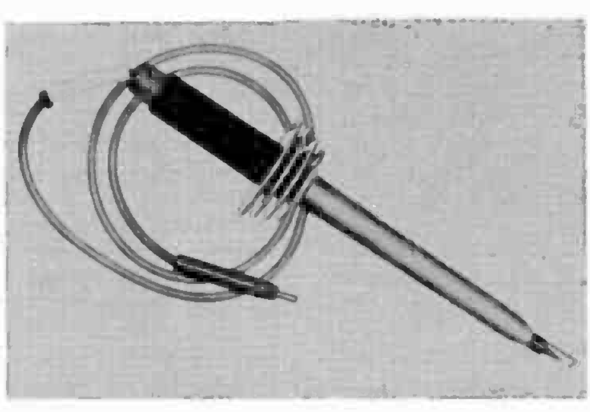


Fig. 22-92. Typical high-voltage probe.

Fig. 22-93A. Singer Co., Metrics Div., Model TA-2 universal spectrum analyzer, with AL-2 plug-in module for 20 Hz to 35 kHz spectrum analysis.

of harmonic and intermodulation distortion, equalizer and filter characteristics, vibration, noise and acoustical analysis, and radio-frequency measurements. A Model TA-2 universal spectrum analyzer, manufactured by Singer Co., Metrics Div., is pictured in Fig. 22-93A, using a plug-in module. This model is designed to cover a bandwidth of 20 Hz to 35 kHz and is calibrated in absolute values of both frequency and level.

A simplified block diagram of the device appears in Fig. 22-93B. A panoramic spectrum analyzer is a scanning superheterodyne receiver displaying level versus frequency of the input signal and may be set to scan through a selected frequency band. Levels are displayed deflections on a calibrated cathode-ray-tube graticule. Frequency is calibrated on the CRT horizontal axis. Horizontal sweep on the CRT is related to the frequency sweep excursion of the analyzer. Both the horizontal deflection and local oscillator are driven by the same sawtooth generator. Thus, a calibrated visual indication of input frequencies is obtained.

Referring to the diagram, the incoming signal is applied to a eleven-position attenuator adjustable from zero to 60 dB in 10-dB steps, and to a 40-dB gain preamplifier, which is selected as required. A sawtooth generator modulates a voltage-controlled oscillator, causing it to scan the selected spectrum segment. The local-oscillator frequency excursions are set by a calibrated center frequency and sweep-width control. The CRT graticule is calibrated in 10-percent increments of sweep width.

The signal from the swept local oscillator combines with the input signals in a diode ring modulator. Balanced modulator operation reduces mixer nonlinear distortion, thus extending the spurious free range. The swept oscillator mixes with input frequencies in the balanced modulator to produce differing frequencies equal to the intermediate frequency and provides the ability to resolve individual input frequencies.

The i-f amplifier section consists of the three 100-kHz variable-bandwidth crystal filters. The bandwidth of these filters is varied automatically for optimum resolution for each sweep width. Following the i-f amplifier is an attenuator for adjusting the display levels, then a linear detector to demodulate the high-frequency i-f output signals into low-frequency envelopes or pips for display. Contained in the detector circuit is a log-compression amplifier which can be switched in to display over a 100:1, 40-dB range. The linear and dB scales are calibrated directly on the CRT graticule.

Each plug-in module contains an internal crystal-controlled oscillator to produce evenly spaced markers (fundamental and harmonics) for accurately checking the sweep width and center-frequency calibrations. The markers are spaced at 2.5-kHz intervals, and may be adjusted independent of the signal deflection. A typical display is shown in Fig. 22-93C. Other plug-in modules covering bandwidths up to 27.5 MHz are available.

22.94 *What is a sound-level meter?* —A sound-level meter is an instrument designed for the measurement of sound-pressure level (SPL). It consists of an omnidirectional microphone, calibrated attenuator, amplifiers, standard weighting networks, and an indicating meter. The frequency characteristics of the weighting networks must meet the standards specified by the United States Standards Institute (ASA) and the International Electrotechnical Commission. Instruments meeting these standards are accepted for the measurement of both product and environmental noise by industry, laboratories, and noise-abate-

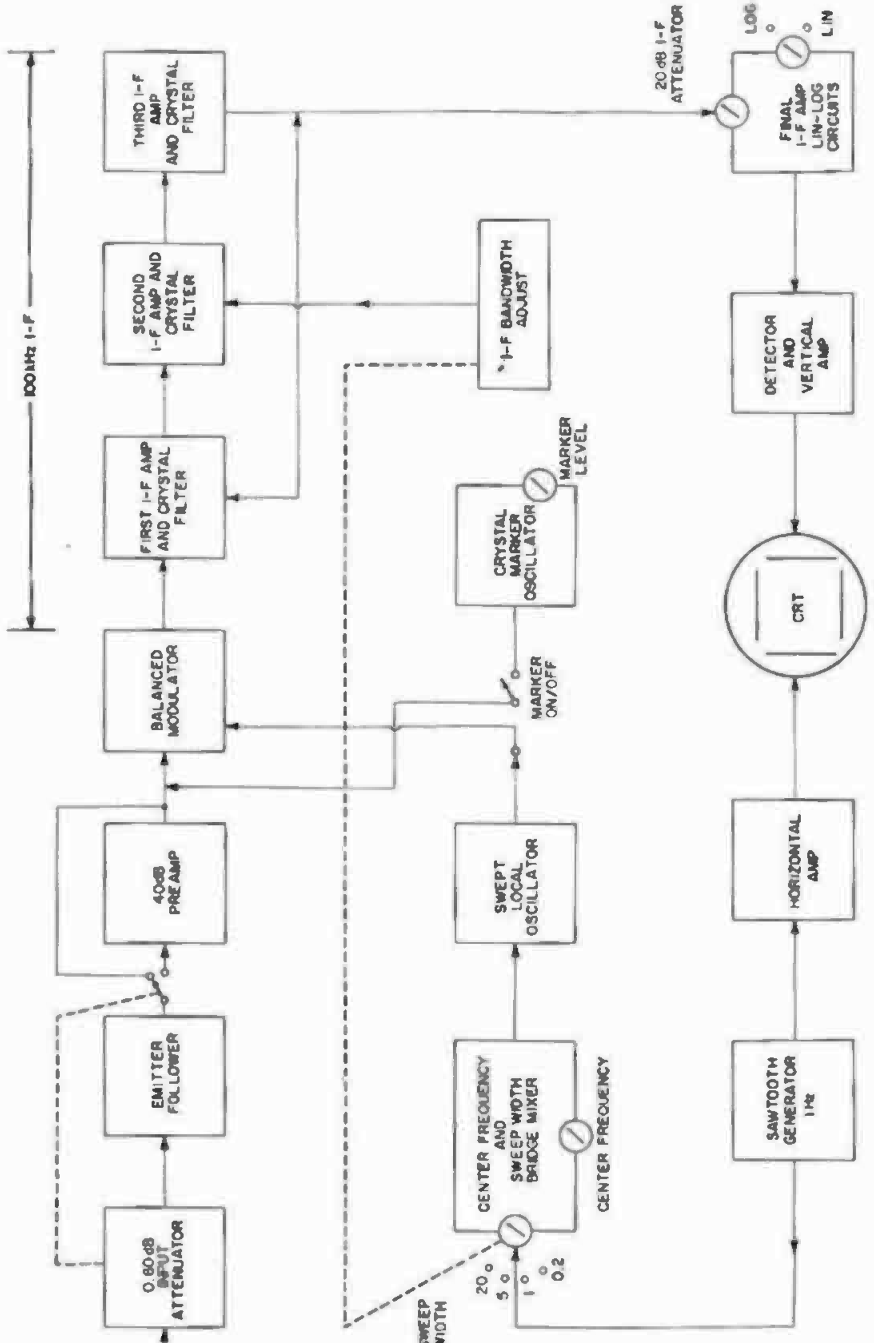


Fig. 22-93B. Block diagram for Singer panoramic spectrum analyzer.

ment groups. These instruments can also be used for the measurement of the acoustic characteristics of sound stages, monitoring rooms, auditoriums, and similar enclosures. Sound-level meters can usually be used in combination with such instruments as spectrum analyzers, vibration pickups, special-purpose microphones, and graph recorders. As a rule, they are completely self-contained and portable. Such a meter, Model 1551C, manufactured by General Radio Co., is shown in Fig. 22-94A.

Referring to the block diagram in Fig. 22-94B, a PZT-type piezoelectric ceramic microphone (essentially non-directional) feeds an attenuator network calibrated 24 to 150 dB (re: 0.0002 μbar), then to a second attenuator and a weighting network. The signal is again amplified and passed through a second weighting network, amplified by a three-stage amplifier, into a third attenuator and weighting network, amplified, and applied to an indicating meter. An internal calibration system permits the adjustment of the instrument before use.

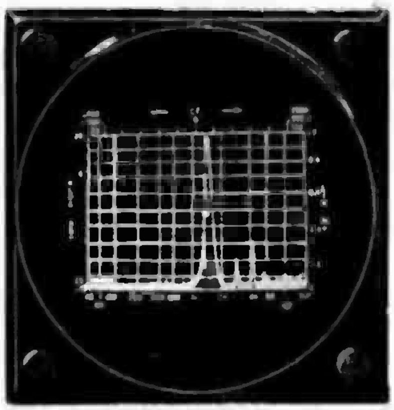


Fig. 22-93C. Typical waveform as seen on the analyzer graticule. The fundamental frequency is the large deflection, with its harmonics appearing to the right of it.

The four-weighting-network frequency characteristics (shown in Fig. 2-93) meet the USASI (ASA) Standard S-1.4 1961. A fourth frequency characteristic provides a flat response from 20 to 20,000 Hz for use with wide-range microphones. Two meter movement speeds are available, fast and slow. In the slow position the meter is heavily damped to show the average level of rapidly fluctuating sounds, and to comply with the above standard.

An output jack is provided for the connection of external equipment. The internal output impedance is 7000 ohms, developing 1.5 volts into an open circuit. It may be used with a lower impedance; however, for greater accuracy the impedance should be 20,000 ohms or more. This output is often used to drive

Fig. 22-94A. General Radio Co. Model 1551C sound-level meter.

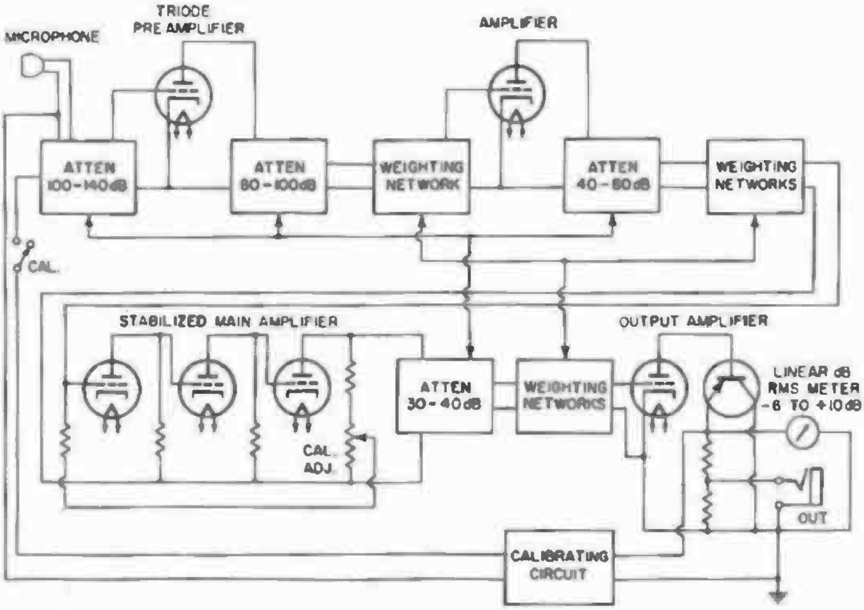


Fig. 22-94B. Block diagram for General Radio Co., Model 1551C sound-level meter.

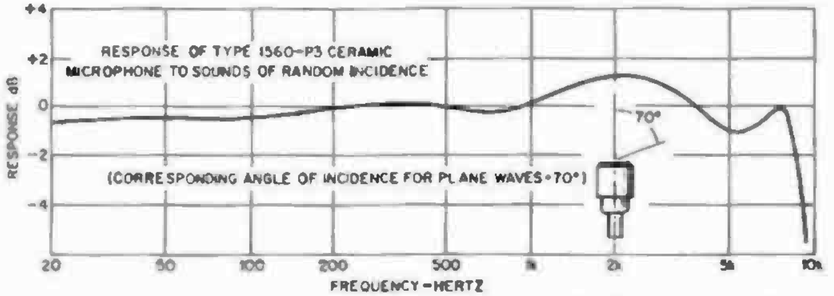


Fig. 22-94C. Frequency characteristics of General Radio Co. sound-level meter ceramic microphone.

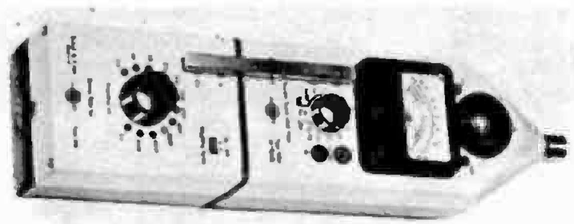


Fig. 22-94D. Brüel and Kjaer Model 2203 sound-level meter with octave band filter.

a tape or graph recorder, wave analyzer, or other devices.

When acoustical measurements using a sound-level meter are made, the observer's body and the meter case have little effect in a semireverberant sound field. However, under free-field conditions and a single sound source both the observer and the instrument case have a large effect on the microphone response. These effects can be minimized if the instrument is held in front of the observer, with the observer oriented so that the sound passes in front of him at a right angle to the axis of the sound-level meter. For greatest accuracy the microphone should be mounted on a tripod and connected to the sound-level meter by an extension microphone cable, to eliminate the effects of the observer's body and also of the meter case.

When the measurements are made in a reverberant field, the angles of incidence of sound reaching the microphone are indeterminate. Under these conditions there is no preferred angle of incidence between the microphone and the sound source. When measurements are made in a free field, an angle of incidence of 70 degrees between the axis of the microphone and sound source will approximate random response.

The instrument described contains six miniature vacuum tubes, one transistor, seven diodes, and is completely battery operated, but it may be used with an external power supply if desired. The frequency response for the microphone at an angle of incidence of 70 degrees is given in Fig. 22-94C.

A smaller version of a sound-level meter, manufactured by Brüel and Kjaer, is pictured in Fig. 22-94D. It has the same standard weighting characteristics, operating over a range of 22 to 134 plus or minus 1 dB, 20 to 18 kHz. The eleven-octave band filter covers a frequency range from 31.5 Hz to 31.5 kHz. Each passband attenuation is adjustable from 0 to 48 dB.

When this meter is used with a sound source principally from one direction, the readings can be adversely affected by the relative positions of the instrument and the observer's body. In this instance the meter should not be held directly in front of the observer with the microphone pointed to the sound source, since the high frequencies are increased to a marked degree due to the observer's body acting as a reflector. Thus, errors of several decibels in the region above 100 Hz are possible. The most uniform response is obtained when the meter is held in front of the observer, but with the sound grazing the microphone (coming from the side rather than the front). Out of doors the microphone is pointed upward (to avoid reflections) as far from the observer as possible.

It should be noted that conventional sound-level meters do not measure intensity; therefore the term "intensity" should not be used in connection with such meters. In accordance with the USASI (ASA) the terms "sound level" and "sound-level pressure" (SPL) are to be used.

22.95 What is an electronic switch and how does it function?—An electronic switch is an instrument which may be used with a conventional cathode-ray oscilloscope for the simultaneous observation of two or more waveforms on the screen of the oscilloscope. Since the cathode-ray tube is essentially a single-signal indicator, a device such as an electronic switch is necessary for the observation of more than one signal waveform simultaneously. Thus, an electronic switch permits the conventional oscilloscope to be used as a multisignal comparison device. Due to the persistence of the human vision and the persistence of the cathode-ray-tube fluorescent screen, two signals can be made to appear on the screen simultaneously. While the employment of a single electronic switch permits the simultaneous observation of two signals, two electronic switches connected in tandem will permit the simultaneous observation of three waveforms. Additional switches may be added, if desired, one for each additional waveform to be observed. However, unless the screen of the oscilloscope tube is 7 inches or larger, observation of three or more traces is not too satisfactory. (See Question 23.184.)

The most frequent use of an electronic switch is for the observation of amplitude, waveform, phase, and frequency relationship between two signals. The signals under study may be those of an electrical or electronic device, or a sound or mechanical device transformed into electrical functions. A block diagram for connecting an electronic switch and an oscilloscope to an amplifier for the simultaneous observation of the input and output waveforms is given in Fig. 22-95A.

Four controls are provided on the electronic switch. They are: a frequency control for setting the frequency of the internal square-wave generator, A and B channel input attenuators, and an image separation or trace position control.

The schematic diagram of an electronic switch is shown in Fig. 22-95B. One of the two signals to be observed, designated A, is connected to the input marked "A" and the other signal, called B, is connected to input "B." The input attenuators of channels A and B are adjusted for a signal level suitable for the operation of the switch.

Starting at input A, the signal is fed in two directions, to the control grid

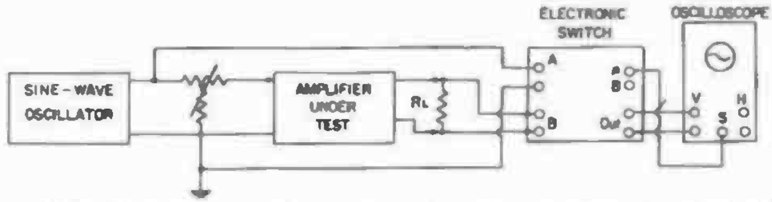


Fig. 22-95A. Block diagram of connections for using an electronic switch with a conventional oscilloscope.

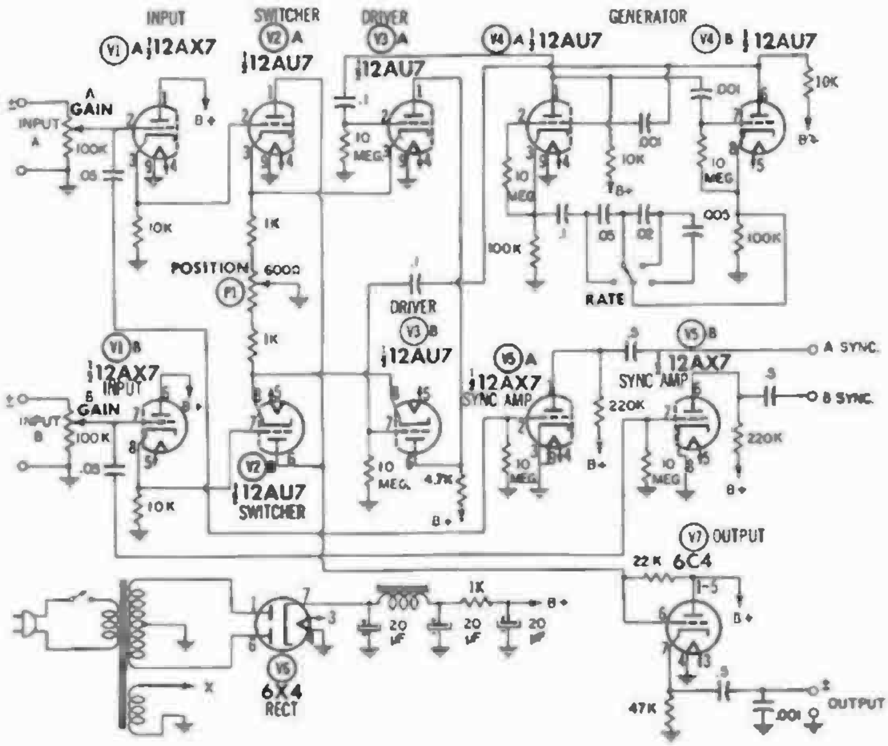


Fig. 22-95B. Schematic diagram of a typical electronic switch.

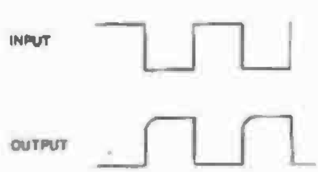


Fig. 22-95C. Appearance of an input and output signal reproduced simultaneously on a cathode-ray tube.

of an input cathode-follower tube V1A and to the control grid of a sync-amplifier tube V5A, and is used for locking the time base or sweep circuit of the oscilloscope to the signal frequency under observation. The B signal passes through a similar type of circuitry.

The cathodes of tubes V1A and V1B are directly connected to the control grids of the switcher tubes, V2A and V2B. A switching generator circuit comprised of tubes V4A and V4B generates a square wave which is applied to the control grid of the driver tubes V3A and V3B. The frequency of the square-wave generator is controlled by capacitors connected to a switch in the cathode circuit. Repetition frequencies of 150, 500, 1500, and 5000 Hz may be generated.

The cathodes of switcher tubes V2A and V2B are directly connected to the cathodes of the driver tubes, V3A and V3B. The square waves generated by the square-wave generator permit the switcher tubes to be operated normally or to be driven to cutoff. When switcher tube V2A in channel A is operating normally, the switcher tube in channel B is driven to cutoff. When the switcher tube in channel B is operating normally, the switcher tube in channel A is driven to cutoff.

The plates of both switcher tubes are fed through a common plate-load resistor; thus any signal applied to the inputs of channels A and B will appear across this common load resistor alternately at the frequency rate of the square-wave generator.

Signals from the switcher tube plates, which are connected in parallel, are fed to the control grid of a cathode-follower output tube for connection to the vertical input of an oscilloscope. Potentiometer P1 is used to separate the two traces (A and B) as they appear on the oscilloscope, by changing the value of the bias voltage on the cathodes of the switching and driver tubes. Thus the traces made by signals A and B may be made to overlap or be separated one above the other. Because the transition from one circuit to the other in the electronic switch does not occur simultaneously, a switching transient in the form of a spike may appear on the trace as seen on the oscilloscope.

A series and a shunt capacitor are connected in the cathode circuit of the output tube, V7. The series capacitor isolates the vertical input of the oscilloscope from the dc voltage across the cathode resistor of the output tube. The shunt capacitor limits the high-freqnency response to transients. The value of this capacitor may be varied for the most desirable presentation on the oscilloscope screen; however, for most purposes the value shown will suffice. The waveforms of Fig. 22-95C show how two traces might appear on the face of the oscilloscope. The upper one is the input signal and the lower one is the output signal. The rounding of the leading edge of the output trace is caused by frequency discrimination within the amplifier.

22.96 Describe the circuitry for a direct-reading frequency meter.—Frequency meters are instruments designed for the direct measurement and indication of the frequency of an unknown signal voltage. The waveform of the voltage under observation may be a sine wave, square wave, sawtooth or pulse. With the proper transducer, rotary motion, acoustical, and other forms of repetitive pulses may be determined. Since the reading of frequency is independent of the input voltage waveform, the meter may also be used to indicate the frequency of random events during a given time period.

The schematic diagram for a frequency meter of this type, manufactured by Hewlett-Packard, is shown in Fig. 22-96A. This meter is designed to cover frequencies between 10 and 50,000 Hz in ten ranges. Basically the circuit consists of a limiter amplifier, electronic switching circuit, pulse counter, and a constant-voltage, constant-current power supply.

The unknown frequency is applied to the input of two limiter tubes, V1 and V2. These tubes amplify and flatten the peaks of the incoming signal voltage, resulting in a square-wave signal voltage at the plate of V2. From the plate of tube V2, the signal is fed in two directions: through capacitor C4 to the control grid of switcher tube V4, and through capacitor C5 to the control grid of V3, a phase inverter. From the plate of V3 the signal is applied to a second switcher tube V5.

Space current for the switcher tubes is obtained from a constant-current regulator tube V7. Alternate half-cycles of the square wave cause a constant current through tube V4 and charge one of capacitors C9 to C18. During the intervening half-cycle, a constant current flows through V5 and causes one of capacitors C19 to C28 to be charged. The time constants of the two RC combinations R33, C9 to C18 and R32, C19 to C28 are equal.

The resistor and capacitors in the charging circuits are such that at the highest frequency to be counted, the capacitors will be fully charged before the end of the half-cycle. The accurately controlled pulses for the two banks of charging capacitors are converted to unidirectional pulses by diodes CR1 to CR4, and the resultant current is ludicated by meter M1. The meter indication is proportional to the number of pulses per unit time and therefore to the frequency of the voltage at the input of the instrument. Resistors S34 to R43 are shunts across the meter to adjust the current through the movement to the correct value for each frequency range. These shunts are factory adjusted.

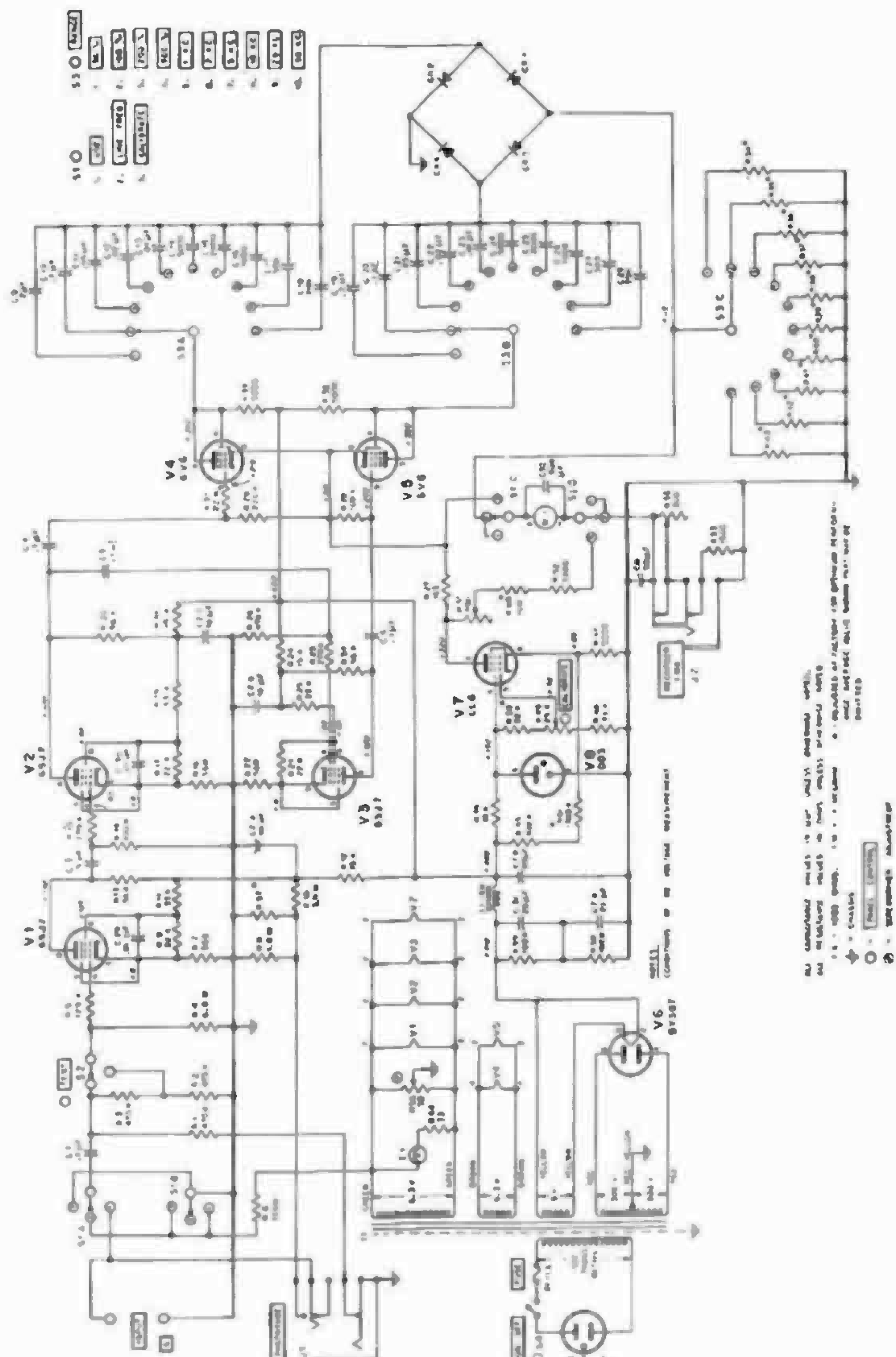


Fig. 22-96A. Schematic diagram for Hewlett-Packard Model 500A electronic frequency meter.

The constant-current regulator circuit consists of tubes V7 and V8. Tube V8, a gas voltage-regulator tube, provides a constant voltage at the screen of tube V7. A voltage divider circuit consisting of resistors R48 to R50 is connected between the screen grid of V7 and ground. Variable resistor R49 adjusts the grid voltage of V7 to produce the desired constant current for the

Fig. 22-96B. Hewlett-Packard electronic frequency meter Model 500B. Frequency range 3 to 100,000 Hz.

switching tubes, which is measured by switching the meter across shunt resistor R27 in the plate circuit of V7.

A graphic level recorder may be driven by connecting it at the recorder jack. Resistor R53 and R56 adjust the current for the recorder and provide a dummy load when it is not in use.

When the instrument is used with a phototube or photocell, black and white stripes are placed on the moving part and illuminated by a bright light. Alternate reflections are picked up by the phototube from the moving stripe and passed from its preamplifier to the frequency meter. Vibration pickups are used in a similar manner.

A front view of a meter similar to the one discussed but having a frequency range of 3 to 100,000 Hz is shown in Fig. 22-96B. This meter has an expanded range that permits the reading of any 10- or 30-percent portion of a selected range to be expanded to full scale, thus increasing the accuracy of the reading.

In both instruments the line frequency voltage is available for checking the calibration. (See Question 22.196.)

22.97 ***Describe the circuitry of a transistor chopper-type dc multivoltmeter.*** — Multivoltmeters are extrmely sensitive meters for the measurement of minute voltages, and they require special treatment for stability and sensitivity. The instrument to be described is a direct-current millivoltmeter (Fig. 22-97A), Model PM2430, manufactured by Phillips of Holland. It is all transistor and battery operated.

Referring to the block diagram of Fig. 22-97B, when a dc voltage is applied to the input terminals it is first impressed on an attenuator network and hum filter, then converted by a chopper to a square-wave voltage. This signal is then amplified and applied to an intermediate attenuator network that determines the 3-mV and 10-mV voltage ranges. The chopper vibrates at a frequency of 50 or 60 Hz. An astable multivibrator generates an ac voltage for the chopper operation and an adjustable voltage to compensate the preliminary deflection (linearity). (See Question 22.59.) The indicating meter circuit consists of three direct-coupled transistor stages, using a Graetz rectifier circuit, incorporated in a negative feedback loop. An automatic phase-sensitive polarity indicator indicates the polarity of the dc input voltage under measurement, and it also may be used as a zero-center indicator.

A complete schematic diagram for this instrument appears in Fig. 22-97C. When the dc voltage is applied to the input terminals (BU1 and BU3), it is attenuated by attenuator network SK2.

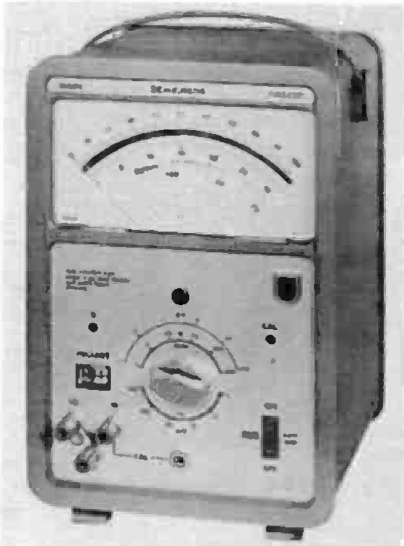


Fig. 22-97A. Dc millivoltmeter Model PM2430 manufactured by Phillips of Holland. (Courtesy, Phillips Electronics Industries Ltd. Canada.)

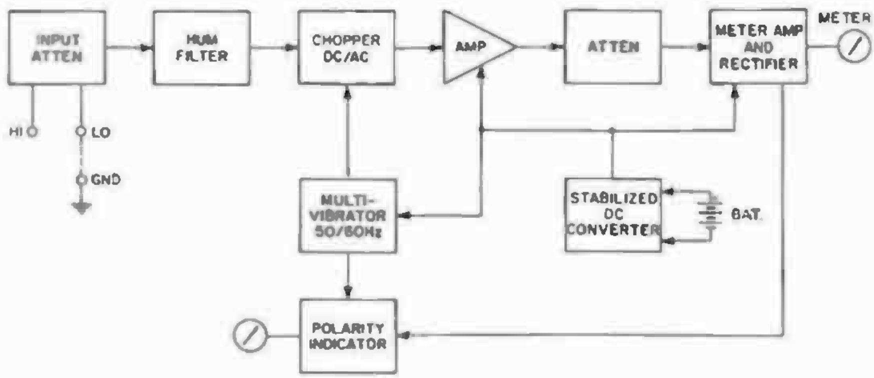


Fig. 22-97B. Block diagram for Phillips Model PM2430 millivoltmeter.

Apart from the 1-, 3-, and 10-mV ranges, attenuation for other voltage ranges is supplied by resistors R301 to R315. Leaving the attenuator, the signal is then applied to a hum filter comprised of R103 to R106, and C101 to C103. The chopper (U4) converts the incoming dc voltage into a square-wave voltage that controls field-effect transistor TS101 via capacitor C105. Zener diode GR101 supplies a constant bias voltage for TS101 through voltage divider resistors R111 and R112. Neon tube B101 provides protection against a too high input voltage. Resistor R101 is factory adjusted to provide a 1-megohm input resistance.

To establish the linearity, a compensation voltage is applied to one contact of the chopper element and adjusted by potentiometer R1. This voltage is supplied from coil S2 of transformer T201, and rectified by diode GR201. The signal from the chopper is amplified by a three-stage amplifier TS102 to TS104. To stabilize these circuits, negative feedback is taken from the emitter of TS104 and applied to the junction of resistors R116 and R117, in the emitter element of TS102. Diode GR102 limits the base-emitter voltage if transistor TS103 is overloaded. The output voltage of TS104 is now applied to the second attenuator (U3) defining the 3- and 10-mV ranges.

The second attenuator is followed by a power amplifier stage (TS105 to TS107) with negative feedback. The Graetz rectifier circuit (GR105 to GR108) for the indicating meter is included in the feedback circuit between the collector of TS107 to the emitter of TS106. Potentiometer R2 adjusts the amount of negative feedback and controls the sensitivity and calibration of the metering circuit. Diode GR104 limits the ac voltage to prevent overloading of TS107 so that the meter indicates full scale. The meter circuit is further safeguarded against current surges by transistor TS108, which will conduct at a voltage exceeding 0.4 volt to act as a protective shunt across the meter movement.

When the instrument is first turned on via switch SK1 a current surge occurs in transformer T202 to charge capacitors C209 and C207 through diode GR206. This current surge causes an induced voltage in secondary S2 of transformer T202. Current then flows in the forward direction through the base-emitter junction of transistor TS-210 through resistor R222 and capacitor C208. This action causes the collector emitter junction to conduct a current through winding S1 of T202. The current rises to the saturation point of the transformer (or the moment when the base of TS210 blocks) because of the decreasing induced voltage.

As the collector-emitter junction is blocked, the voltage across transformer T202 rises strongly as a consequence of self-induction, and capacitor C209 is charged through diode GR206. During the discharge phase of the transformer a new induction voltage arises across winding S2. However, as the polarity of this voltage is reversed, the base of TS-210 is blocked and there is no current. In this manner only the induction current flows through the diode GR206 to charge capacitors C209 and C207. When the induction voltage drops below the battery voltage, the battery again sup-

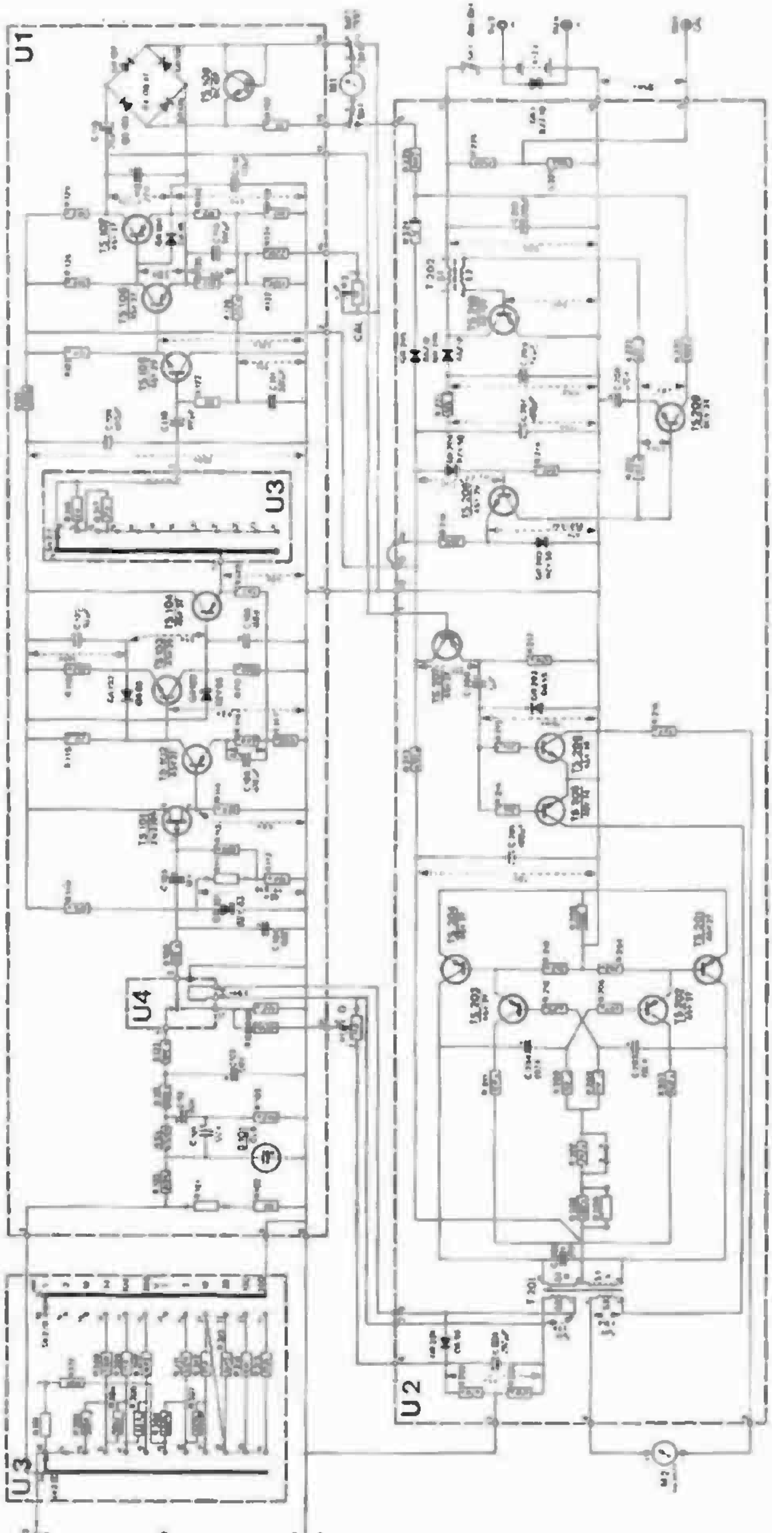


Fig. 22-97C. Schematic diagram for Phillips Model PM2430 dc millivoltmeter.

plies a new current surge, since diode GR206 functions as a valve through which current can flow only in the charge direction. Diode GR205 and resistor R224 permit a fast charging of C207 equal to the battery voltage; thus, oscillation is reestablished.

The output voltage at C207 is controlled by a bridge circuit. The reference voltage is formed by the total value of the zener voltages across diodes GR203 and GR204. When the output voltage across C207 exceeds the reference voltage, the collector current of transistor TS208 decreases. Consequently, transistor TS209 is recontrolled, which causes a decrease of the average base curreut in transistor TS-210.

Because the oscillating frequency of TS210 will be lower at a small base current than at a greater base current, electrolytic capacitors C207 and C209 are charged less, affecting the output voltage and maintaining the reference voltage at an almost constant value. At a higher current consumption or at a lower battery voltage the oscillating frequency will be higher than with a smaller load current or with new batteries.

When the supply voltage is applied to capacitor C205, current flows through resistor R208, the emitter-base circuits of TS201 and TS202, and resistors R206 and R204 to charge capacitor C204. The same procedure occurs in the other part of the symmetrical circuit when one capacitor is more highly charged than the other as a consequence of a small asymmetry of transistor characteristics.

The charging process of the less intensely charged capacitor is immediately suppressed by a collector current caused by the bias-emitter current of TS201. As a result, capacitor C203 is charged with a polarity opposite to that of C204. As the charge drops, the base-emitter current of TS201 decreases and the charging current through the collector-emitter circuit of transistor TS 201 is blocked. The capacitors uow interchange their charges as current through the base-emitter circuits of transistors TS204 and TS203 causes C203 to become positive, and current flows through winding S1 of transformer T201.

Because of the alternate conduction and blocking of transistors TS201 and TS204, voltages are introduced in transformer windings S1a and S1b. The voltage induced in winding S2 feeds the field coil of the chopper unit. Winding S3 serves to compare the phase for the polarity indicator.

The polarity meter is independent of the applied dc voltage, and its movement is five to ten times more sensitive than the instrument. It functions as follows. The multivibrator generates 50 or 60 Hz, which is applied to transformer T201, which in turn drives the chopper element. Winding S3 supplies an ac voltage to the collectors of transistors TS205 and TS206. The dc voltage to be measured (now converted to a square wave) is amplified and controls the base of transistor TS207, an emitter-follower, to achieve further current amplification.

Depending on the phase of the voltage at the bases of TS205 and TS206 with respect to the collector side, the current for the polarity-indicator movement will flow in one direction or the other. Since the voltage at the bases of TS205 and TS206 is negative, diode GR202 will conduct and block these two transistors. For each positive alternation through the diode, the transistors will conduct. The current through the polarity indicator changes direction according to the phase of the base-emitter voltage, with respect to collector voltage, thus indicating the polarity of the dc voltage applied to the input terminals of the instrument.

The field coils of the chopper unit (U4) are normally fed from 60 Hz generated by the astable multivibrator. To obtain sufficient output power, both transistors of the multivibrator circuit are preceded by an emitter follower. To avoid interference phenomena which might cause vibration of the indicating meter, the frequency of the multivibrator is adjusted to 50 Hz when used around 60-Hz equipment, and 60 Hz when used with 50-Hz equipment. This is accomplished by strapping across resistor R207. With the strap removed, the frequency of the multivibrator is 50 Hz. Resistor R229 is factory selected to adjust the circuit to the correct frequency.

The instrument is designed to be operated above ground (floating), or the low-side terminal may be grounded as required. The total range of this in-

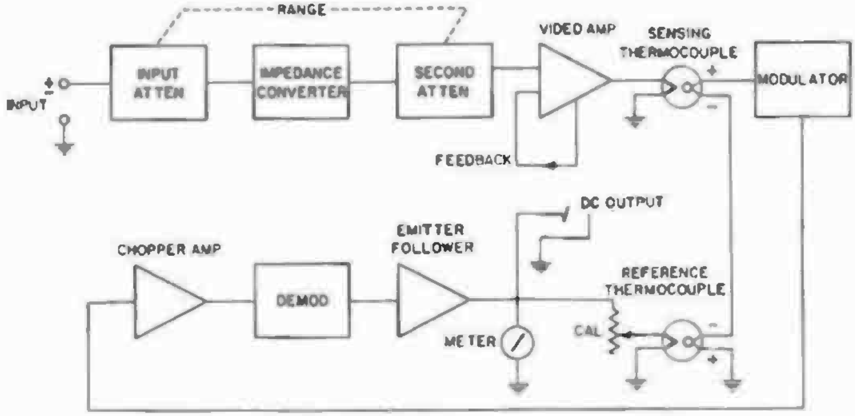


Fig. 22-98A. Basic circuit for Hewlett-Packard Model 3400A rms voltmeter.

strument is 1 millivolt to 300 Vdc in twelve ranges, with an accuracy of 1 percent. Provision is also made for use with a high-voltage probe, increasing the voltage range to 30 kilovolts. The input resistance for the 1- to 10-millivolts scales is 1 megohm, and for the 1 to 300-volt scales is 100 megohms. The polarity indicator will operate for an indication of less than 3 percent full-scale. This instrument may be obtained for either battery or ac line operation. The batteries are rechargeable.

22.98 What is a true-responding voltmeter?—Electronic voltmeters are classified into three broad categories—rms-responding, peak-responding and average-responding. The majority of ac voltmeters are usually either average- or peak-responding meters, with the meter scale calibrated to read the rms

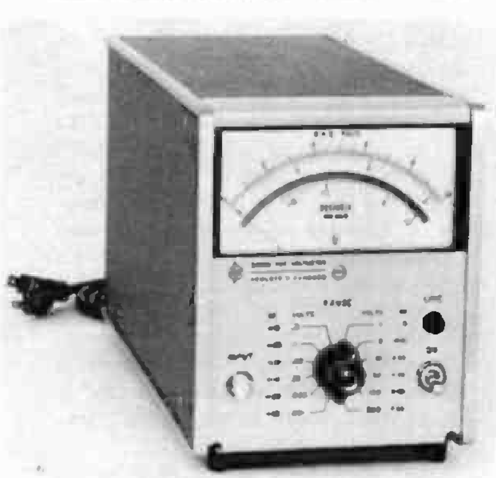


Fig. 22-98B. Hewlett-Packard Model 3400A rms voltmeter.

value of a sine wave. Electronic voltmeters are ac-to-dc converters, which derive a dc current proportional to the ac input signal and use this current for the meter deflection. The conversion from ac to dc eliminates serious errors which otherwise result from a meter movement sensitive to frequency. The difference in the preceding three meters lies in their interpretation of the value of the input signal.

The rms value of a sine wave was established as an equivalent dc voltage which generates the same amount of heat in a resistive load that an ac voltage generates. For this reason, rms voltage is synonymous with effective voltage. Rms voltage is defined as the square root of the average of the squares of the quantities being measured. Theoretically, this can be done by measuring the voltage point-by-point along the waveform of one cycle, squaring the numerical value of the voltage at each point, and then finding the average value. Regardless of the waveform shape this procedure leads to the rms or effective value. The rms value of a sine wave is 0.707 times the peak or maximum value of the voltage waveform. (See Question 25.149.)

The average value of an ac voltage is simply the average of the voltage values measured point-by-point along the waveform (omitting the squaring and root extraction). The average value of a sine wave is really zero, because the waveform has equal positive and negative values when averaged for one whole cycle. Since the equivalent dc or energy content in the waveform is usually the quantity of interest, the average value of the sine wave is taken to mean the average rectified value. The average value of one half cycle is 0.636 times the peak value.

The peak-responding voltmeter is of little interest to the audio engineer; however, at times it can be useful. The preceding discussion leads to a meter quite useful for the measurements of audio frequencies, because it reads the true rms value of a sine wave, is not affected by the waveform, and is usable over a frequency range of 10 Hz to 10 MHz. The basic circuit for such a voltmeter, the Hewlett-Packard Model 3400A, is given in Fig. 22-98A. In this instrument are two thermocouples in combination with an ac and dc amplifier. Nonlinear effects are cancelled in the measuring thermocouple by similar nonlinear effects in the second thermocouple.

The voltage under measurement is applied to an input attenuator through a coaxial connector on the front panel. The input attenuator has an impedance greater than 10 megohms, and provides two ranges of attenuation. The output of the input attenuator is applied to an impedance converter, a noninverting unity-gain amplifier. The purpose of the impedance converter is to present a high impedance to the input signal and provide a low impedance for driving the second attenuator network. The second attenuator provides six ranges in a 1, 3, and 10 sequence. These two attenuators are ganged and provide twelve ranges of attenuation, or a voltage range of 1 millivolt to 300 volts, with a decibel range of minus 72 to plus 52 dBm.

The output of the second attenuator is amplified by a wideband amplifier comprising five stages. The overall gain of this amplifier is controlled by an ac feedback loop. The ac output of this amplifier is applied to a sensing thermocouple. The dc output from the thermocouple is applied to a modulator consisting of two photocells which are alternately illuminated by two neon lamps (90 to 100 Hz), which are in turn controlled by an oscillator. The output of a sensing thermocouple is also applied to one photocell. The resultant output of the modulator is a square wave whose amplitude is proportional to the dc input level.

The square wave from the modulator is amplified by a chopper amplifier composed of a high-gain three-stage ac amplifier. The output of this amplifier is applied to a demodulator. The output of the demodulator is a dc level whose magnitude is proportional to the amplitude of the ac input signal. The output of the demodulator is applied to a two-stage direct-coupled emitter follower. The emitter-follower stage supplies an impedance transformation from the high impedance of the demodulator to the low impedance of a direct-reading meter and a reference thermocouple.

The reference thermocouple acts as a summing point for the ac output of the video amplifier and the dc output of the emitter follower. The difference in the heating effect of these voltages is fed as a dc input to the modulator. The difference input is amplified and fed to the reference thermocouple and the meter. This amplified voltage represents the rms value of the ac signal applied to the input of the instrument.

A regulated power supply provides plus 75 volts and negative 6 and 17.5 volts. Since the voltage at the dc output jack is proportional to the meter deflection, the instrument can also be used as a linear rms ac-to-dc converter for driving a graph recorder. As loading does not affect the meter, both the meter and recorder may be used simultaneously.

A front view of the complete instrument is shown in Fig. 22-98B.

22.99 ***Describe an audio-frequency vacuum-tube voltmeter.*** — Fig. 22-99A shows a Heathkit Model IM-21 vacuum tube voltmeter, designed to operate over full-scale ranges of 10 millivolts to 300 volts rms, with an input impedance of 10 megohms shunted by 22 pF on all ranges. The frequency response is plus or minus 1 dB from 10 Hz to 500 kHz, and plus or mnus 2 dB from 10 Hz to 1 MHz for all ranges. A decibel scale is also provided, reading from minus 40 dB to plus 50 dB (re: 1 milliwatt) at 600 ohms. As for most rectifying-type ac meters the meter deflection is proportional to the average value of the input waveform.

Referring to the schematic of Fig. 22-99B, the voltage to be measured is applied to a frequency-compensated 1000:1 voltage divider and then to the control grid of an input cathode follower V1A. Input voltages for the

Fig. 22-99A. Heathkit Model IM-21 audio-frequency vacuum-tube voltmeter.

lower six ranges are coupled directly to the control grid, while for the four higher the ranges are divided by 1000 and coupled to the grid of V1A from the lower tap of the voltage divider.

Cathode-follower stage V1A provides a low-impedance source for the signal applied to the voltage divider feeding the input of amplifier section V1B and V2. The voltage divider network in the cathode of V1A divides the signal into six different levels to provide ten scales with readings from 10 millivolts to 300 volts.

Approximately 19 dB of negative feedback is returned through the meter circuit from the plate of V2 to a potentiometer in the cathode circuit of V1B. This negative feedback loop provides high stability and uniform gain over a wide frequency range. The meter circuit consists of a 200-microampere meter movement with a full-wave bridge rectifier of four germanium diodes. For calibration purposes (setting the gain), the amount of meter current is adjusted by potentiometer R18 in the cathode of V2B, which varies the amount of negative feedback.

The power supply consists of a half-wave rectifier containing a single silicon rectifier and filter capacitor C17. To minimize hum voltages the heater winding is balanced to ground through resistors R28 and R29.

22.100 Describe the schematic for a cathode-coupled vacuum-tube voltmeter. —A block diagram of such a voltmeter, manufactured by Hewlett-Packard, is shown in Fig. 22-100A. The essential components consist of an input voltage divider controlled by a range switch, a cathode-follower input tube, a precision step attenuator controlled by the range switch, a broad-band amplifier, an indicating meter, and a regulated power supply. The voltage applied to the input terminals for measurement is divided by 1000 before application to the input cathode follower, when the range switch is set to the 1-volt range and higher. The input voltage is applied directly to the cathode follower on the lower ranges. Voltage from the cathode follower is divided in the precision attenuator to be less than 1 millivolt for application to the voltmeter amplifier. The output of the amplifier is rectified in a full-wave bridge rectifier, with a dc milliammeter across its midpoints. The resultant direct current through the meter is directly proportional to the input voltage. Both the block and schematic diagrams should be referred to in the following discussion.

Referring to Fig. 22-100A, the input voltage divider limits the signal level applied to the input cathode follower to less than 0.3 volt rms, when voltages above this level are measured with the range switch is set to the 1-volt range above. The divider consists of a resistive branch with one element made adjustable to obtain an exact 1000:1 division at middle frequencies, and a parallel capacitive branch with one element made adjustable to maintain exact 1000:1 division to beyond 4 MHz. The input impedance of the voltmeter is established by this divider and is the same for all positions of the range switch. On the six low-voltage positions of the range switch, the input divider provides no attenuation of the input voltage.

The step attenuator in the cathode circuit of the input cathode follower reduces the voltage to be measured to 1 millivolt or less for application to the voltmeter amplifier. Each step of the attenuator lowers the signal level by exactly 10 dB, by the range switch rotor, which has two contactors. The first contactor contacts each resistor in turn while the input divider is in the

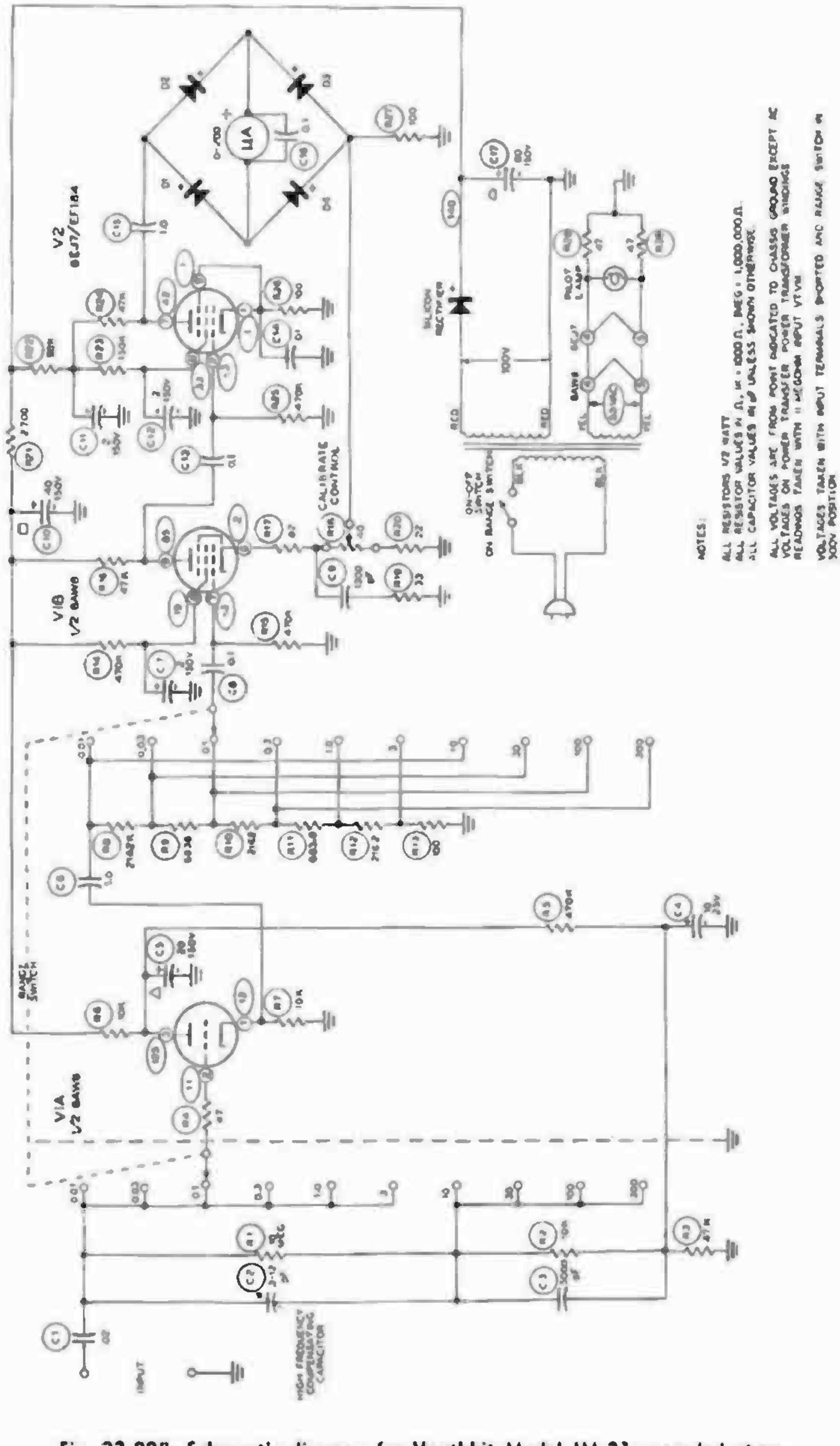


Fig. 22-99B. Schematic diagram for Heathkit Model IM-21 cascoded vtvm.

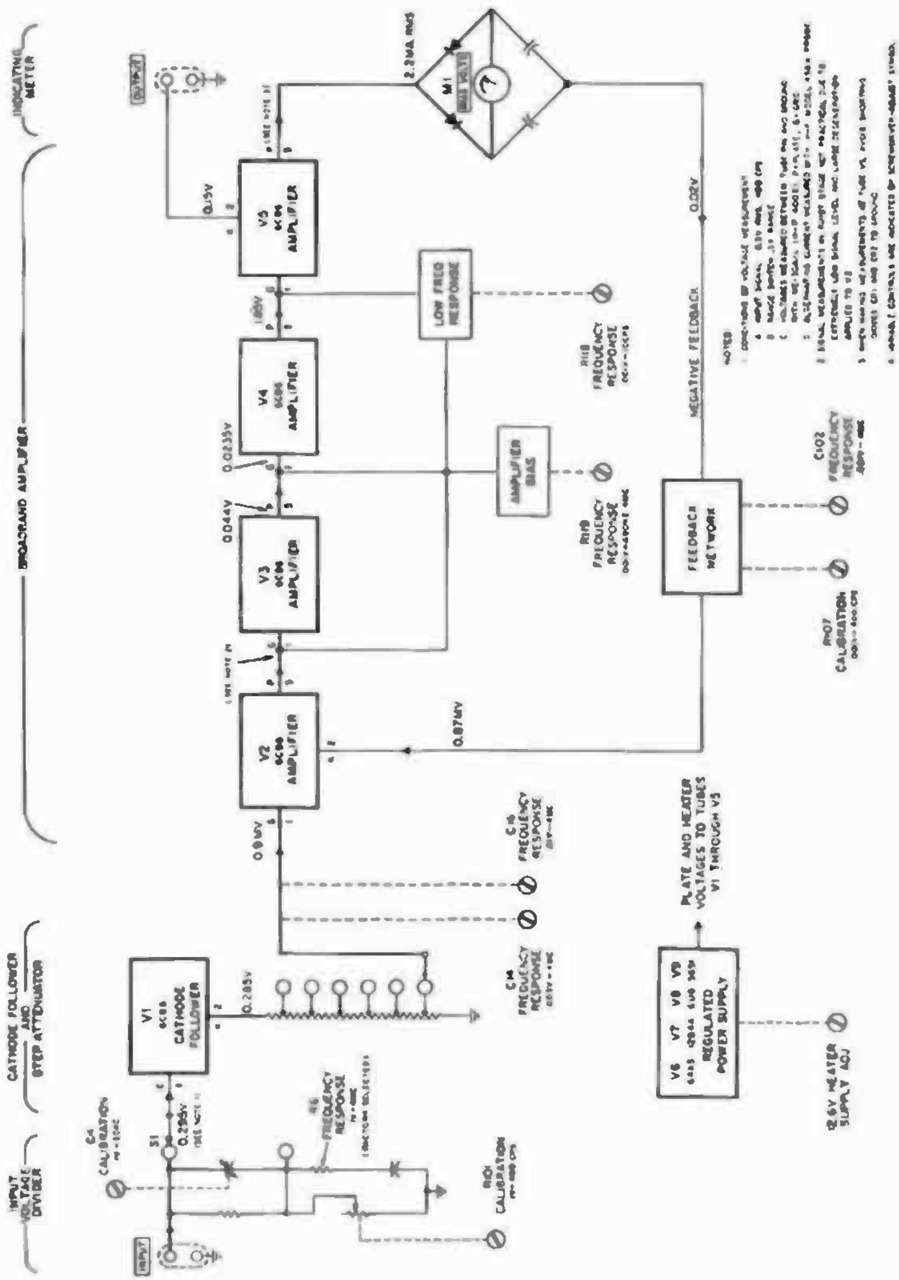


Fig. 22-100A. Block diagram for Hewlett-Packard 400D vacuum-tube voltmeter.

nonattenuating position; the second rotor finger repeats these contacts while the input attenuator is in the attenuating position. On the 0.001-volt range C15 is automatically connected to provide a flat frequency response beyond 4 MHz. In the 0.003- and the 0.01-volt ranges, separate adjustable capacitors C14 and C16 are automatically connected to the attenuator to permit setting the frequency response at 4 MHz. C14 and C16 are also connected to the attenuator on the 3- and 10-volt ranges. Fixed capacitor C106 (permanently connected) flattens the frequency response on the 0.03- and 30-volt ranges.

Cathode follower V1 provides a constant high input impedance to the input voltage and provides a relatively low impedance at its cathode circuit to

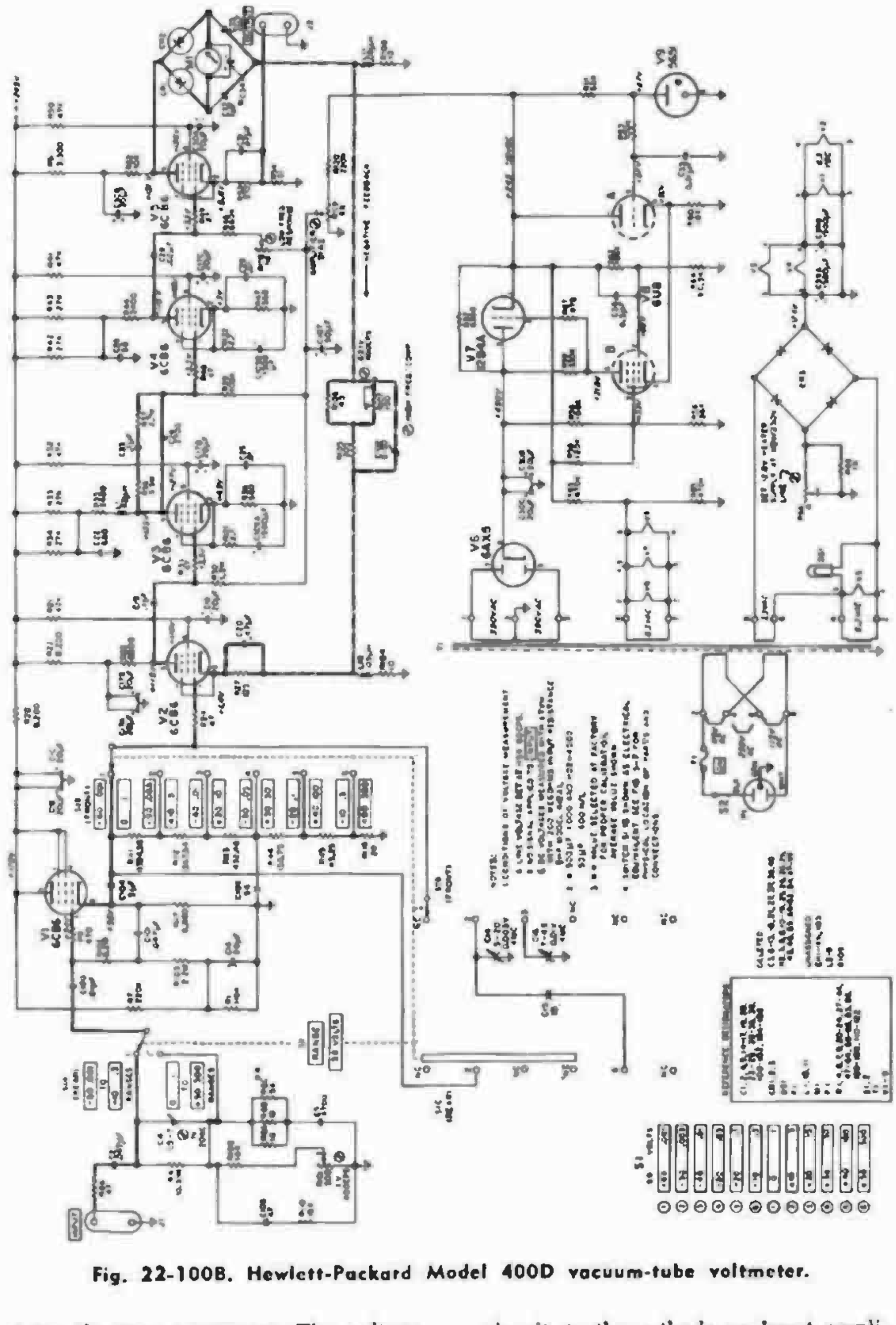


Fig. 22-100B. Hewlett-Packard Model 400D vacuum-tube voltmeter.

drive the step attenuator. The voltage gain factor across V1 is 0.95.

Amplification of the signal voltage is provided by a four-stage stabilized amplifier consisting of tubes V2 through V5. The amplifier provides between 55 and 60 dB of gain, with about 55 dB of negative feedback at midfrequencies. The feedback signal is taken from the plate of output amplifier V5, through the meter rectifiers and gain-adjusting circuit, to the cathode on input amplifier V2. Variable resistor R107 in the feedback network adjusts the negative feedback level to set the basic gain of the amplifier at midfrequencies, while adjustable capacitor C102 permits setting the amplifier gain at 4 MHz. Variable resistor R118 in the coupling circuit between V4 and V5 permits adjusting the gain of the amplifier at 10 Hz by controlling the phase shift of low-

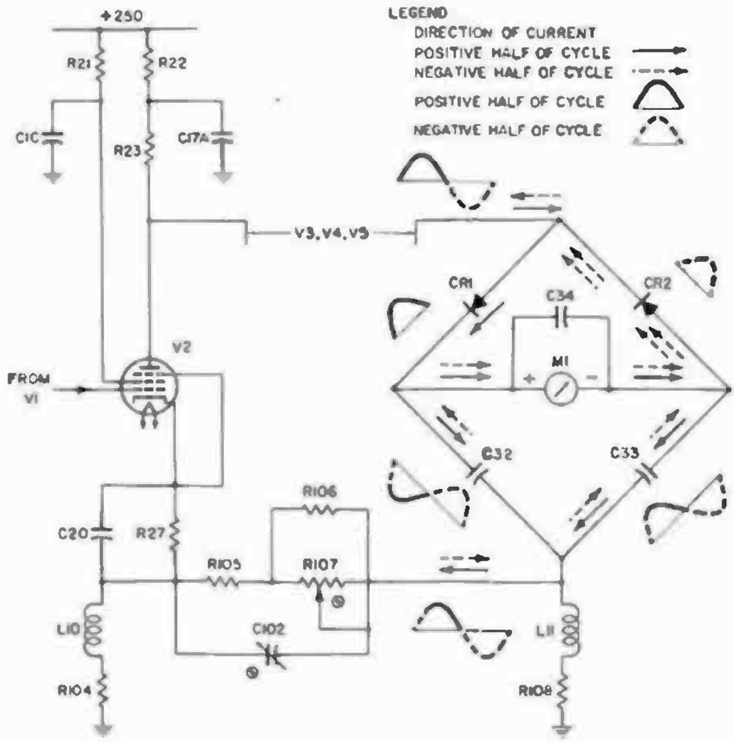


Fig. 22-100C. Simplified schematic of meter bridge circuit.

Fig. 22-100D. A front view of Hewlett-Packard Model 400D vacuum-tube voltmeter.

frequency signals between these two stages (increasing phase shift decreases degeneration and increases gain).

Variable resistor R119 in the grid-return path for V3, V4 and V5 adjusts the total transconductance of these tubes in order to restrict the maximum gain-bandwidth product of the amplifier. The gain-bandwidth product must be restricted to give a smoth frequency response rolloff above 4 MHz and to prevent possible unstable operation at frequencies far above 4 MHz when tubes having usually high transconductance are used. The plate voltage from V5 is fed to the meter output terminals for monitoring purposes. The current through V5 and thus the signal voltage at the cathode is affected by the loading of the meter rectifiers. For signal levels causing third-scale or more meter deflection, this distortion consists of a very small irregularity near 0 volts on the waveform as each diode begins conduction.

The meter rectifier circuit consists of two silicon diodes and two capacitors connected as a bridge, with the indicating meter across the midpoints as shown in Fig. 22-100C. The diodes provide full-wave rectification of the signal current for operating the meter. Electron flow through the meter is supplied in the following manner. During the positive half-cycle of the plate voltage on V5, rectifier CR1 conducts electrons from both C32 and C33 back to the B+ bus. The portion of electrons from C33 flows through the meter on the way to B+. At this point in the cycle, both C32 and C33 are charged to the potential of B+ minus some small drops in R51 and R52.

During the negative half-cycle of the plate voltage of V5, rectifier CR2 conducts electrons back to both C32 and C33 from the plate of V5. That portion of electrons going back to C32 flows through the meter on the way, in the same directions that the electrons flowed in the first positive half-cycle. At this point in the cycle, both C32 and C33 are discharged. The pulsating current through the meter is smoothed by C34 to prevent meter-pointer vibration when low-frequency signals are measured. The current is proportional to the arithmetic average value of the waveform amplitude of the signal. Meter calibration in rms volts is based on the mathematical ratio between the average and rms values of true sine-wave current.

In addition, the bridge circuit serves as a segment of a voltage divider (in series with L11 and R108) connected across the output of the amplifier. The negative-feedback voltage fed to the input of the amplifier is obtained across L11 and R108. The alternating charge and discharge of C32 and C33 produce, at their junction with L11, an alternating current of the same phase and waveform as that at the plate of V5. This phase is negative with respect to the input signal applied to the first stage of amplifier V2, and drives the negative feedback network.

The power supply consists of tubes V6 through V9 and the associated circuits, as shown in the schematic diagram of Fig. 22-100B. The power supply furnishes regulated +250 Vdc for the plate and grid bias circuits of tubes V1 through V5, and unregulated 12.6 Vdc for the heater supply of tubes V1 through V4, and 6.3 Vac for heater supply of V5 through V8. The power supply is designed to operate from either a 115-volt (plus or minus 10 percent or a 230-volt ac power source of 50 to 1000 Hz.

The output of rectifier V6 is applied to the voltage-regulator circuit consisting of V7 through V9, which supplies a constant +250 Vdc to the stabilized amplifier circuit of the voltmeter. Tube V7 is the series regulator tube, and V9 provides a fixed reference voltage drop with which the output voltage is applied to the control grid of V8B while the reference voltage is applied to its cathode. The difference between the control-grid and cathode voltages controls the operating point of V8B and thus its plate voltage, which in turn supplies the grid voltage for regulator V7. Any change in the regulated output of V7 produces a correction in the grid bias of V7 through the action of V8B. This action results in an essentially constant output voltage despite changes in line voltage or load on the supply. The gain of V8B is high enough to keep the output at the V7 cathode regulated to within plus or minus 1 Vdc as the V7 plate voltage is varied plus or minus 10 percent, with about 60 mA of load current. The response of the regulating circuits is fast enough to reduce the ripple in the output voltage to less than 1 millivolt, supplementing the filtering action of C30. Capacitor C36 couples the ripple component in the regulated output directly to V8B to avoid attenuation by R62. Resistor R57 shunts a small portion of the load current around V7 to prevent excessive V7 plate dissipation at high line voltages. Resistor R63 and capacitor C35 constitute a low-pass filter which prevents noise generated in V9 from reaching V8B.

The heater supply for the voltmeter tubes is divided into two sections. One section supplies dc voltage for the tubes in the input cathode follower and the amplifiers, while the other section supplies ac voltage for the tubes in the power supply. The voltage required for the heaters of tubes V1 through V4 is obtained from the 6.3- and 7.3-volt secondary windings of transformer T1. The voltage developed across the two series-connected windings is rectified by full-wave rectifier CR3, reduced to 12.6 volts by R66 and R68 in parallel, and applied to the series-parallel-connected heaters of V1 through V4. The series-parallel connection of the four heaters establishes a voltage of 6.3 Vac from one of the windings which drives CR3. The heaters of V6, V7 and V8 receive 6.3 Vac from a separate 6.3-volt secondary winding of T1. A front view of the instrument appears in Fig. 22-100D.

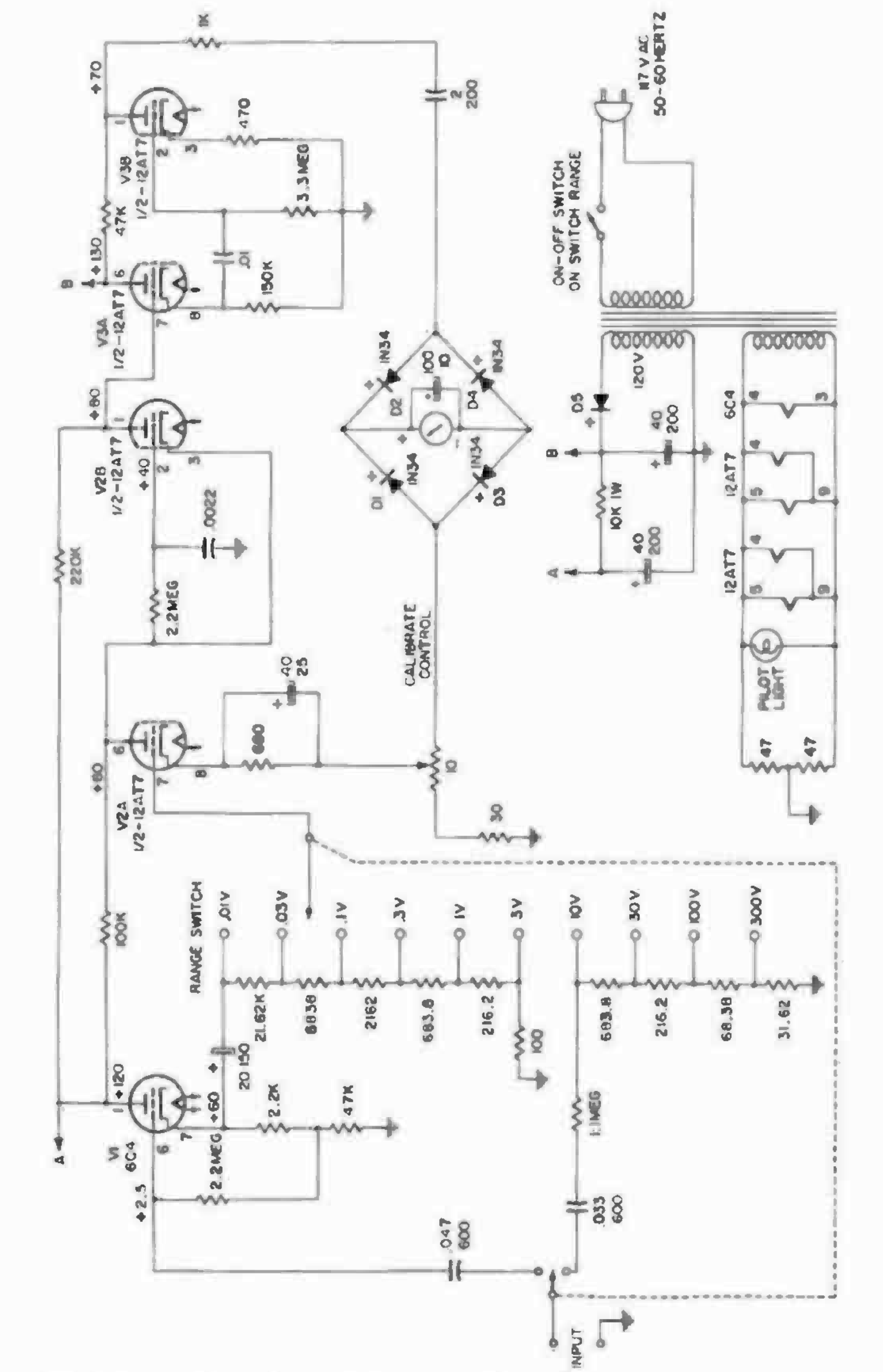


Fig. 22-101. Schematic circuit for a cascode-coupled vacuum-tube voltmeter.

22.101 ***What is a cascoded vacuum-tube voltmeter?***—It is a vacuum-tube voltmeter using a cascoded amplifier, as described in Question 12.88. The purpose of its use in a vacuum-tube voltmeter is to provide high stability, low internal noise, and increase the high-frequency response. The instrument to be described is quite similar to that of Question 22.99. It has the same voltage and decibel range but the high-frequency response is not as great, as it is limited to about 300 kHz on the lower ranges, and 40 kHz on the higher ranges. Voltage ranges are 10 millivolts to 300 volts rms.

Referring to Fig. 22-101, the input signal voltage is applied to the control grid of V1, a 6C4 tube connected as a cathode follower. This tube presents a high impedance to the voltage under measurement (about 1 megohm), and a low impedance to the voltage divider in its cathode circuit. The output from the voltage divider is applied to the control grid of V2A (half a 12AT7 tube) which is cascode coupled to its other section V2B. The plate of V2B is direct coupled to the control grid of V3A, a second 12AT7, then cathode coupled to its other half, V3B, and serves to drive the indicating meter. The meter circuit with its four diodes is returned in a negative feedback loop to a 10-ohm potentiometer in the cathode circuit of V2A. This control is used to calibrate the instrument against a source of known voltage. The capacitor connected across the meter movement aids in removing low-frequency flutter of the meter movement when frequencies below 40 Hz are measured.

Input voltages between 10 millivolts and 3 volts are applied to the input of tube V1 and measured at its cathode. For input voltages of 10 to 300 volts the voltage to be measured is applied to a voltage divider connected directly to the voltage source, through a series capacitor and resistor. It is important that the four diodes (D1 to D4) in the metering circuit be of the type specified, or ones having a good high-frequency response. The diode D5 used in the half-wave rectifier circuit to supply the operating voltages can be a silicon diode having an rms rating of 400 volts at 75 milliamperes.

22.102 ***Describe the circuitry for a vacuum-tube voltmeter using a linear decibel scale.***—Linear decibel scales are used with vacuum-tube voltmeters to eliminate crowding of the calibrations at the lower readings, and they are very useful in estimating fractional parts of a decibel because of its linear deflection. To accomplish linear deflection for a logarithmic quantity, the pole pieces of the meter are shaped to respond logarithmically (see Questions 22.19). For this type of operation, the meter has no zero calibration. Its normal position (without an input signal) is at the left-hand off-scale stop. Such a meter, manufactured by Ballantine Laboratories, Inc., is pictured in Fig. 22-102A. It will be noted that the meter scale is calibrated logarithmically for voltage and linearly for decibels. Both voltage and decibels are calibrated in terms of the rms values of a sine wave.

Since the input amplifier always operates with the same basic sensitivity (0.001 to 0.01 volt), an attenuator is provided for reducing higher voltage levels to this range. Actually, the voltage divider network consists of three separate attenuators providing reductions of 10:1, 100:1, and 1000:1.

At very low frequencies the attenuation ratio is determined entirely by the resistors. However, at the higher frequencies capacitive compensation is necessary to compensate for circuit strays, input capacitance of the amplifier, and other factors. The crossover from resistive to capacitance attenuator occurs at approximately 25,000 Hz. A schematic diagram of the instrument is given in Fig. 22-102B. It will be observed that no compensation is employed on the $\times 1000$ and $\times 100$ steps because of the relatively low resistance involved and the negligible reactive component of the resistors. Thus capacitor C6 serves to adjust the high-frequency response for the $\times 10$, $\times 100$, and $\times 1000$ steps.

The amplifier section consists of four capacitively coupled pentode stages. The first three stages are operated to provide voltage gain, while the final stage

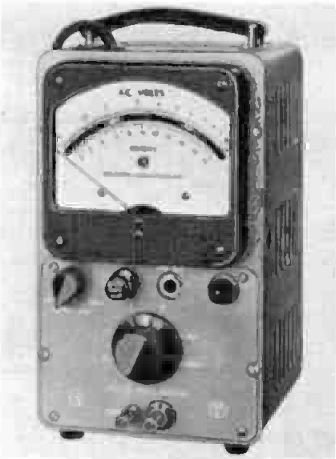


Fig. 22-102A. Ballantine Laboratories Model 300G logarithmic vacuum-tube voltmeter.

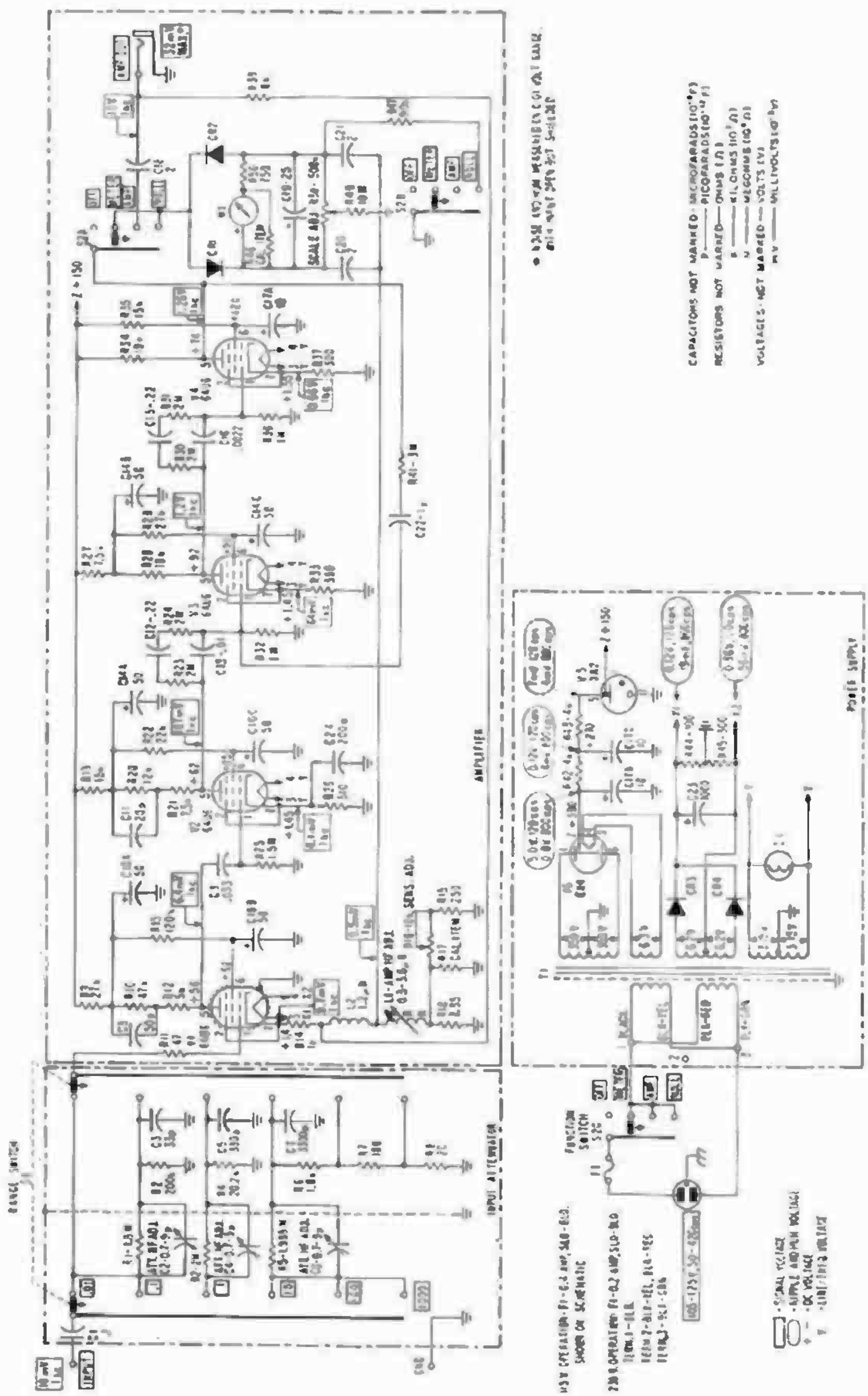


Fig. 22-102B. Schematic diagram for Ballantine Model 300G vacuum-tube voltmeter.

is used as a transducer to convert voltage to current for the rectifier meter circuit. When the instrument is operated as an amplifier only, the fourth stage also provides voltage gain.

Local feedback is achieved up to 8 dB or greater by the use of unbypassed cathode resistors, with an overall feedback of 32 dB or greater to minimize distortion and gain changes. The amplifier response to frequency without feedback is shaped to fall off from the midband value at a rate of approximately 6 dB per octave. As a result, with the amount of feedback available, the stability of the amplifier at the band extreme (10 Hz to 250 kHz) is essentially the same as in the midband.

The high-frequency response is adjustable over a narrow range by means of inductor L1 in the cathode circuit of tube V1. The gain and sensitivity (for calibration purposes) is adjustable over a small range by means of R16 in the feedback uetwork. The heater of the input stage is operated on direct current to minimize the line-frequency component injected in the low-level stage when voltages at the line frequency are measured and to reduce its harmonics.

The ac current available from the final amplifier stage is passed through the rectifier meter circuit and the feedback network to ground. The rectifier circuit is composed of two silicon diode rectifiers in full-wave bridge circuit. Capacitor C19 connected in parallel with the meter tends to reduce flutter at the lower frequencies. The response of the meter circuit is the average although it is calibrated to read in terms of a sine wave. Resistor R38 and R40 provide adjustment of the meter scale on range 1. A small current determined primarily by the value of R40 and the potential at the top of the rectifier circuit is passed through the rectifier to ground. The division and direction of this current through the meter is determined by the position of the arm on potentiometer R38. The nonlinear characteristic of the diode rectifiers is reduced to a negligible amount by connecting the rectifier circuit within the feedback loop.

Amplifier plate and screen voltages are supplied by a full-wave rectifier 6X4, employing an RC filter network, and an OA2 (V5) gaseous regulator tube. Heater voltage for tube V1 is obtained from a full-wave rectifier employing silicon junction diodes and a capacitance filter. All other heaters are operated from ac and balanced to ground by a center tap on the heater winding.

The voltage range for this meter is 1 millivolt to 1000 volts, in six full-scale decade ranges. Frequency response is 10 Hz to 250 kHz, with accuracy of 1 to 2 percent, depending on the scale used and the frequency. Input impedance is 2 megohms shunted by 15 pF, decibel range is minus 60 to plus 60 dB (re: 1 volt). The logarithmic voltage scale from 1 to 10 provides a 10-percent overlap at both ends of the scale. The linear decibel scale indicates 0 to 20 dB. Linear decibel scales may be applied to either vacuum-tube or transistor-type voltmeters.

22.103 What is the effect of nonsinusoidal waveforms on the accuracy of voltmeters calibrated to read in terms of the rms value of a sine wave?—The following data apply to both transistor and vacuum-tube voltmeters calibrated to read in terms of rms value of a pure sine wave, unless they are of the true-rms-responding type. If the voltage under measurement contains an appreciable amount of harmonics, the error in the reading will depend on the magnitude and phase of the harmonics. (See Questions 22.98 and 23.166.)

For a meter calibrated to read the average voltage, which is the calibration generally used for vacuum-tube voltmeters, a greater variation will be noted when an appreciable amount of third harmonic is present than when the waveform contains second harmonic components. The reading of the meter will always be lower for the second harmonic than the rms value. The reading of a waveform containing a third harmonic may be either higher or lower for harmonic components up to 75 percent.

Third harmonics will cause a greater variation than any other harmonic. The extremes of error with small amounts of odd harmonics are given by the percentage of the harmonics divided by the order of the harmonic. For the worst condition (third harmonic), the accuracy of an average-reading vacuum-tube voltmeter is still good. Third harmonics up to 10 percent will only cause errors up to 3.3 percent.

Voltage values which might be read on a vacuum-tube voltmeter calibrated to read rms value of a sine wave compared to the voltage as read on a peak-reading meter and the true rms value are given in the table in Fig. 22-103. The first column is the magnitude of the harmonic in terms of the fundamental voltage. The second column is the actual effective value of the voltage. This is the value which would be indicated by a thermocouple meter. The third column gives the range of readings which will be obtained using an average-reading meter. Where two values are given, these are the limits which are determined by the phase relationship of the particular harmonic. The fourth column is the indication which would be obtained with a peak-reading meter calibrated in terms of the rms value of a sine wave.

22.104 What is the percentage error when a nonsinusoidal waveform is applied to a vtvm calibrated in rms or peak voltage of a sine wave?—The expected errors are given in the following table:

Harmonic	*Error Average Meter (rms)*	*Error Peak Meter*
Second	0 to 1.8%	−20 to +20%
Third	−6.6 to 6.6%	−20 to +20%
Fourth	0 to 1.7%	−18 to +20%
Enharmonic	1.0%	20%

The values given are for the harmouic phase angles giving maximum positive and negative errors. The last line pertains to extraneous current whose frequency is not an integral multiple of the fundamental frequency. The advantage of the average-value indication is apparent, especially when measurements are to be made using ac voltage sources of questionable purity. (See Question 22.103.)

22.105 How do nonsinusoidal waveforms affect a peak-reading voltmeter?—With a peak-reading type vacuum-tube voltmeter, the limits of error can theoretically range from 100 percent low to infinitely high. Since the deflection of a peak-reading meter is proportional to the peak of the applied waveform, the maximum reading will be obtained when the relative phases of the waveform components are such that a peak of the harmonic coincides with the peak of the fundamental frequency. The maximum reading for a given magnitude of harmonic will thus be the same, regardless of the order of the harmonic.

The peak maximum value will be obtained when the peak of the harmonic is in phase opposition to the peak of the fundamental. The lowest minimum peak value will be obtained with low-order harmonics. As the order of the harmonic is increased, the minimum peak value will increase until it approaches the maximum peak value (when the fundamental and the harmonic peaks coincide). The reason that the minimum reading increases as the amount of the harmonic is increased, is that the harmonic ultimately causes neighboring peaks to be formed. These peaks are what the meter responds to and their amplitude increases with the increased harmonic. The higher the order of the harmonic, the smaller is the percentage at which these peaks form. Possible errors occur when the phases of the second or third harmonics are such as to cause the minimum possible reading of the meter. Higher harmonics will give minimum readings that progressively approach the maximum possible readings.

If the error for a peak-reading meter is compared with that of an average-reading meter, the superiority of the average-reading meter in approximating

Percent Harmonic	*Actual Rms Value*	*Average Value VTVM*	*Peak-Indicating VTVM*
0	100	100	100
10 2nd	100.5	100	90-110
20 2nd	102	100-102	80-120
50 2nd	112	100-110	75-150
10 3rd	100.5	96-104	90-110
20 3rd	102	94-108	82-120
50 3rd	112	90-116	108-150

Fig. 22-103. True rms values versus peak-indicating meter readings.

the true rms value of the waveform will at once be apparent. (See Question 22.103.)

22.106 What effect do hum frequencies have on measurements made with an average- or a peak-reading voltmeter?—The hum voltage will probably be measured with the signal voltage under measurement. If a high frequency is being measured and the hum voltage is 10 percent of the signal voltage, it will increase the reading of an average-reading meter about one-half as much as it would increase the reading of a true rms meter.

On the other hand, a peak-reading meter will add the hum voltage linearly to the desired voltage reading. Thus, the meter reading is increased by an amount approximately equal to the hum-voltage amplitude.

22.107 How may a vacuum-tube voltmeter be used as an ammeter?—By connecting a shunt resistor across the input of the vacuum-tube voltmeter and then connecting the shunt in series with the circuit in which the current is to be measured. (See Fig. 22-107.)

The voltage drop across the resistor is read on the vacuum-tube voltmeter and the current is computed by using Ohm's law:

$$I = \frac{E}{R}$$

where,
R is the shunt resistance in ohms,
E is the voltage drop measured on the meter.

The shunt resistor must be capable of dissipating the power of the circuit under measurement.

If the meter scales are calibrated in ratios of ten, and the shunt resistors are selected in ratios of ten, the current may be read directly from the meter in ratios of ten. Thus, if a 1-ohm shunt is used, 1 ampere would be indicated as 1 volt. A table of current readings for shunt resistors ranging from 0.10 ohm to 1000 ohms read on a meter scale of 1 volt is given below.

Resistance	*Current*
0.10 ohm	10 amperes
1.0 ohm	1 ampere
10 ohms	100 milliamperes
100 ohms	10 milliamperes
1000 ohms	1 milliampere

22.108 Describe a phase meter and how it functions.—A phase meter is an electronic instrument used for the measurement of phase difference between two ac signals. A precision phase meter, Model 406H, manufactured by Ad-Yu Electronics, is pictured in Fig. 22-108A. A block diagram of this instrument is given in Fig. 22-108B.

Input signal E_1 is applied to a four-stage, cathode-coupled limiter, CL11 to CL14. The function of the cathode limiter is to produce a square wave with the instants during which the waveform intersects with the zero axis identical to those of the input signal. A schematic diagram of a cathode-coupled limiter stage is shown at (a) in Fig. 22-108C.

The control grids of the duotriode tube are biased with positive potentials

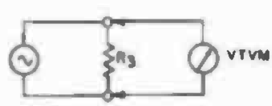


Fig. 22-107. Vacuum-tube voltmeter used as an ammeter.

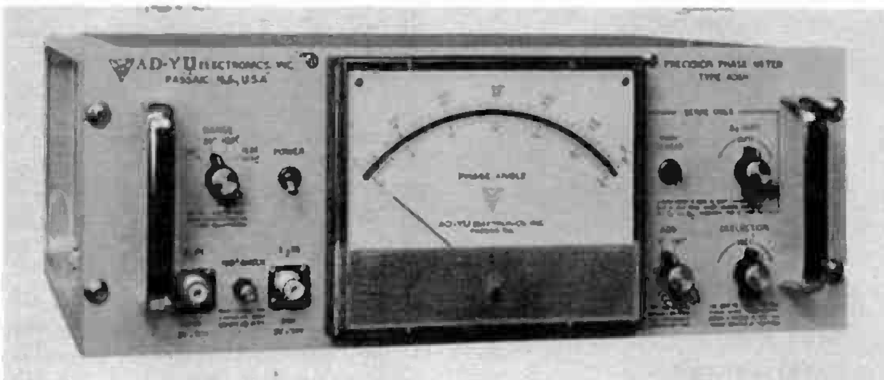


Fig. 22-108A. Ad-Yu Electronics Model 406H precision phase meter.

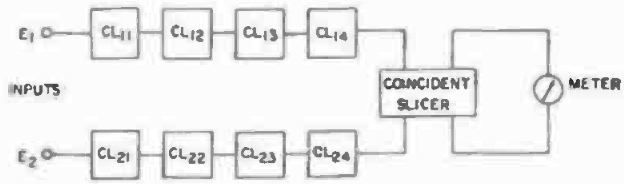


Fig. 22-108B. Block diagram for Model 406H precision phase meter manufactured by Ad-Yu Electronics.

E_{c1} and E_{c2}, respectively. When the input signal increases from zero, the potential at the cathode of both sections increases. The control grid of the second section is held constant by E_{c2}; thus the potential at the plate of the second section increases. As the signal continues to increase, the potential at the cathodes will be high enough to cut off the plate current of the second section; therefore, the waveform at the plate of the second section becomes flat-topped.

When the input signal increases negatively from zero, the cathode potential decreases at a slower rate and the plate current of the second section increases because its control grid is held constant by E_{c2}. As the signal continues negatively, the control grid to cathode potential of the first section will reach a value high enough to cut off the plate current and the output waveform becomes flat-topped again. Before limiting action occurs, the gain of a cathode-coupled limiter is approximately equal to:

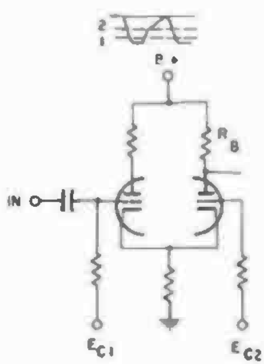
$$\frac{\mu R_B}{2(R_B - R_p)}$$

where,

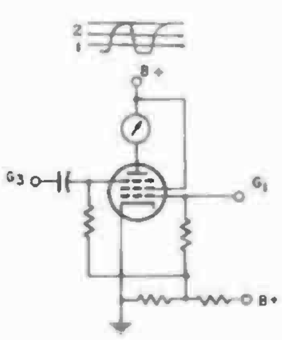
R_B is the plate-load resistance of the second section,

$_p$ is the plate resistance of the second stage,

μ is the amplification factor of the tube.



(a) Cathode-coupled limiter.



(b) Coincident slicer.

Fig. 22-108C. Basic circuits for Ad-Yu Electronics Model 406H phase meter.

With conventional duotriode tubes, a gain of 10,000 can be secured with four stages of cathode-coupled limiters. Both input signals are transferred into two square waves by means of separate four-stage, cathode-coupled limiters and then fed into a coincident slicer circuit. The circuit diagram of the coincident slicer is given at (b) in Fig. 22-108C. A gated-beam tube with two control grids, G_1 and G_3, is employed. Both control grids, G_1 and G_3, are biased positively with respect to its cathode and both applied signal voltages are negative. The plate current of the gated-beam tube cannot flow unless G_1 and G_3 are simultaneously above their cutoff values.

When both input signals are at their most negative positions, plate current cannot flow. Assume the dotted line (1) in the waveform shown at (b) of Fig. 22-108C represents the cutoff level and the dotted line (2) the saturation level of both grids G_1 and G_3. When the potentials at G_1 and G_3 are being increased from a value below cutoff, the plate current will start to flow as soon as the voltages of G_1 and G_3 are either equal to or above the dotted line (1). Thereafter, the plate current continues to increase if the potential at G_1 and G_3

continues to increase. As soon as the potentials at G_1 and G_2 reach dotted line (2), the plate current reaches its maximum value and will no longer increase. If the irregularities of both input signals, rounded corners, and overshoots are outside the region of dotted lines (1) and (2), the plate current will not be affected by such irregularities.

Fig. 22-108D shows the waveforms at various points of the circuits. E_1 and E_2 are the input signals; E_1' and E_2' are output signals of the two four-stage, cathode-coupled limiters. The rounded corners are exaggerated for low-frequency signals, but not for high-frequency signals. Dotted lines (1) and (2) represent the cutoff and saturation levels, respectively, of the gated-beam tube used in the coincident slicer. I_B represents the waveform of the plate current of the coincident slicer. The complete schematic diagram appears in Fig. 22-108E. Assume T to be the period of the applied signals and ϕ to be their phase angle. The average plate current (I_{BA}) can be expressed:

$$I_{BA} = I_{BP}\, t_1/T = k\phi$$

where,

I_{BP} is the peak value of the plate current of the gated-beam tube,
k is a constant.

Therefore the output meter can be calibrated to read the phase angle between E_1 and E_2 directly.

It is known that stray capacitance and lead inductance can produce rounded corners and overshoots on a square-wave signal. This effect is very

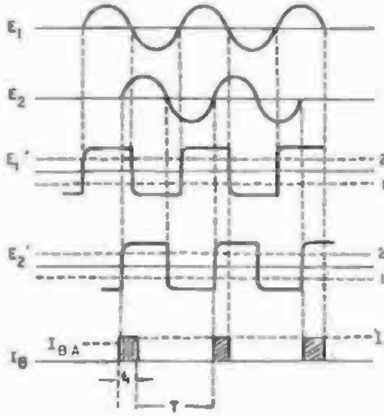


Fig. 22-108D. Waveforms at various points in the circuit of the Ad-Yu Electronics Model 406H phase meter.

serious if a sum amplifier is used for phase measurement instead of a coincident slicer as used in conventional phase meters. Another advantage of using a coincident slicer is that the output meter reading can be made virtually independent of the variations in the output amplitudes of the cathode-coupled limiters or voltages of the tubes. Since the instrument employs two four-stage cathode-coupled limiters, a gain of about 10,000 is achieved for each input channel. The output meter reading is completely independent of variations in signal amplitudes.

Used in conjunction with an accurate phase shifter, this instrument may be employed for the measurement of phase angles with an accuracy of better thau 0.25 degrees at fractions of 1-degree phase difference between two signals.

22.109 What is an in-the-circuit capacitor tester?—A capacitor tester designed to test capacitors installed in a circuit without disconnecting them. The device will test capacitors for internal short circuits and opens, even though the capacitor is connected in parallel with a resistor of 50 ohms. The schematic diagram for such a tester is shown in Fig. 22-109A. The circuit consists of two parts: one for testing shorts and the other for testing opens.

For the open-circuit test, tube V1 operates as a Hartley oscillator on an approximate frequency of 20 MHz. Inductance L2 is a coupling coil to the tank circuit of oscillator coil L1. The circuit, composed of C5, C6, L3, and the test cable, is designed to appear to coupling coil L2 as a quarter-wave line, and has the characteristics of a quarter-wave line; that is, an open circuit at one end of the line (open-circuit capacitor) appears as a short circuit at the other end, and a short circuit (capacitor with an internal short circuit) appears as an open circuit.

A good capacitor will reflect an open circuit to the junction of L2 and CR2, which in turn develops a voltage which is rectified by CR2. Therefore, when the test switch is closed, a negative voltage is applied to the control grid of V2, a 6E5 electron-ray tube, causing the eye to close.

In the short-circuit test position, if the test leads are connected to a good capacitor, the ac voltage from the filament winding of transformer T1 ap-

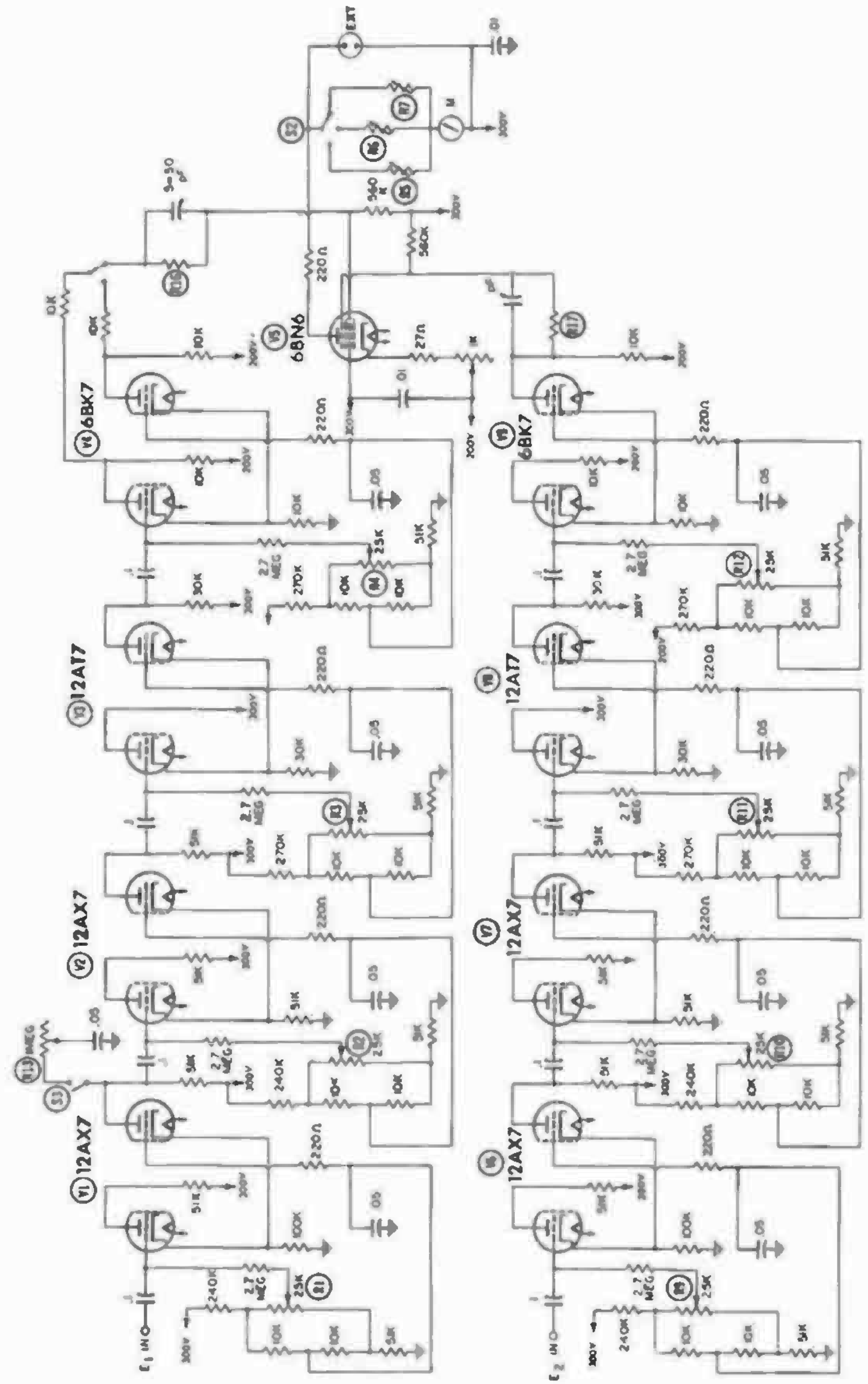


Fig. 22-108E. Schematic diagram for Ad-Yu Electronics Model 406H phase meter.

pears at the control grid of V2, causing the eye to close. If the test leads are connected to a capacitor with an internal short circuit, the ac voltage from the filament winding of T1 is shorted across resistor R1 and no voltage appears at the control grid of V2. This permits the eye to remain open.

Intermittent capacitors are indicated by a fluttering of the eye.

Any type of capacitor from 20 pF up to and including 2000 μF may be checked for shorts or an open. Capacitors of 0.10 μF may be tested with a resistor of 20 ohms in parallel. Good capacitors of less than 20 pF will indicate an open circuit.

The device with the suspected capacitor must be disconnected from the power line when making tests.

22.110 What is a direct-reading capacitance meter?—In a direct-reading capacitance meter, as shown in Fig. 22-110, the measurement of capacitance is

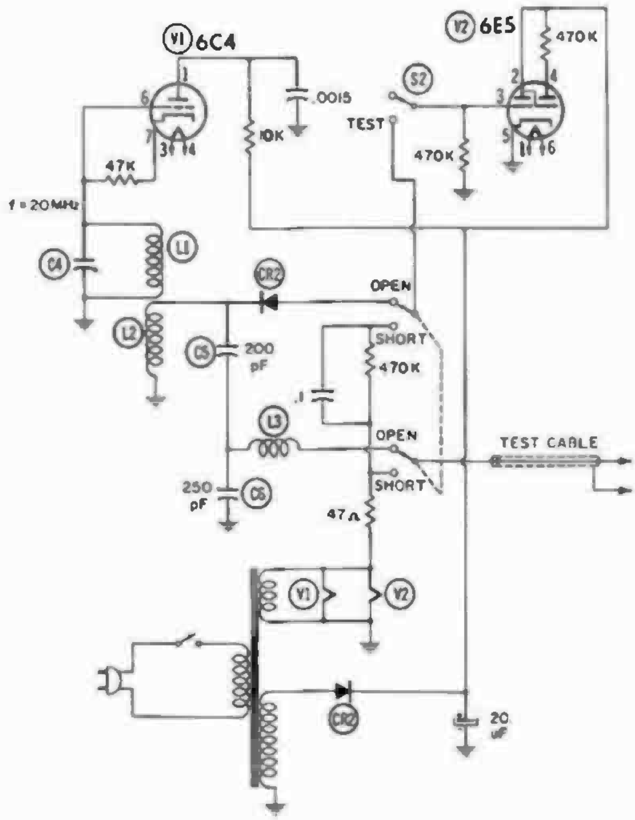


Fig. 22-109A. Schematic diagram for an "in-the-circuit" capacitor tester.

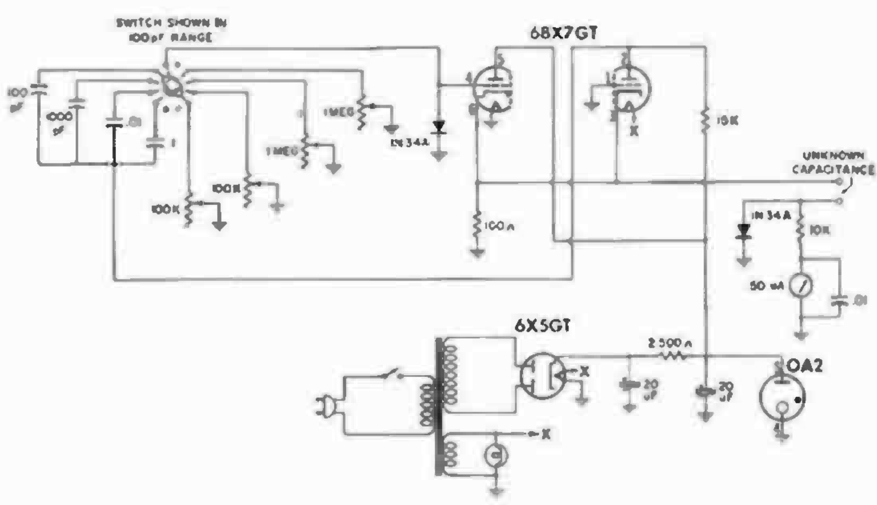


Fig. 22-110. Schematic diagram of a direct-reading capacitance meter.

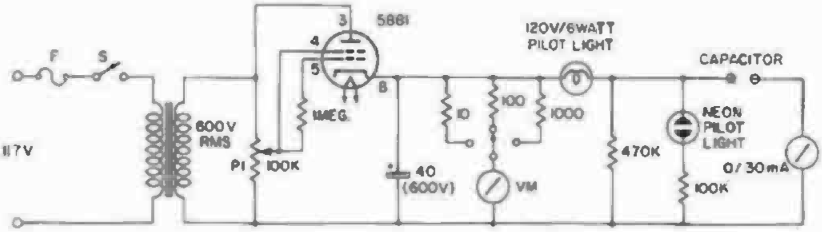


Fig. 22-111A. Electrolytic capacitor leakage tester.

accomplished by rectifying a square-sided pulse and reading the average value of the voltage on a meter calibrated in microfarads. The impedance of the meter circuit, in conjunction with the unknown capacitor, determines the shape of the pulse and, consequently, the average value of rectified voltage.

The capacitor meter consists of a type 6BX7GT tube in a cathode-coupled multivibrator circuit. A 100-ohm resistor in the cathode circuit provides a common impedance for both sections of the tube, which is necessary to maintain oscillation, and is the source impedance of the pulse used for the measurement of capacitance. A range switch permits the selection of various capacitors in a feedback loop between the plate of the second section and the control grid of the first section. Four calibrating resistors are provided, one for each range of capacitance measurement. The value of the calibrating resistor establishes the pulse repetition rate and the maximum value of capacitance that may be read. Oscillator frequencies range from 80 Hz to 80 kHz.

The pulse repetition rate is such that, at the maximum deflection of the meter for any range, the waveform of the rectified pulse will have decayed to zero during the interval between pulses. This is necessary in order to maintain linearity of the readings throughout the full range of the instrument. A conventional power supply with a resistance-capacitance filter and a OA2 voltage regulator complete the circuit. The tolerance of the internal capacitors used as a standard and the adjustment of the variable resistors determine the accuracy of the instrument. Before using the instrument, it should be permitted to warm up for about 10 minutes. The unknown capacitor is connected to the test terminals. If the approximate value of the capacitor is known, the range switch is set to the proper scale; if unknown, it is set to the highest scale and then to the next lower scale, until the proper range is found. The capacitance may then be read directly from the meter scale, noting the position of the range switch.

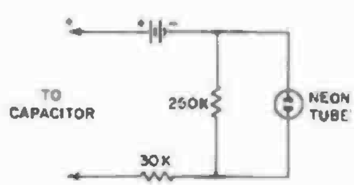


Fig. 22-111B. Circuit for testing capacitor leakage.

22.111 *Describe a capacitor leakage tester.* — The leakage tester circuit shown in Fig. 22-111A consists of a 600-volt transformer, a type 5881 tube used as a series current regulator, a series pilot lamp to protect against short circuits, a voltmeter, and a milliammeter.

The milliammeter has a range of 0 to 30 mA, and the voltmeter has three scales covering full-scale deflections of 10, 100, and 1000 Vdc. The 5881 tube is used both as a half-wave rectifier and as a series current regulator, with a potentiometer P1 used for controlling the voltage at the control grid. In this manner the output voltage may be varied from zero to full voltage across the capacitor.

In operation, the test voltage is slowly brought up from zero to the rated voltage of the capacitor, while the leakage current is kept below 10 mA, thus reducing the in-rush current and preventing the capacitor from overheating. The leakage current is read after several minutes to allow it to become normal, with allowance made for the capacitor age and ambient temperature.

If the leakage current is high, the capacitance should be measured, as electrolytic capacitors will often measure up to 60 percent higher than their rated capacitance. If the capacitance is satisfactory, the capacitor can then be considered to have high leakage. If the current fluctuates and refuses to settle down when the rated voltage is reached, it is generally an indication of intermittent leakage and the capacitor should be discarded.

The maximum permissible leakage current may be calculated as:

$$I = K \times C + 0.30$$

where,

I is the permissible leakage current in milliamperes,
K is a constant taken from the table below,
C is the rated (or measured) capacitance in microfarads.

(See Question 25.30.)

K	*WVDC*
0.01	3 to 100
0.02	101 to 250
0.03	251 to 350
0.04	351 to 500 and above.

A simplified capacitor leakage tester consisting of a battery (or other source of voltage), an NE-51 neon lamp, and two resistors is given in Fig. 22-111B. If voltage is applied to a capacitor and the current through the capacitor creates a voltage drop across the 30,000-ohm resistor causing the neon lamp to glow, leakage is present. If the leakage current is small, the lamp will remain dark. If it flashes intermittently, leakage is present, or if it glows continuously it is a sign of an internal short. When first applying the tester to a capacitor, the lamp will glow for a few seconds and then go dark.

Capacitors using oil, paper, or one of the other low-leakage dielectrics may be tested in a similar manner; however, these capacitors generally show no leakage, or may flash once in 10 seconds or so, or not at all.

22.112 What is a graphic level recorder and what is its purpose?—A graphic level recorder is an instrument for automatically plotting the electrical response of a device under test. When operated in conjunction with a sound-level meter or some other source of sound pickup, it can be used to plot the acoustical delay characteristic of recording stages, auditoriums, or for recording mechanical vibration by connecting the proper transducer to the input. The plot is made on a moving strip of paper calibrated to read in decibels, voltage, current phons, sones, or other units. It can also be used for plotting the electrical response of amplifiers, filters, equalizers, or any other device in a rectilinear or logarithmic manner, and in some instances the recorder plots in both manners. Such an instrument is ideal for routine testing of microphones and speakers in conjunction with an anechoic or quiet room. The paper strip is driven by a synchronous motor and is generally capable of moving at several different speeds.

Pictured in Fig. 22-112A is a Model 2305 level recorder, manufactured by Brüel and Kjaer of Denmark. The basic diagram for this instrument is given in Fig. 22-112B. The recorder has been de-

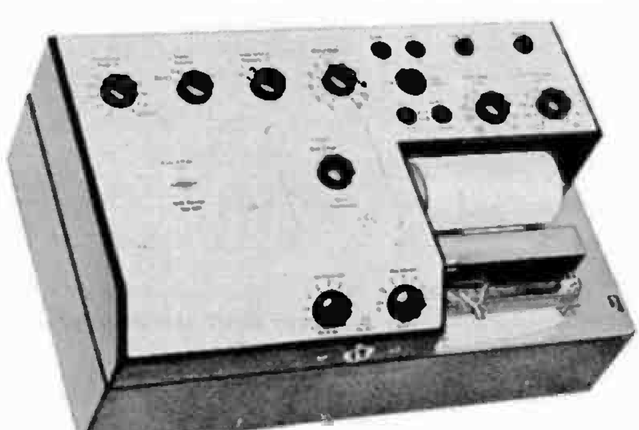


Fig. 22-112A. Brüel and Kjaer Model 2305 level recorder.

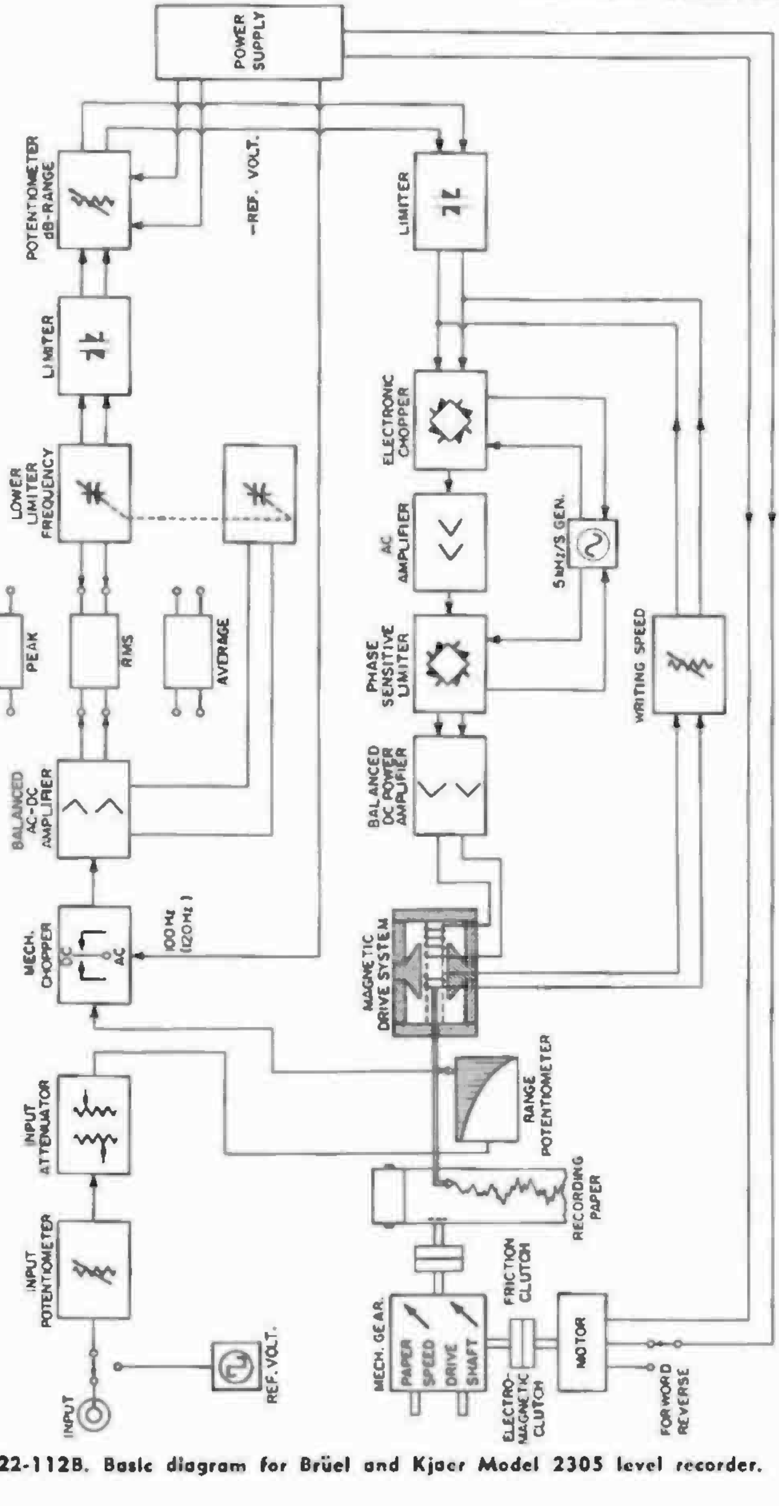


Fig. 22-112B. Basic diagram for Brüel and Kjaer Model 2305 level recorder.

signed to respond to frequencies between 2 Hz and 200 kHz.

The operation of the recorder is based on the servo principle. When the magnitude of the voltage applied to the input terminals is changed, a current will flow through the driving coil of the writing system, thus moving a stylus (pen) mechanically coupled to a range potentiometer. By the movement

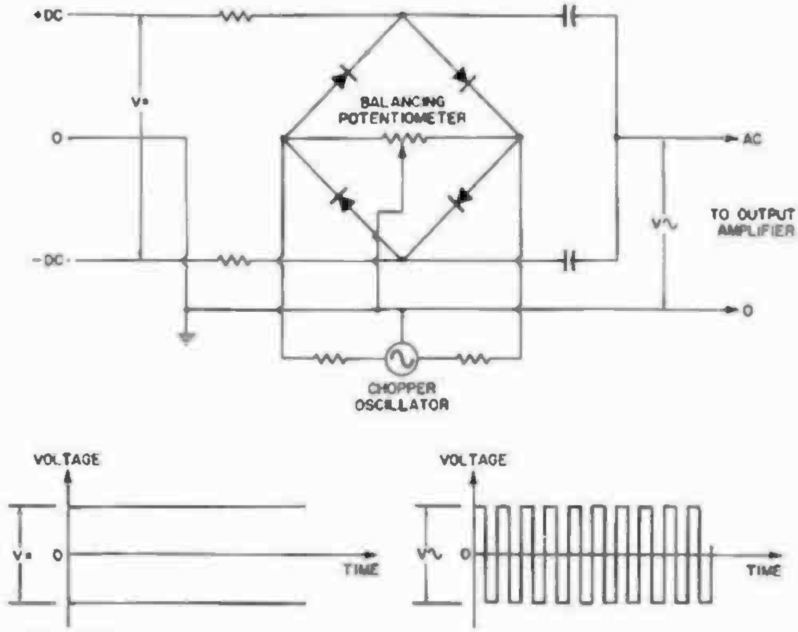


Fig. 22-112C. Basic circuit for electronic chopper used in the dc output amplifier for the Brüel and Kjaer level recorder.

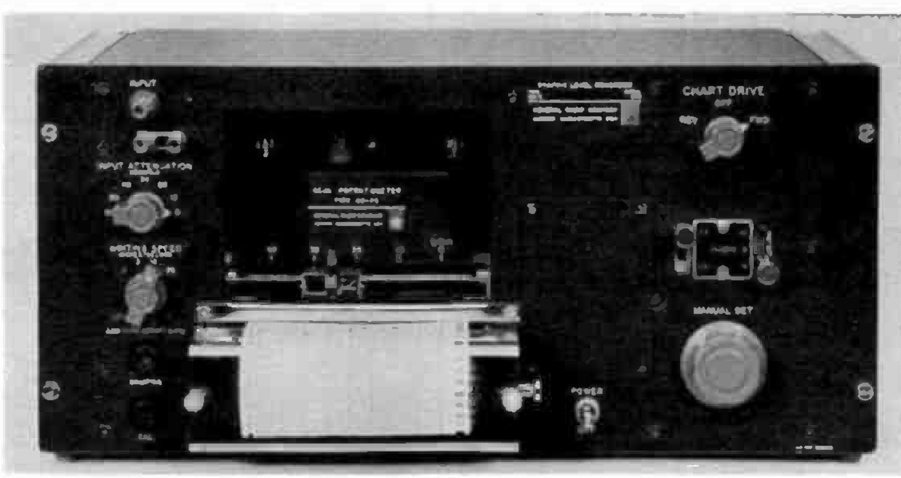


Fig. 22-112D. General Radio Co. Model 1521B graphic level recorder.

of the stylus, a voltage delivered from a potentiometer to an input amplifier is altered until a stable condition is obtained. In this way it is possible to make recordings for different ranges of input voltage by employing different range input potentiometers.

Electronically, the recorder consists of an input potentiometer, attenuator, and a direct-coupled ac amplifier. A signal rectifier and dc output amplifier drive the electrodynamic writing system. The input signal is fed to the continuously variable input potentiometer and then to a calibrated input attenuator, whose impedance is approximately constant, varying between 16,000 and 18,000 ohms. The input attenuator decreases the input signal in six 10-dB steps for a total range of 60 dB. Interchangeable input potentiometers are available for a wide range of operation for either linear or logarithmic operation.

The voltage developed between the variable arm of the range potentiometer and ground is fed to au electromechanical chopper. When an ac signal is recorded, the chopper arm is held in one of its two contact positions. However, when switched to dc, the chopper will operate at twice the line frequency, supplyiug an ac signal to the following amplifier. The nonchopped or chopped signal is applied to a balanced direct-coupled amplifier. Unwanted blocking of the amplifier due to overdriving (resulting in an overshoot in the plot) is eliminated, and the influence of the variations in supply voltage are decreased to a minimum. A balanced amplifier enables the output from the signal rectifier, which follows after the amplifier, to be symmetrical with respect to ground. By using ac as well as dc negative feedback, the lower limiting frequency of the amplifier is easily controlled. Because of the feedback arrangements, the dc amplification is approximately 270, while the ac amplification is on the order of 4500.

The rectifier can be switched to measure either true rms, arithmetical average, or half the peak-to-peak value of the input signal. The lower frequency limits of the recorder may be set for 2, 10, 20, 50, or 200 Hz. To avoid overdriving by the high-level rapidly fluctuating signals, an amplitude limiter is included. The characteristics of the limiter are such that the drive system has full power when the fluctuations reach their limit.

The output signal from the limiter is now dc and is compared with a built-in, balanced, dc reference voltage. The difference between these two signals is used to drive the output amplifier

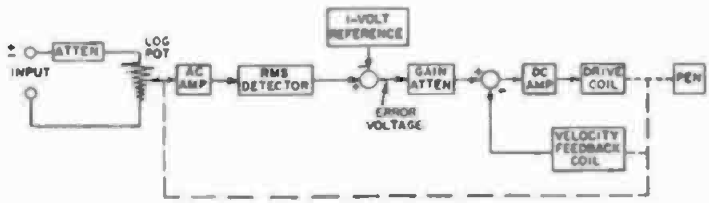


Fig. 22-112E. Block diagram for General Radio Model 1521B graphic level recorder.

Fig. 22-112F. Heathkit Model EUW-20 servo recorder.

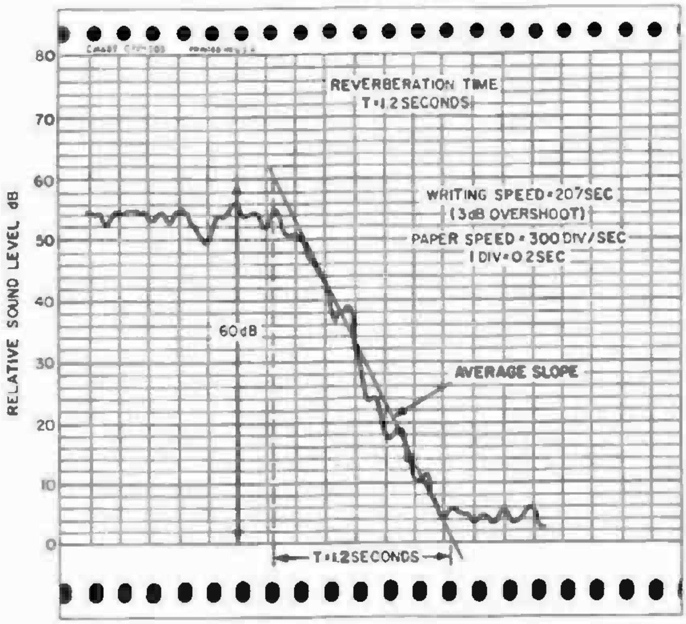


Fig. 22-112G. A typical graphic recording made by an automatic level recorder.

through a second limiter. The difference signal is obtained by comparing the rectified and limited input signal with the built-in reference voltage, which may be attenuated by a step range potentiometer. The latter element controls the resolution of the recorder. From the range potentiometer attenuator, the signal goes to a second limiter. This limiter is to ensure that the drive signal to the output amplifier is independent of the magnitude of the servo error signal, as soon as it reaches a given value.

The output consists of a chopper amplifier and a push-pull dc power amplifier. The chopper amplifier prevents zero-level drift during the amplification of very small dc signals. A balanced modulator operating at a frequency of 5000 Hz is used for the chopper. After the chopper signal is amplified, it is rectified by a phase sensitive rectifier, then used for controlling the dc power stage. This latter stage operates a single-ended push-pull stage, thus permitting one terminal of the drive coil in the electromagnetic recording system to be grounded.

A velocity-dependent signal is developed from a special winding in the electrodynamic drive system. This signal is introduced as a negative-feedback signal to the input of the output amplifier. Thus the feedback signal controls the output signal from this section as long as the signal from the second amplitude limiter has a constant value, which is the situation when the limiter is in operation. In this manner constant speed of the writing system is obtained. The speed and thereby the damping (average time) of the writing system is adjusted by the writing speed control. The basic diagram for the chopper circuit is shown in Fig. 22-112C.

The power supply section contains two full-wave and two half-wave rectifiers. One full-wave rectifier supplies dc for the magnetic clutch, the event-marking arrangement, and the lifting magnet. An ac ripple voltage is used to drive the electromechanical chopper in the recording input circuit. The second full-wave rectifier supplies the plate voltage to the amplifiers, with the exception of the tubes in the dc output stage. The positive reference voltage for the servo system is also supplied from the rectifier and stabilized by a zener diode.

The maximum sensitivity of the recorder is 5 mVac and 10 mVdc, with a dynamic range of 10 to 75 dB. The range potentiometers are of the plug-in type and may be changed to fit a particular requirement. The input impedance is 16,000 to 18,000 ohms. Writing speed is 2 to 1000 mm/s for 50-mm paper, and for 100-mm paper 4 to 200 mm/s. There is an internal 60-Hz calibrating signal.

Means of coupling the recording mechanically to other instruments such as an oscillator or analyzer are also provided, as well as remote control of the principal operating controls. Standard chart paper may be used, either of the ink- or wax-surface type.

Fig. 22-112D shows a Model 1521B graphic level recorder, manufactured by the General Radio Co. This recorder is somewhat similar to the one previously described. Its basic diagram appears in Fig. 22-112E. The input signal is fed through an input step attenuator to a potentiometer which is automatically positioned to maintain a constant 1-mV signal at the slides arm. This voltage is then amplified and rectified by a quasi-rms detector. The detector output voltage is compared with a 1-volt reference, and the difference (error) voltage is amplified by a dc amplifier. The push-pull output current from the dc amplifier passes through a drive coil suspended in a magnetic field. Interaction between the coil current and the magnetic field moves the arm on the potentiometer to reduce the error voltage to zero (null condition), and also positions the pen mounted on the coil assembly. Since the potentiometer output is a constant 1 mV at null, the attenuation of the input potentiometer is directly proportional to the level of the input signal. By suitable shaping of the potentiometer winding a linear scale in decibels is achieved.

A feedback voltage proportional to velocity is subtracted from the error signal at the input to the dc amplifier. This voltage provides damping so that the drive coil will not oscillate and varies the servo bandwidth and maximum writing speed. The moving coil then responds to the changes in the input level of a voltage applied to the recorder, and a pen fastened to the coil traces out these changes on paper. For instance, if the input level increases by 10 dB, the drive coil (along with the pen) moves the potentiometer arm to restore the voltage to about 1 mV. Since the potentiometer is logarithmic, the potentiometer arm moves downscale 10 dB, indicating a level change of 10 dB directly on the chart paper. The recorder may be used for either ac or dc input signals. The input impedance is 10,000 ohms and has a sensitivity of 20 to 80 dB, depending on the input potentiometer employed. The frequency response is 7 Hz to 200 kHz, with writing speeds up to 20 ips for a 0.10-inch overshoot. Paper speeds are 2.5 to 75 ips. The direction of the paper drive may be reversed when necessary. Provision is also made for mechanically connecting the instrument externally to a wave analyzer, oscillator, or other device. (See Fig. 22-51E.

Fig. 22-112F shows a Model EUW-20 servo recorder, manufactured by Heathkit. This instrument is similar in its basic design to the graph recorder just discussed, except that it is driven by a dc signal. It may be employed for the measurement of speeds, pressure, temperature, strain, light radiation, and many other uses. The reference voltage is obtained from a standard 1.35-volt mercury cell. An ordinary cartridge-type fountain pen is used.

A typical recording made on an automatic graphic recorder is shown in Fig. 22-112G. (See Question 22.140.)

22.113 Describe the basic design of oscilloscope cameras. — Oscilloscope camera frames are designed in two basic styles: the manually operated type which can be used for both moving and stationary waveforms, and the automatic which can be used either manually or tripped externally by the signal for photographing transient waveforms.

One of two means of photographing the image is used. The most common is the Polaroid Land camera, which processes the picture in 10 seconds, or a 35-mm photographic film camera, which requires development after removal of the film from the camera. As a rule, the camera frame is designed to use either type of camera by means of an adapter. The front end of the camera is held to the oscilloscope by studs or a clamping ring attached to the bezel on the oscilloscope. Images photographed, using a Polaroid Land camera are photographed from the CRT without reversal.

In certain designs an ultraviolet (UV) light is included for illuminating the internal graticule of a CRT. The UV light excites the phosphor on the CRT face, causing it to glow uniformly over its entire surface, appearing in the photograph as an intermediate gray, contrasting with the black graticule lines and the white image display. The intensity of the UV illumination is variable and does not degrade the intensity of the trace. Ultraviolet light produces a twofold increase in the film speed by presensitizing (prefogging) the film at the time the image is photographed. Ultraviolet light peaks in the 3560-angstrom region, and the uniform glow of the excited phosphor lowers the apparent threshold sensitivity of the film. This enables it to record dimmer traces, thus sharpening the image of both repetitive and single sweeps. (See Question 18.8.)

The efficiency of a given phosphor screen can be quite important when fast single or low repetition sweeps are to be photographed. Phosphor type P-11 is the most suitable and has the highest photography efficiency and should be used when maximum writing speed is desired. (See Question 22.83.) Other phosphors in their descending order of efficiency are: P-31, P-2, P-7, and P-1. Ultraviolet light cannot be used to excite phosphor P-1.

Graticule filters should be removed for photography, as they detract from the quality of the photograph and require a longer exposure. Use of an external graticule increases the effects of parallax distortion. The exception would be the use of a blue filter to eliminate cathode glow on some type of nonaluminized phosphor CRT's. Blue filters may also be used to eliminate yellow afterglow of long-persistence tubes.

If the use of an external graticule is necessary, the distance separating the phosphor and the graticule causes varying degrees of parallax. This effect can be minimized only by positioning the graticule as close to the phosphor as possible. The face of the CRT should be in contact with the graticule, with no filter between them. For maximum accuracy of measurement, only the scale graduations near the center are used.

Because white light is used for graticule illumination, whereas the trace fluorescence is colored, matching of the trace and graticule intensities should be determined by evaluation of prints rather than by observation of the display. Best results are usually obtained with the trace appearing to the eye as slightly dimmer than the graticule.

Shown in Fig. 22-113A is a Model 197A oscilloscope camera, manufactured by Hewlett-Packard. The camera employs an electronic shutter-control circuit to provide exposure times from 1/30 to 4 seconds. The shutter may be operated electrically from a remote source. A sync-output connection provides synchronization for use with external equipment. Focusing is accomplished by using a split-image focusing plate in the camera. Normally the camera is used with a Polaroid Land camera film-pack. However, this may be removed and replaced with a 4 × 5 Graflex back. The back can be moved vertically through eleven positions for

Fig. 22-113A. Hewlett-Packard Model 197A oscilloscope camera.

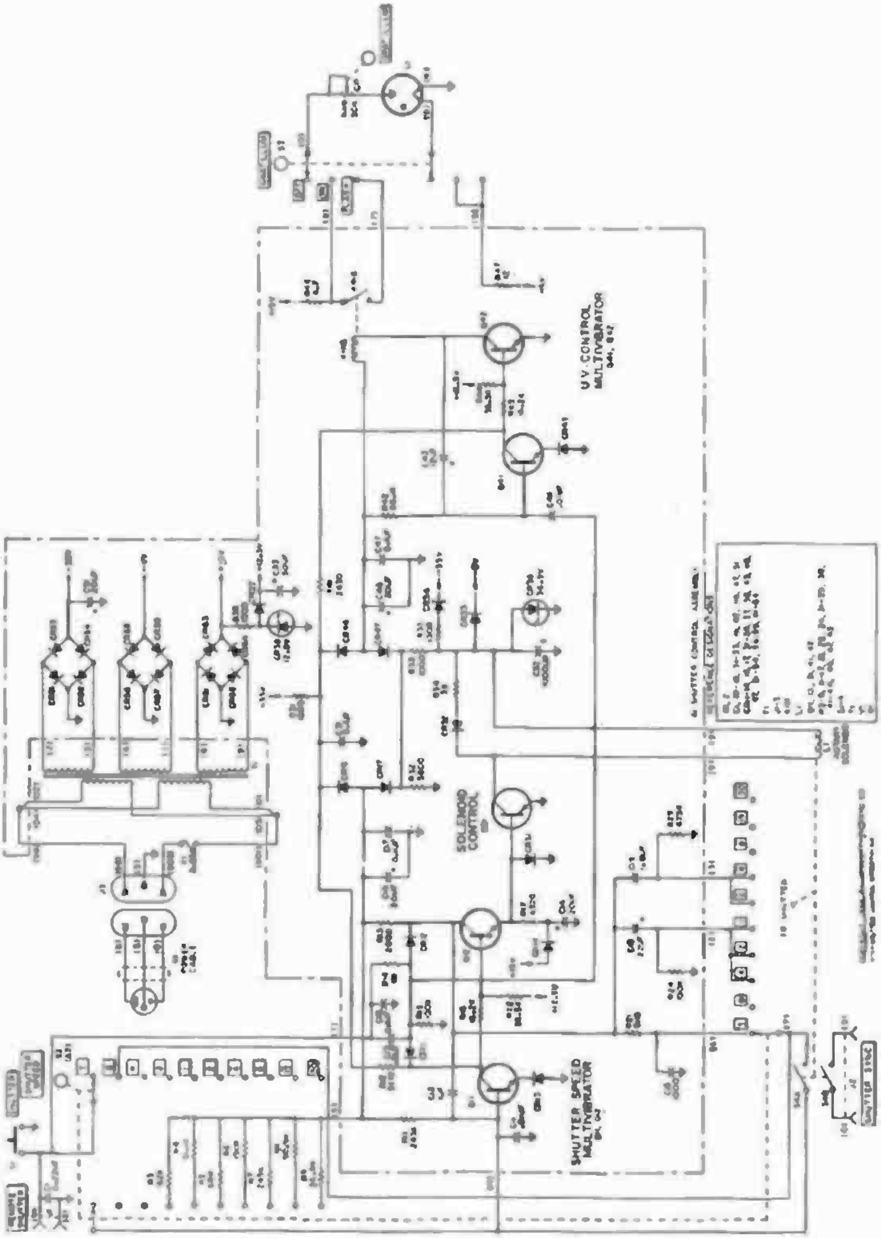


Fig. 22-113B. Schematic diagram for Hewlett-Packard Model 197A oscilloscope camera with electronic shutter control.

multiple exposures, or it may be rotated from the horizontal position to vertical, permitting several smaller pictures to be taken on one photograph. Polaroid film processing starts automatically when the print is withdrawn from the camera and it takes 10 seconds. The camera frame can be swung to one side when not in use.

Referring to the schematic diagram of the electronics in Fig. 22-113B, shutter timing is controlled by a shutter-speed multivibrator, transistors Q11 and Q12, a nonstable multivibrator circuit. Initially, with the shutter closed, Q11 is biased on, and, due to the cross coupling, Q12 is biased off. When the shutter switch S1 is closed, Q11 is turned off and Q12 is turned on, thus causing the shutter to open. The time the shutter is held open is determined by the time constant of resistor R11 and ca-

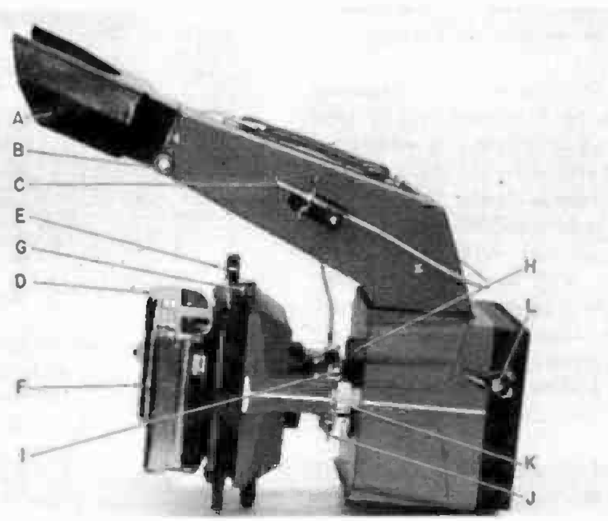


Fig. 22-113C. Tektronix Inc. Model C-12 oscilloscope camera system.

pacitor C12. Various shutter speeds are established by changing the time constants. This is accomplished by shutter-speed switch S2, which connects additional resistance in parallel with R11 and, for lower speeds, through switch S4A for additional capacitors in parallel with C12.

The shutter is operated electrically by the action of solenoid L1 and closed by a spring action within the solenoid. Solenoid L2 is actuated when solenoid control transistor Q31 is turned on by the output signal from Q12. Solenoid L1 is initially actuated by a 35-volt charge stored in capacitor C32, which ensures a fast and positive action of L1. The charge on C32 is established by the action of diodes CR34 and CR35 and takes approximately 2 seconds to reach full potential. When L1 is actuated by the discharge of C32, the voltage decays very quickly, but L1 is held the required time by the 6 volts supplied through diode CR33. Shutter action times are: 5-ms start to open, half-open 8.5 ms, and fully open 12-ms. These delays are caused by the mechanical action of the shutter. The shutter may also be operated from a remote source, and a sync-output circuit provides a contact closure when the shutter is operated in synchronization with other equipment. Exposure times range from 1/10 second to 4 seconds.

Fig. 22-113C shows a Model C-12 oscilloscope camera system manufactured by Tektronix, Inc. The camera frame mounted on the oscilloscope is of the lift-on type. The whole unit may be removed or swung to one side when not in use. The camera mount is designed to accomodate a Polaroid Land camera, or film-pack, and may be rotated through nine detented positions, or increments of 90 degrees. A focusing control is provided and, when once adjusted for a particular oscilloscope, requires no further attention.

Pictures are taken directly from the CRT screen through a beam-splitting mirror. The major components are viewing head A, knob for opening viewing door B, shutter control C, Polaroid Land camera D, slide indexing lever permittiug the camera to be rotated to nine different positions in 90-degree increments E, camera back (Polaroid) F, rotating slide latch G, locknut for holding rear portion of camera frame to front portion K, and mounting latch L (locks camera frame against graticule for lighttight seal).

By use of an adapter a standard 35-mm film camera may be substituted for the Polaroid camera. Automatic operation of the shutter mechanism is obtained by the use of a shutter actuator.

22.114 What is an emission-type tube tester?—A type which connects all elements such as the plate, screen grid, suppressor grid, and control grid together and uses them as a diode to check the electron emission from the filament or cathode circuit. The indicating meter is generally calibrated to read Good or Bad. This reading is often referred to as an English-reading tube tester.

22.115 Describe a dynamic mutual-conductance or transconductance tube tester.—It is a tube tester in which the test circuits are such that the test conducted ou a tube is similar to its normal operating conditions. The test is made by applying a small 60-Hz signal to the control grid, with the other elements operating under somewhat normal conditions. The transconductance (mutual conductance) is measured and indicated directly on a meter calibrated to read in both transconductance and English Good/Bad.

Fig. 22-115A. Dynamic mutual conductance tube tester Model 9-66A manufactured by Stark Electronic Instruments (Canada).

A group of switches provides a means of selecting the tube elements and the proper circuitry for the particular tube under test. Dials permit the bias, plate, and heater voltages to be individually selected, with push buttons for making the test.

Two full-wave internal power supplies provide dc for the tests. A tapped winding on the power transformer supplies heater voltage ranging from 0.6 to 117 Vac. The line voltage is indicated on the meter by actuating a push button, and it is adjusted by a variable control in the primary of the power transformer.

Several tube sockets of different configurations permit most of the conventional tubes to be tested. Generally, a roll-type tube chart indicating the switch settings and voltages is included in the instrument. In addition to the tests mentioned previously, an element short test, gas, pilot-light, ballast-tube, and, in some designs, elementary transistor and diode tests are included.

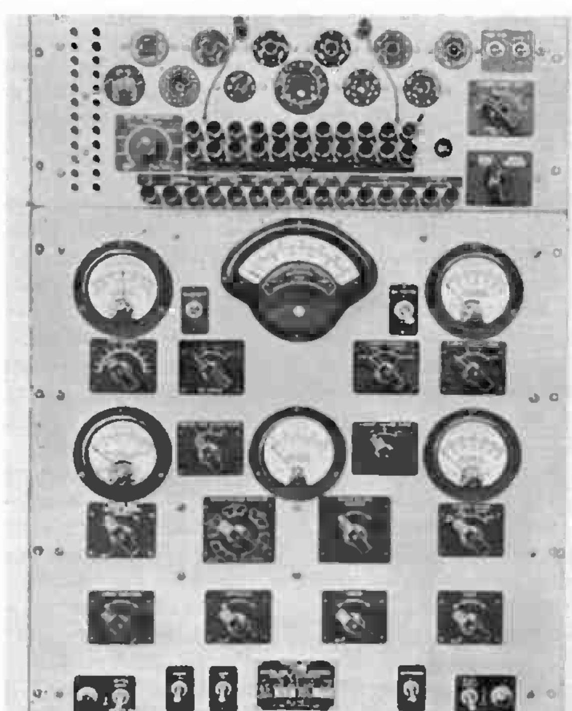


Fig. 22-115B. The Weston Electrical Instrument Corp. Model 686 laboratory-type vacuum-tube test panel.

A typical dynamic mutual-conductance tube tester, manufactured by Stark Electronics Instrument (Canada) is pictured in Fig. 22-115A. A laboratory-type testing panel is shown in Fig. 22-115B. This latter instrument is designed to provide means of duplicating the tube manufacturer's specifications for a particular tube. Each individual element voltage may be set for a given condition and the transconductance measured. The socket configuration is selected by the use of small patch cords. Many tests not available on the conventional tube tester are possible on this tester.

22.116 What is a signal tracer?—An instrument consisting of a high-gain audio-frequency amplifier, speaker, and test probe. It is used for tracing a signal in audio-freqnency recording and reproducing equipment when trouble-shooting.

A signal is applied to the device under test and the probe of the signal tracer is applied to various points in the signal path. If the device is functioning up to a point, the signal will be heard on the signal tracer speaker. When the point is reached where the signal disappears, indicating the trouble spot, the signal will no longer be heard on the signal tracer. Usually other tests are provided besides the listening test.

The circuit of a typical audio-frequency signal tracer is shown in Fig. 22-116. The probe of the signal tracer has a crystal diode in series with the input lead. This will permit the instrument to be used in checking rf devices when necessary.

22.117 What is an audio analyzer?—A test instrument used by motion-picture-theater sound engineers. The device contains terminations of various values capable of carrying up to 150 watts of audio power. Included is a vacuum-tube voltmeter for measuring the frequency response of amplifiers, equalizers, filters, and the overall frequency response of the sound system. The vacuum-tube voltmeter is employed to measure the signal-to-noise ratio of the system and is also a part of a self-contained harmonic distortion analyzer.

The vacuum-tube voltmeter measures both ac and dc and is of sufficiently high input impedance that it may be used for measuring photocell voltages and the voltage at different points in a resistance-coupled amplifier stage.

The audio oscillator circuit contains an attennator calibrated in decibels for making gain/frequency measurements. Included are circuits for measuring resistance, inductance, capacitance, and leakage of paper and electrolytic capacitors.

22.118 What is a "Z" angle meter?—An electronic instrument used for measurement of impedance and phase angle in ohms and electrical degrees.

22.119 What is a vacuum-tube bridge?—A device which makes possible the measurement of the low-frequency dynamic coefficients of vacuum tubes and transistors over a wide range of values and under a variety of operating conditions.

The device is constructed so that the circuits are independent direct-reading measurements of the forward and reverse voltage amplification factor, resistance, and transconductance. Interelectrode capacitance is balanced out in such a manner that correction factors are unnecessary.

The instrument is strictly a laboratory device used in the development and research of vacuum tubes. The device is constructed to permit setting up of special circuits for the measurement

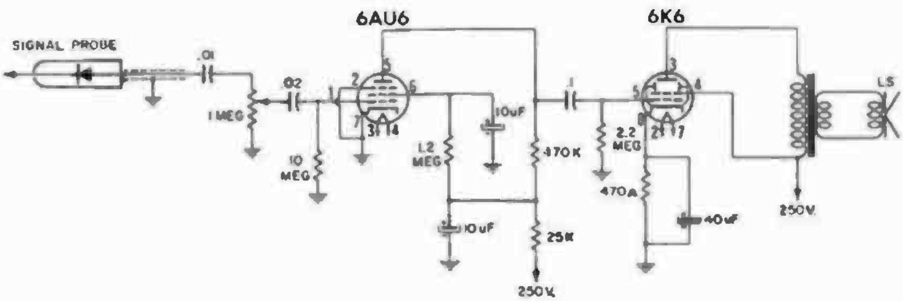


Fig. 22-116. Schematic diagram of audio frequency signal tracer.

Fig. 22-120A. General Radio Co. Model 1531A Strobotac. This instrument may be used for measuring the speed of rotating machinery and for observing irregularities in the motion.

of vacuum-tube parameters of conventional and nonconventional design.

22.120 What is a strobe light?—It is a device containing a high-intensity flashing light which is used to illuminate a moving object. The light flashes may be adjusted to a frequency equal to the motion or rotation of the equipment nnder observation, or to a multiple or submultiple of its period. When the time between the light flashes is a multiple of the period of the moving object, the object appears to be at rest. In this manner, irregularities in the motion may be studied. In certain designs the speed of the rotating object may be read directly from a calibrated dial.

The device pictured in Fig. 22-120A is a stroboscope manufactured by the General Radio Co. and is so designed the flashing light can be synchronized with any fast repeating motion, so that the rapid movement appears to be at rest.

To understand how the device stops or slows down motion, consider a fan rotating at 1800 rpm and a light that is switched on and off at the rate of 1800 times per minute. Since the light-flash frequency is the same as the fan rotation, every time the light flashes the fan blades are in the same position they were the last time the light was on. Thus the fan blades appear to be standing still. Because the retina of the human eye holds the image until the next image appears, there is little or no flicker.

If the light is switched on and off at 1801 times a minute (the fan turning at 1800 rpm), it is flashing slightly faster than the fan is turning, therefore each time the light is on, the fan blade has not quite reached the last position and the blade is seen at progressively earlier parts of its cycle and will appear to be slowly turning backward. If the light flashes 1799 times per minute, the fan blade will appear to be slowly moving forward.

If the flashing rate of the lamp is known heforehand, it can be calibrated to read directly in rpm. The practical significance of the slow-motion effects is that since it is a true copy of the high-speed motion all irregularities such as vibration, torsion, chattering, and whip are present and can be studied. The schematic diagram for a Model 1531A Strobotac electronic stroboscope is given in Fig. 22-120B. The flashing rate is controlled by an internal oscillator (V1) which constitutes a bistable circuit. In such a circuit one section conducts while the other section is off, and vice versa. Each section is alternately turned on and off at a rate determined hy the value of the resistors and capacitors in the circuit and the voltage setting of resistor R3, the rpm control. Range switch S2 introduces the proper-value timing capacitor into the circuit to increase or decrease the flashing rate by a factor of six.

The output of oscillator V1 is applied to a thyratron tube (V2) through capacitor C8. The thyratron together with C9 and pulse transformer T2 produces high-voltage pulses for firing the flash tube. The high-voltage output from the trigger circuit is capacitively coupled from T2 to the strobotron tube, V3. The coupling capacitance is between pins of the strobotron socket and a brass slug in the center of a ceramic insulator. The energy to flash the strnhotron is obtained from the discharge of capacitors C10, C11, and C14. The correct value of capacitance for each rpm range is con-

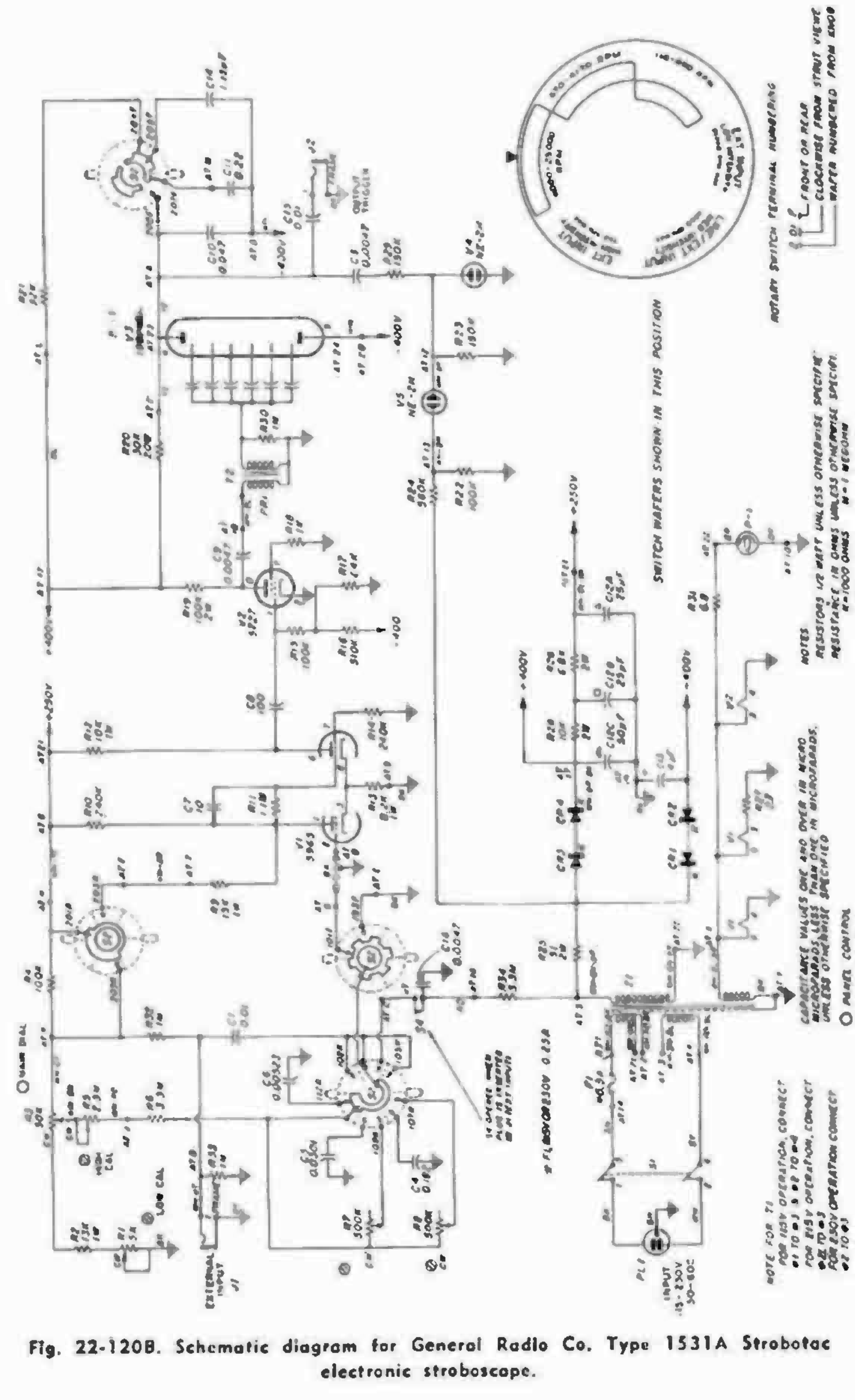


Fig. 22-120B. Schematic diagram for General Radio Co. Type 1531A Strobotac electronic stroboscope.

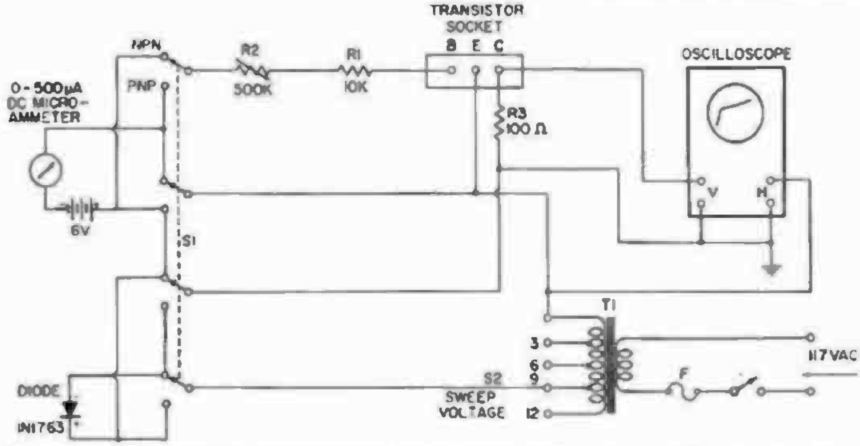


Fig. 22-121A. Transistor curve tracer for both pnp and npn types. The circuit polarities are selected by means of switch S1.

nected across the strobotron by the range switch. After each flash the active capacitors are recharged to 800 Vdc. The unused capacitors are kept charged at the same voltage to reduce arcing at the switch contacts when the range switch is changed.

The power supply is a voltage doubler, furnishing plus and minus 400 Vdc, and plus 250 Vdc. For operation with an external synchronizing source, the oscillator circuit is converted automatically to a conventional amplitude-sensitive bistable circuit. The strobotron can be flashed from 110 to 25,000 flashes per minute, over three ranges. Rotating speeds up to 250,000 rpm can be measured. The flash duration is approximately 0.8, 1.2, and 3 microseconds, with a peak light intensity of the high-, medium-, and low-speed ranges of 0.21, 1.2, and 4.2 million beam candlepower measured at a 1-meter distance from the center of the beam. For a single flash the intensity is 7 million beam candlepower at the same distance.

22.121 ***Give the circuit for a transistor curve tracer.***—The design of transistor amplifiers and similar equipment requires a combination of graphical and mathematical analysis, the former being taken from the transistor characteristic cnrves supplied by the manufacturer. The most useful of these curves is the I_c/V_{ce} or common-emitter curves.

With the curve tracer such curves may be easily and quickly displayed on the face of an oscilloscope. Tracing transistor characteristics on an oscilloscope prevents overheating the transistor as would occur if the characteristics were plotted point by point. Referring to the schematic diagram of Fig. 22-121A, series resistor R2 controls the base current (I_b) supplied by a 6-volt battery. Resistor R1 is a swamping resistor connected in the base to limit the base current. The sweep voltage is supplied by transformer T1. The 1N1763 diode permits only the negative (for a pnp transistor) half of the ac voltage to be applied to the collector circuit, resulting in a voltage drop across R3 proportional to the collector current. Thus the changing values of the collector voltage trace the I_c/V_{ce} characteristic on the oscilloscope.

When the cathode of the diode is positive (reverse biased) the collector current is cut off, resulting in no deflection of the oscilloscope trace. However, when the cathode of the diode is negative for the other half-cycle, the diode is forward biased and the voltage is supplied

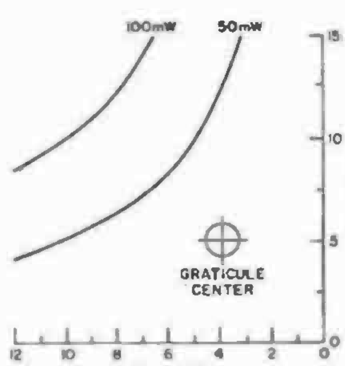


Fig. 22-121B. Graticule for transistor curve tracer.

to a collector element causing a current and starting the horizontal sweep. For measuring a pnp-type transistor the polarity of the 1N1763 diode is reversed.

Resistor R3 is a 1-percent precision resistor. Since the current range in the common-emitter configuration will range from 20 to 300 milliamperes, a 500 microampere meter is required.

A plastic graticule (Fig. 22-121B) is prepared and mounted on the oscilloscope. The X axis of the oscilloscope (horizontal) is calibrated to read 0 to 10 volts peak-to-peak, and the Y axis (vertical) reads from 0 to 10 milliamperes peak-to-peak. Curves indicating the power dissipation in milliwatts are also included for easy reference. The curves for the graticule should match those shown; however, they may be made any size as long as the ratios are maintained.

After the graticule is placed on the oscilloscope face, the vertical input control is calibrated by applying 1 volt peak-to-peak (60 Hz) to the input and adjusting the deflection to match the 10-milliampere graticule scale. The same is done for the horizontal deflection, using 10 volts peak-to-peak. Both controls are then left in position. The oscilloscope positioning controls are then adjusted to bring the start of the transistor curve to the graticule starting point.

Transistors with internal short circuits will be indicated by a diagonal straight line, while an open circuit will show only a straight line aloug the 10-volt axis. Although the described tracer can be used with any oscilloscope, if a long-persistence or storage type is available, several different traces may be run and the image retained for study.

22.122 ***Describe the basic circuit and operation of a gain set.***—A gain set, or *transmission set* as it is sometimes called, is a device used for measuring amplifier gain or loss, the frequency characteristics of filters, equalizers, amplifiers, recording and reproducing circuits, and similar devices. A gain set may be used to supply a signal of known magnitude in voltage or decibels from a given impedance with respect to a given reference level.

Gain sets may be designed to use either one or two VU meters. The two-meter instrument is both easier and faster to use, as both the send and receive level can be viewed simultaneously during a measurement, with certain precautions.

A typical two-meter gain set is pictured in Fig. 22-122B, manufactured by the Daven Co., designed for laboratory, radio-network, recording-studio, and general routine maintenance work.

Basically a gain set consists of a group of fixed and variable attenuators calibrated in decibels, one or more VU meters, a group of switches for selecting various sending and receiving impedances, attenuators for the VU meters, and jacks for internal and external connections with the gain set components. An external oscillator supplies the test signal.

The schematic diagram of Fig. 22-122B is that of a single-meter unbalanced gain set, consisting of two heavily shielded sections (A and B), with C the send-receive switch, and D the VU meter attenuator network. The reason for the use of the term "unbalanced" is that unbalanced attenuator networks are used with one side at ground potential (balanced gain sets are discussed in Question 22.123). The sending im-

Fig. 22-122A. The Daven Co. Model 6C two-meter transmission or gain set.

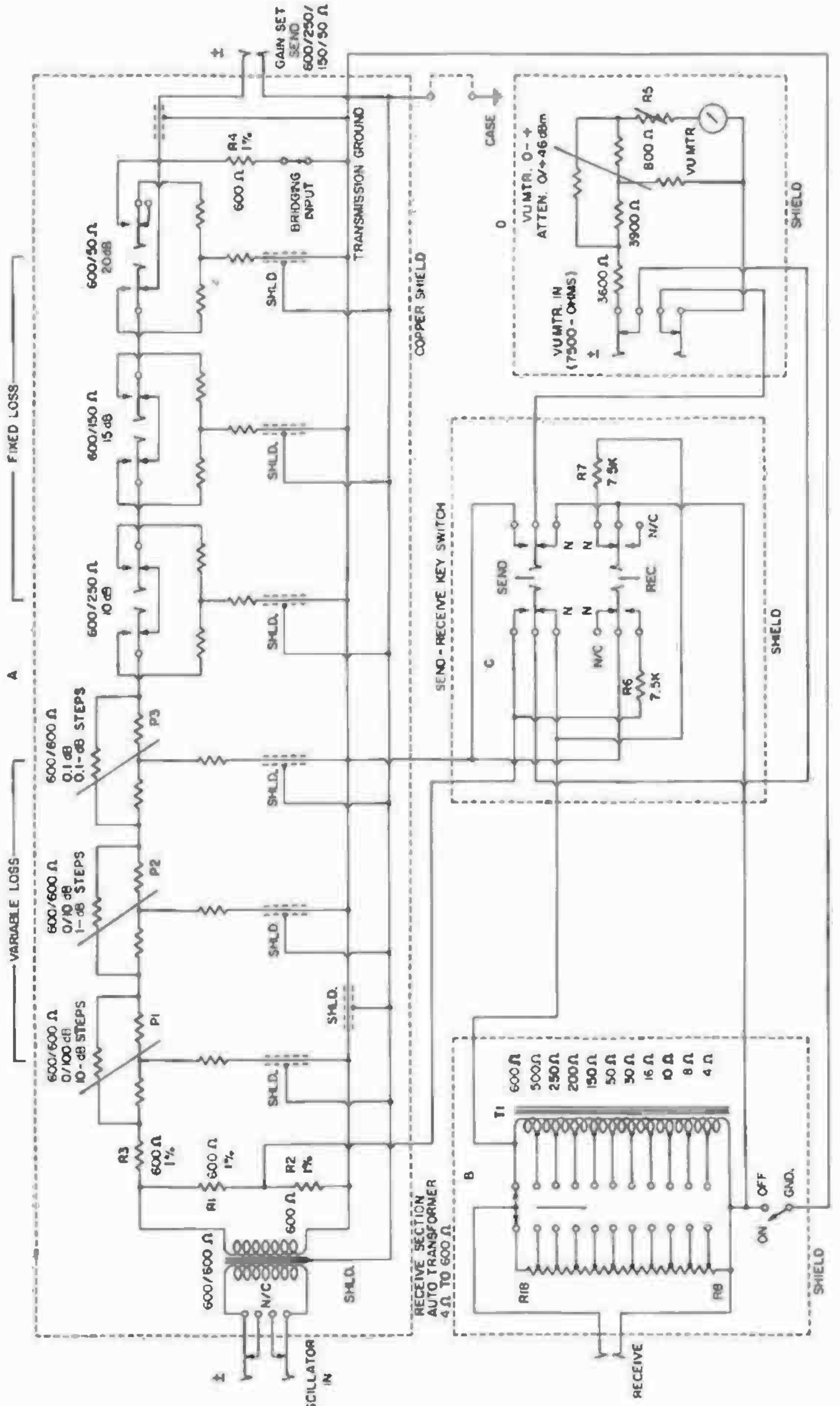


Fig. 22-122B. Schematic diagram for a single VU meter gain set.

pedances for this particular instrument are 600, 250, 150 and 50 ohms, although any impedance may be matched by patching an external pad with the correct impedance ratio (600 ohms to the desired impedance).

A brief description of how a gain set is used will aid in the understanding of the circuit diagram. Suppose that a frequency-response measurement is to be made on an audio amplifier. The attenuators in the gain set are set to their maximum loss position before the amplifier is connected. The amplifier is grounded to the system ground. The send terminals of the gain set are connected to the input terminals of the amplifier, and the receive section of the gain set is connected to the output terminals of the amplifier. The send and receive sections of the gain set are set to match the amplifier input and output impedances.

Next, the VU meter key switch shown in Section C of the diagram is thrown to the send position. The audio oscillator is set to 1000 Hz and its output control advanced until the VU meter reads a plus 4 dBm or 100 percent. The point of 100 percent on the meter will be called the *zero-level point* and used as a reference.

The VU meter key is now thrown to the receive position and loss slowly removed from the attenuators until a reading of zero level (100 percent) is obtained. When the same level is obtained for both the send and receive positions, the total loss of the attenuators equals the gain of the amplifier at 1000 Hz. This is true because the oscillator signal-level voltage has been reduced by the attenuators at the input of the amplifier to a level equal to the voltage gain of the amplifier. As an example, if the input and output impedances are correct and 40 dB of loss in the attenuators is required to obtain equal readings on the VU meter, the gain of the amplifier equals the loss of the attenuators. The same operational procedure is used for all frequencies measured. The input signal level (send) is always set to zero and the attenuators adjusted for a similar reading in the receive position. Fractional parts of a decibel are read on the right side of the meter zero and added to the attenuator losses. If the gain set includes a 0- to 1-dB attenuator calibrated in 0.10-dB steps, the receive meter is brought to zero and then the losses added.

Now, again referring to the diagram in Fig. 21-122C, the output of the oscillator is fed through normal jacks, an isolating transformer, and a building-out network composed of resistors R1, R2, and R3, which serve two purposes. First, the network isolates the output impedance of the oscillator from the attenuators, thus preventing an impedance mismatch when only small amounts of loss are inserted in the attenuators (one or two dB). Second, the network serves as a level-correction network for the VU meter. The three network resistors comprise an L-type attenuator network. Looking into the oscillator terminals of the gain set, resistors R1 and R2 are seen in series and equal 1200 ohms. Looking from 600-ohm variable attenuator P1 toward the oscillator, R3 is in series with attenuator P1, which is also 1200 ohms and matches the network of R1 and R2.

Under the foregoing conditions, there will be a 6-dB or 50-percent voltage drop between the output of the oscillator and attenuator P1 across R3. Hence, if a zero-level signal is to be maintained at the input of attenuator P1, the output of the oscillator must be 6 dB higher to overcome the loss across R3. Thus, if the VU meter were to be connected across R1 and R2, it would result in a reading 6 dB too high, when the level at the input of P1 was actually zero level. To overcome this condition, the VU meter is connected across the resistor R2. Doing so drops the signal level to the VU meter 6 dB, or 50 percent. Now, if the level across the first attenuator is zero, the meter will also read zero, although the actual signal level from the oscillator is 6 dB higher. Attenuators P1 and P2 have 600 ohms impedance and are of the bridged-T type. The first attenuator has a total loss of 100 dB, variable in steps of 10 dB. The second attenuator has a total loss of 10 dB in steps of 1 dB, while P3 has a total loss of 1.0 dB in steps of 0.1 dB.

The normal send impedance of this particular gain set is 600 ohms, with three fixed pads for matching input impedances of 250, 150, and 50 ohms. The fixed loss of these pads is 10, 15, and 20 dB respectively, and it must be added to the indicated loss of the variable attenuators when they are employed. The three variable-loss attenuators present a maximum loss of 111.10 dB.

This total loss is variable in steps of 1 dB. The preceding losses will cover most applications, and if one is encountered where more loss is required, external pads may be used and their loss added to the loss of the attenuators in the gain set. Resistor R4 is used to terminate the send position of the gain set when devices having a bridging input impedance are measured.

The VU meter with its attenuator is shown in Section D of the diagram. It is designed to cover a range of plus 4 dBm to plus 46 dBm in steps of 2 dB each. The design and operation of VU meters is discussed in Section 10. Series resistor R5 in the meter circuit is used for adjusting the meter calibration against a standard.

The VU meter key switch is shown in Section C. Resistors R6 and R7 serve only to replace the load of the meter as it is switched from the send to receive positions or vice versa. Their value in ohms is equal to the input impedance of the VU meter. The use of these dummy load resistors prevents level changes when the meter load is removed.

The receive section (Section B) of the gain set consists of the several terminating resistors (R8 to R18) and autotransformer T1, with a provision for connecting the VU meter across its 600-ohm terminals. The terminating resistors are 1 percent, wirewound, vitreous type, capable of continuously dissipating 20 watts of power because they carry the greater portion of the output power of the device being tested by the gain set.

Eleven standard output impedances are provided which are selected by means of a switch serving a dual purpose: that of selecting the proper terminating resistor and the proper impedance tap on the autotransformer. The principal purpose of the autotransformer is to provide a 600-ohm terminating impedance for the VU meter and at the same time match the output impedance of the device nnder test. Because of this latter feature, no correction is required for the VU meter when reading the output level, regardless of the terminating impedance. Although the diagram shown calls for only one VU meter, two meters may be used, one across the send section and the second across the receive section, thus eliminating the need for the transfer switch.

Because of the great difference in the power levels between the send and receive sections, the two sections must be carefully isolated and shielded to prevent feedback between the input and output circuits. Therefore, the output level should not exceed 20 watts if the receive section is included in the gain set proper. If the receiving section is a separate unit external to the attenuator section, it may be designed to handle considerably more power. Generally, if the power output level is more than 25 watts, an external noninductive load resistance is used and the output power is measured, using a vacuum-tube voltmeter and the wattage computed.

It will be noted the return connection of the attenuators is brought to a common ground connection, using separate ground wires. Each ground wire is individually insulated and shielded to prevent leakage at the high frequencies. Equipment being tested should be grounded to common ground point where it connects the actual ground.

If measurements are consistently being made where the frequency variation must be read to within 0.10 dB, a variable attenuator with a total loss of 1 dB, variable in steps of 0.10 dB, may be included in the attenuator group.

An unbalanced gain set may be used for measuring balanced circuits by the insertion of a repeat coil between the send terminals and the input of the device being tested. The Daven gain set shown in Fig. 22-122A has a repeat coil permanently built into the send circuit to facilitate its use with balanced circuits. It also has the advantage that the ground return of the attenuator section is isolated from grounds in the equipment under test.

At times it may be desirable to ground the lower side of the receive section; if so, it is grounded at the common ground point of the system. The techniques of using a gain set are discussed in Section 23.

22.123 Describe the basic circuitry for a balanced gain set.—Balanced gain sets as a rule employ balanced bridged-T variable and fixed attenuator networks. In the early design of gain sets, variable H-type attenuator configurations were used; however, the use of such networks increased the contacts per attenuator (consequently noise) and increased cost of manufacture. Therefore, an improved type of bridged-T balanced variable attenuator is used in modern designs. A balanced gain set developed by A. C. Davis and manufactured by Altec Lansing is pictured in Fig. 22-123A. It is of the two-meter type, permitting the simultaneous observation of both the send and receive levels.

In the send section of this instrument four variable attenuators are employed which may be switched from an unbalanced configuration. The first decade attenuator covers a range of 100 dB in steps of 10 dB; the second covers 10 dB in steps of 1 dB, and the third covers 1 dB in steps of 0.10 dB. The remaining attenuator (at the extreme left) is a matching network in the send section for supplying impedances of 600, 250, 150, and 50 ohms, either terminated or unterminated, to the device under test. The terminated outputs are for use with devices employing bridging inputs. The send-section attennator networks are isolated from the oscillator by a 1:1 repeat coil. The attenuator for the send meter (left) is calibrated plus 4 to 24 dBm. Levels of plus 18 dBm down to minus 120 dBm are possible in steps of 0.10 dB.

The receive section is well shielded from the send section and consists of a load network capable of dissipating 30 watts. This network is composed of an autotransformer, a group of terminating resistors as described in Questions 22.122 and 22.124, and a VU meter calibrated plus 4 to 44 dBm. Terminating impedances of 600, 150, 16, 8, and 4 ohms either terminated or unterminated are available. The VU meter has jacks in the circuitry for use as an external meter.

The accuracy of the send attenuator network is plus or minus 0.10 dB for any setting over a range of 10 to 50,000 Hz. All circuitry is passive (no amplifiers), with an induced distortion of less than 0.20 dB. The frequency response of the two VU meters is matched and meets the standards for such meters. The attenuator section may be used separately if desired by means of jacks on the front panel, with a link for ground separation if required. Connections for the oscillator send and receive and ground are provided at the rear for rack mounting.

Since no coil is used in the send section, the output is purely resistive, and if used in the balanced position the circuit is balanced to ground since the centers of the networks are grounded. Therefore, if a repeat coil is necessary in the send section, it must have a broad, uniform frequency characteristic. The use of repeat coils with gain sets is discussed in Questions 23.14 and 23.15.

A basic configuration for a typical balanced gain set is given in Fig. 22-123B. (See Question 23.24.)

22.124 Describe the circuitry for a gain-set receive section employing an autotransformer.—Fig. 22-124 shows the circuit connections and component values for an autotransformer-type gain-set receive section, similar to that described in Question 22.122.

Fig. 22-123A. Balanced two-meter gain set Model 9704 manufactured by Altec Lansing Corp.

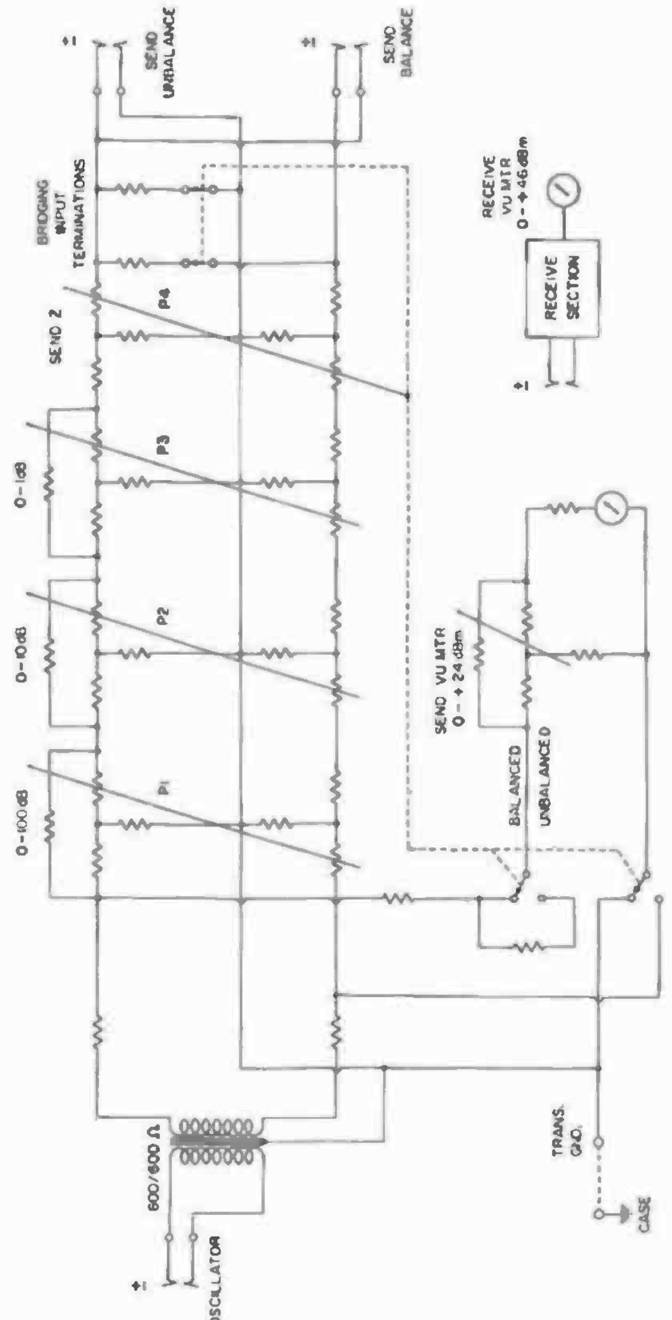


Fig. 22-123B. Basic configuration for a balanced gain set.

The autotransformer is designed to provide an impedance match between a 7500-ohm VU meter and 4, 8, 16, 30, 150, 200, 250, 500, and 600 ohms. Resistors R1 to R9 are 25-watt, wirewound, vitreous-type resistors and carry the full output power of the device being tested. Note that the value of the terminating resistor is greater than the impedance for a given tap on the autotransformer, because the autotransformer has both ac impedance and dc resistance. Being in parallel with the impedance of the autotransformer winding, the terminating resistance must be slightly greater in value to obtain the exact impedance. The primary purpose of the autotransformer is to provide an

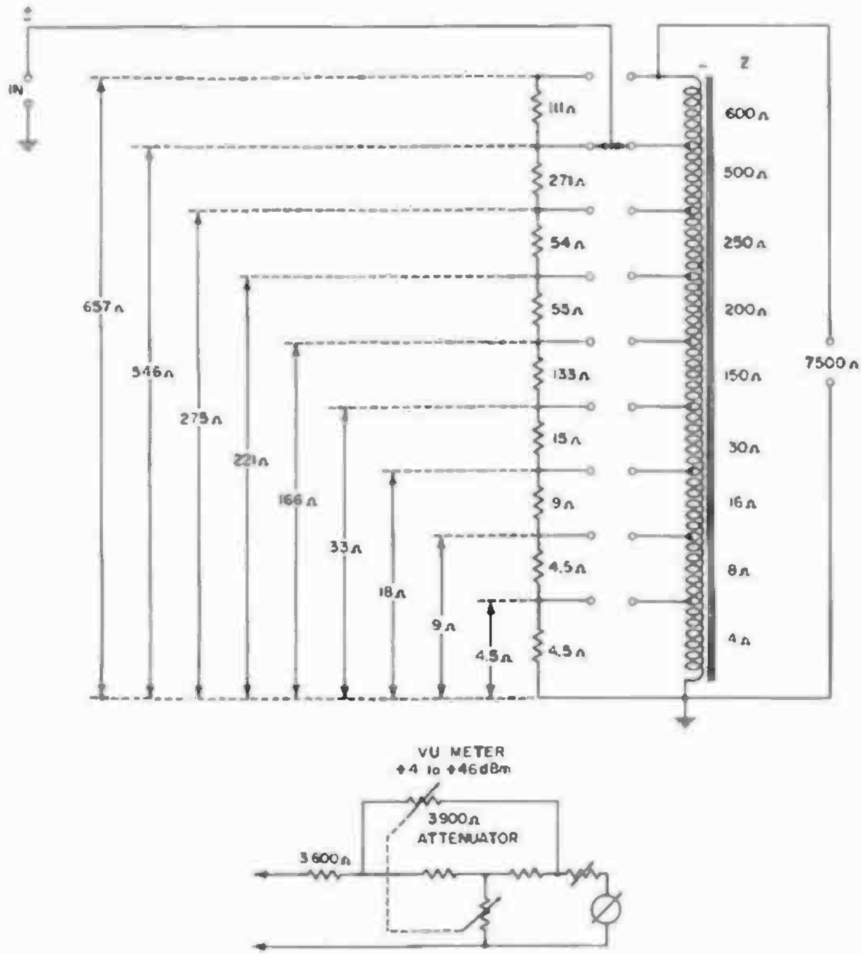


Fig. 22-124A. Variable termination panel using an autotransformer and a group of terminating resistors.

impedance match between a given output impedance and the 7500-ohm impedance of the VU meter.

The proper terminating resistance for any tap may be determined by terminating the 600-ohm tap with a 7500-ohm resistor, then connecting a resistor across a given impedance tap and varying the value of the resistance until the correct impedance is obtained when the two are in parallel. This measurement is best made with an impedance bridge.

The autotransformer should be designed to carry about 25 percent of the maximum power to be dissipated in the terminating resistors. As a rule, autotransformer receive sections have good frequency response because of their design. The frequency response should be at least 20 to 20,000 Hz.

If the receive section described in the foregoing is to be used with a single meter-type gain set where the meter is switched from input to output and vice versa, provision must be made to terminate the VU meter connections in 7500 ohms when the meter is removed. If this is not done, the impedance seen by the output of the device under test will not be correct.

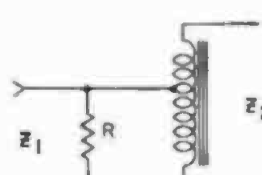


Fig. 22-124B. Autotransformer with terminating resistor.

The terminating resistor for an autotransformer panel may be calculated:

$$R = \frac{ZX_L}{\sqrt{X_L^2 - Z^2}}$$

where,

R is the terminating resistor in ohms,
Z is the line impedance,
X_L is the inductive reactance of the coil at the lowest frequency of interest.

The basic circuit for these calculations is given in Fig. 22-124B.

22.125 Can an autotransformer-type termination panel be used when making distortion measurements?—Yes, provided the autotransformer is not overloaded. If it is, harmonics will be generated due to the saturation of the core material.

When making intermodulation measurements involving frequencies below 100 Hz, it is best to use a noninductive resistor as a termination, rather than the autotransformer termination. If the autotransformer is to be used, measurements should be made at various power ratings and then compared to similar measurements made using a resistive termination, to ascertain whether intermodulation products are being generated by the autotransformer.

22.126 What is a transmission set?—In the past the term "transmission set" was taken to mean a gain set. However, this term is now applied to measuring equipment designed for use with telephone and carrier system measurements. Fig. 22-126 shows such an instrument, manufactured by Waveforms, Inc., consisting of a signal generator, amplified VU meter, attenuator system, and power supplies. The instrument may be used for a variety of measurements, such as gain and loss, frequency response, distortion, impedance, noise, and signal-to-noise ratio in equipment and systems. The design is such that it will accept or transmit at a number of impedances, both balanced and unbalanced. The oscillator is of the RC bridged-T type, covering a frequency range of from 9 to 120,000 Hz, explained in Questions 22.49 and 22.50.

22.127 What is a variable termination panel?—A resistive or autotransformer network designed for terminating audio-frequency devices when frequency and power characteristics are measured. A typical variable terminating network is the receiving section of the gain set described in Question 22.124.

22.128 What is an attenuator panel?—An instrument containing a number of attenuators or loss pots, calibrated in decibels, voltage, or resistance. The purpose of an attenuator is to provide known values of loss for the accurate measurement of small voltages. A known voltage is applied to the input of the attenuator group and is reduced in known quantities by inserting known values of loss. Attenuator panels as a rule contain a meter, whereby the input voltage may be set to a given value before attenuating the signal. In this manner, the ratio of attenuation to the input voltage is established. (See Question 22-43.)

22.129 What is an intermodulation analyzer?—An instrument used for the measurement of intermodulation distortion generated within audio-frequency devices. An elementary block diagram of a typical intermodulation analyzer connected for the measurement of intermodulation products generated by an audio amplifier is shown in Fig. 22-129A. The analyzer section consists of two sections, a signal generator and an ana-

Fig. 22-126. Waveforms Inc Model 452A Transmission set designed for audio and telephone carrier use.

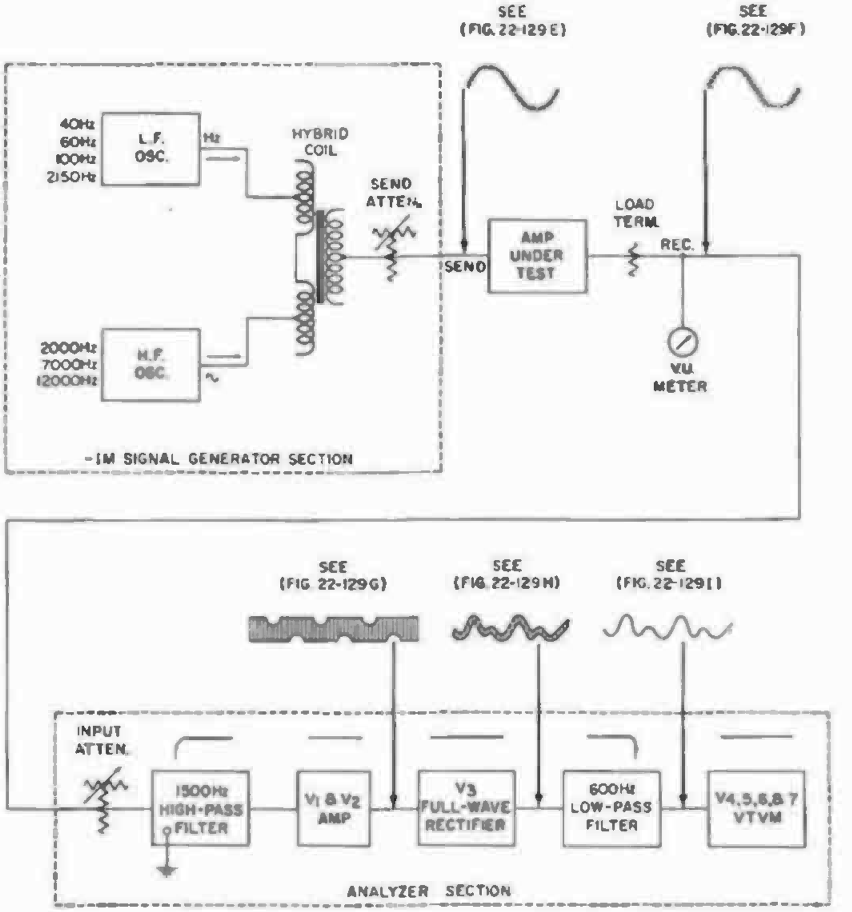


Fig. 22-129A. Elementary block diagram of an intermodulation analyzer and signal generator, showing how the two sections of the analyzer are connected to an amplifier, using the SMPTE method of measuring.

lyzer section. The signal generator section generates two signal frequencies, one high and one low, which are mixed in a predetermined ratio and then applied to the input of the device or circuit to be tested. The analyzer section is connected across the output termination of the amplifier and measures the intermodulation distortion generated by the amplifier in percent of intermodulation.

The signal frequencies supplied by the generator section are usually set for a ratio of 4:1, the lower frequency being 12 dB higher in amplitude than the higher frequency. For certain types of tests, it may be desirable to use a ratio other than 4:1. The low frequency may be any frequency between 40 and 100 hertz, while the higher frequency may lie anywhere between 1000 and 12,000 Hz. The choice of frequencies will depend on the type of equipment being tested, its frequency range, and its characteristics.

To better understand the operation of an intermodulation analyzer, a brief review of intermodulation theory is in order. The most commonly used method of measuring harmonic distortion is shown in Fig. 22-129B. A single sine-wave frequency is applied to the input of the device to be tested and then the harmonics at the output are measured in percent of the fundamental frequency, using a distortion-factor meter

Fig. 22-129B. Block diagram of the method of measuring harmonic distortion of an amplifier.

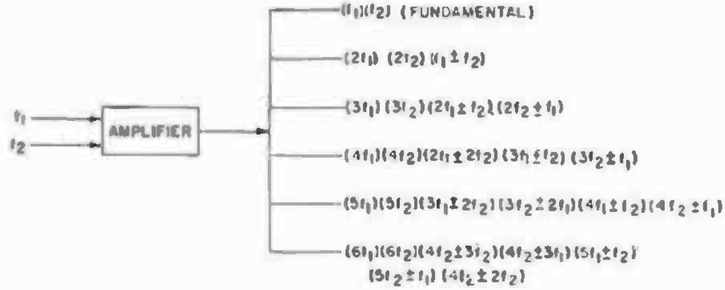


Fig. 22-129C. Intermodulation products generated within a nonlinear device when two frequencies f_1 and f_2 are applied to the input.

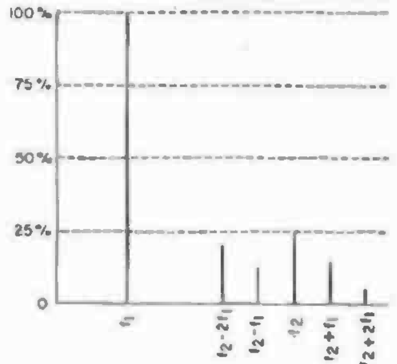


Fig. 22-129D. Intermodulation spectrum as the analyzer section sees it using the SMPTE method. Sum and difference frequencies for f_1 and f_2 have been omitted for clarity.

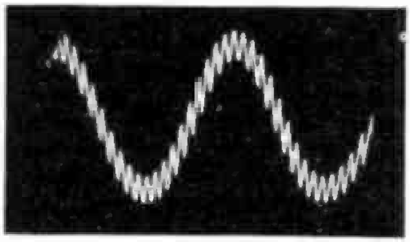


Fig. 22-129E. Undistorted intermodulation test signal at output of the signal generator "send" terminals. The waveform has been expanded to show the high-frequency component (f_1 equals 40 Hz, f_2 equals 2000 Hz).

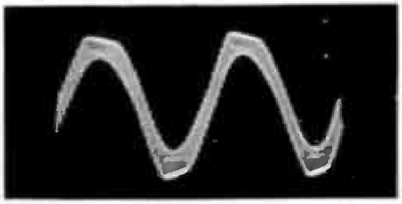


Fig. 22-129F. Distorted intermodulation signal at the output of the amplifier under test (f_1 equals 40 Hz, f_2 equals 2000 Hz, the frequency ratio is 4:1).

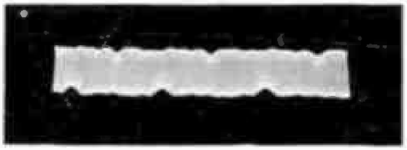


Fig. 22-129G. Appearance of the distorted signal at the plate of the second amplifier in the analyzer section.

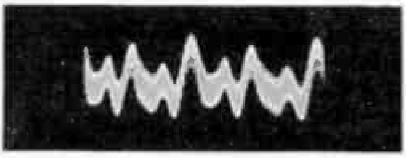


Fig. 22-129H. Intermodulation products at the output of the full-wave rectifier.

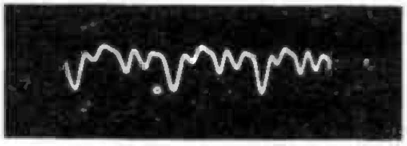


Fig. 22-129I. Distorted signal after passing through 500-Hz, low-pass filter.

or harmonic wave analyzer. (See Questions 22.60 and 22.65.)

Nonlinear distortion is sound recording and reproducing equipment affects the reproduction by introducing frequency components not present in the original program material. The effect of these added components is to annoy the listener because of masking effects and the interference created with the original frequencies of the signal. This results in unpleasant reproduction and often induces listener fatigue.

Although the single-frequency method of measuring distortion has been in use for many years, it is an acknowledged fact that such tests are not completely satisfactory and do not present a complete picture of the distortion characteristics of a given device. Even systems which have low-harmonic distortion will not sound just right to the listener. It is not uncommon to take two amplifiers of the same manufacture, design, and harmonic distortion characteristics, listen to their reproduction and find they do not have the same listening quality. What is the difference? The difference is generally in the percentage of intermodulation distortion generated within the amplifiers.

The intermodulation method of measuring distortion in audio-frequency devices more nearly approaches the manner in which the human ear hears than any other system of measurement. Also, the intermodulation system is several times more sensitive than the conventional method of measuring harmonic distortion.

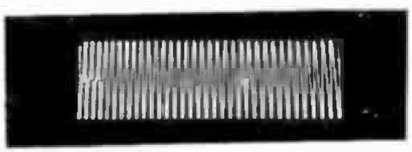


Fig. 22-129J. Appearance of signal at the output of the full-wave rectifier when an undistorted signal is applied to the analyzer input.

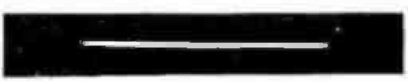


Fig. 22-129K. Appearance of an undistorted signal after passing through the 500-Hz, low-pass filter.

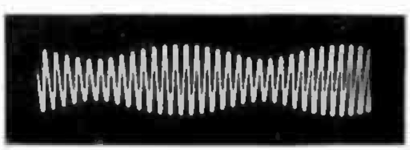


Fig. 22-129L. Appearance of the 10% intermodulation calibration signal with oscilloscope sweep frequency set to show the modulation pattern.

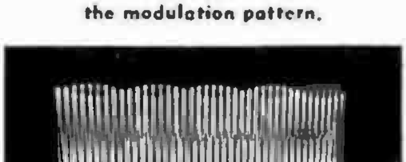


Fig. 22-129M. Appearance of the 10% intermodulation signal, consisting of a 2000-Hz and 2150-Hz combination to produce a 10% calibrating signal.

Intermodulation measurements have been used in the motion picture industry for several years and have been adopted by the Society of Motion Picture and Television Engineers (SMPTE) and standardized by the USASI, formerly known as the American Standards Association (ASA). Intermodulation measurements relative to motion picture recording and reproduction are discussed in Questions 18.274 to 18.281, and general measurements are covered in Section 23. A second system known as the CCIF, recommended by the International Telephonic Consultative Committee, also employs two frequencies but in a different manner than that used by the SMPTE method.

Intermodulation distortion may be defined as: "The production in a nonlinear circuit element of frequencies corresponding to the sum and differences of the fundamentals and harmonics of two or more frequencies which are transmitted through that element." Thus, when two frequencies are applied to the input of an amplifier, not only do the fundamental frequencies appear in the output but also sum and difference frequencies which are, as a rule, not harmonically related to the fundamental frequencies. (See Fig. 22-129C.)

For two fundamental frequencies f_1 and f_2 the intermodulation products will consist of second-order terms, $(f_1 + f_2)$ and $(f_1 - f_2)$; third order terms $(2f_1 + f_2)$, $(2f_1 - f_2)$, $(f_1 + 2f_2)$, and $(f_1 - 2f_2)$, and so on for the higher sum and difference frequencies. Frequencies higher than fifth-order terms need not be measured. However, it is important that the fifth-order harmonics be measured to obtain complete distortion information.

Fig. 22-129D shows a portion of the frequency spectrum as it appears to the input of the analyzer section. Frequencies f_1 and f_2 are the fundamental frequencies. The others are the sum and difference frequencies generated within the amplifier. It will be noted the lower fundamental frequency f_1 is taken as 100-percent amplitude and all other frequencies are plotted relative to this frequency. The amplitude of the fundamental frequency f_2, being 12 dB lower in amplitude, corresponds to 25-percent amplitude. This is a ratio of 4:1.

The amplitudes of the sum and difference frequencies will be dependent on the amount of intermodulation distortion present in the amplifier. If a frequency of 100 Hz is selected for f_1, and 5000 Hz for f_2, sum and difference frequencies will appear at intervals of 100 Hz. Other frequencies above and below f_2 will also appear but these have been omitted for clarity. Under certain conditions the sum and difference frequencies may assume an amplitude equal in value to the fundamental frequencies, f_1 and f_2.

Referring again to Fig. 22-129A, assume the amplifier is to be measured using 40 Hz for the low frequency and 2000 Hz for the high frequency, mixed in a ratio of 4:1; that is, the high frequency will be set 12 dB lower in amplitude than the 40-Hz signal and applied to the input of the amplifier in this manner. The appearance of this signal at the output of the signal generator section will be as shown in Fig. 22-129E. It will be noted the peaks of the 2000-Hz signal superimposed on the 40-Hz waveform are uniform in amplitude and undistorted and that the combined waveform is symmetrical.

Assume the amplifier has 10-percent intermodulation distortion. The waveform at the output terminals of the amplifier will appear as shown in Fig. 22-129F. Here it will be seen that both the positive and negative peaks of the 2000-Hz component have been flattened after passing through the amplifier. To understand the functioning of the analyzer components, this waveform will be followed through the various sections of the analyzer circuit.

The distorted waveform of the amplifier is applied to the input of the analyzer, which consists of a 1500-Hz high-pass filter which removes the 40-Hz component and its harmonics, leaving only the distorted 2000-Hz component.

The signal is then amplified in the two-stage amplifier V1 and V2 (Fig. 22-129A). Fig. 22-129G shows the waveform at the output of the amplifier section. The notches in the positive and negative halves of the waveform are caused by the flattening of the peaks as shown in Fig. 22-129F.

The high-frequency component is then applied to a full-wave rectifier circuit. Fig. 22-129H illustrates the appearance of the high-frequency component at the output of the full-wave rectifier. Here it may be seen how the 2000-Hz component was distorted in the amplifier due to modulation of the 40-Hz component. Both second- and third-order harmonics are present. Because of full-wave rectification, the 2000-Hz signal frequency is now 4000 Hz.

The 4000-Hz component is now passed through a 600-Hz, low-pass filter which removes the 4000-Hz component, leaving only the modulation envelope caused by the 40-Hz component. Its appearance is shown in Fig. 22-129I as a ripple voltage or pulsating dc voltage, and its waveform is similar to the waveform shown in Fig. 22-129H. The only difference is that the high-frequency component has been removed. The amplitude of this ripple voltage is measured by a vacuum-tube voltmeter which is calibrated to read percent intermodulation distortion.

To further illustrate the functioning of the intermodulation analyzer section, suppose that a distortion amplifier is measured. Applying the same test frequencies and ratio, instead of the distorted waveform of Fig. 22-129F appearing at the amplifier output, the waveform will appear the same as that applied to the amplifier input. (Fig. 22-129E.)

This undistorted signal is passed through the 1500-Hz, high-pass filter removing the low-frequency component and then to the full-wave rectifier. Because the amplifier is distortionless, there will be no modulation of the 2000-Hz component by the 40-Hz signal; therefore the peaks of the 4000-Hz signal at the output of the full-wave rectifier all have the same amplitude and appear as shown in Fig. 22-129J. The 600-Hz, low-pass filter removes the high-frequency component, leaving only a pure dc voltage as shown in Fig. 22-129K.

As the vacuum-tube voltmeter reads only alternating current or pulsating voltages, no indication appears on the meter; thus, no distortion is indicated.

The waveforms of Figs. 22-129F to I are for a small push-pull amplifier with approximately 10-percent intermodulation distortion. For single-ended amplifiers and those of other classifications, the waveform will take on a variety of appearances.

The appearance of the 10-percent calibration signal used for the initial calibration of the analyzer before the 4:1 ratio is set is shown in Fig. 22-129L. When the picture shown in Fig. 22-129L was made, the oscilloscope sweep was set to show the modulation of the higher frequency by the lower frequency. The frequencies shown are 2000 and 2150 Hz.

Fig. 22-129M shows the same two frequencies, except that the sweep of the oscilloscope was set to show the beating between the two signals of 2000 and 2150 Hz.

Intermodulation measurements are discussed in Questions 23.117 to 23.125.

22.130 Describe the signal generator section for an intermodulation analyzer.—The schematic diagram of the signal generator section of a typical intermodulation analyzer is shown in Fig. 22-130A. This section consists of two Wien-bridge or resistance-tuned oscillators. (See Question 22.50.) Tubes V1, V2, and V3 constitute the low-frequency oscillator and supply frequencies of 40, 60, and 100 Hz. Tubes V4, V5, and V6 constitute the high-frequency oscillator and supply frequencies of 2000, 7000, and 12,000 Hz.

Included in this section is a conventional power supply using diode rectifiers, two LC filters (sections L1 and L2), and two gaseous voltage regulator tubes, V7 and V8, which supply a source of regulated 300 volts to the plates of the oscillators and the amplifiers of the analyzer section. The plate voltage for oscillator output amplifier tubes V3 and V6 is supplied from the junction point of the two filter chokes. The ripple voltage at the junction of the filter sections must not exceed 15 millivolts, and 5 millivolts at the regulator tubes to prevent hum modulation of the signal voltages. The oscillators are composed of resistors R1 to R9 and capacitors C1 to C8, for the low frequencies, and R26 to R32 and C18 to C23 for the high frequencies, connected in the control grids of tubes V1 and V4. The various combinations of test frequencies are selected by switches S1 and S2.

In addition to the foregoing test frequencies, a frequency of 2150 Hz is provided in the low-frequency oscillator which is used in combination with 2000 Hz in the high-frequency oscillator for setting the intial calibration of the analyzer section. When these two frequencies are combined in a ratio of 10:1 (20 dB), they produce a 10-percent intermodulated signal (Fig. 22-129L). The output tubes, V3 and V6, are connected for cathode-follower operation. The signal is takeu from the cathodes, carried to a plug, and then to the analyzer section over a flexible cable. The oscillator signals are mixed in the analyzer section by means of a hybrid coil. A fraction of one cycle of a test signal consisting of 40 and 2000 Hz is shown in Fig. 22-130B.

Decoupling circuits composed of capacitors C17 and C32 and resistors R24 and R46 preveut the oscillator signals from entering the power supply, thus preventing the mixing of the signals before they enter the mixing circuits in the analyzer. The harmonic distortion for any one frequency of the oscillators must not exceed 0.25 percent.

Jacks for the external connection of variable oscillators are also provided. However, the fixed frequencies iu general usage throughout the electronic industry are those specified by the SMPTE, and are adequate for most measurements.

It is important that the frequencies of 40, 60, and 100 Hz be closely adjusted to their values. Any variation in the 40-Hz frequency should be to the plus side rather than the minus side. The calibration should be held to 40 Hz, plus 2 Hz, minus zero. The 10-percent calibration frequencies, 2000 and 2150 Hz, should be held to plus 1 percent minus zero. The exterior and interior views of the signal generator section of an intermodulation analyzer manufactured by the Altec Lansing Corp. are shown in Figs. 22-130C and 22-130D.

22.131 Give a schematic diagram for an intermodulation analyzer section for use with the signal generator, described in Question 22.130.—The schematic diagram for such an analyzer appears in Fig. 22-131A. Starting at the input terminal upper left, these terminals present an unbalanced 7500-ohm noninductive bridging input to the device under test and receive the output signal. It is imperative that the output of the device being tested is terminated in its normal load impedance with a noninductive resistor capable of carrying the full output power of the device under test. The input of the analyzer being a bridging impedance is in reality

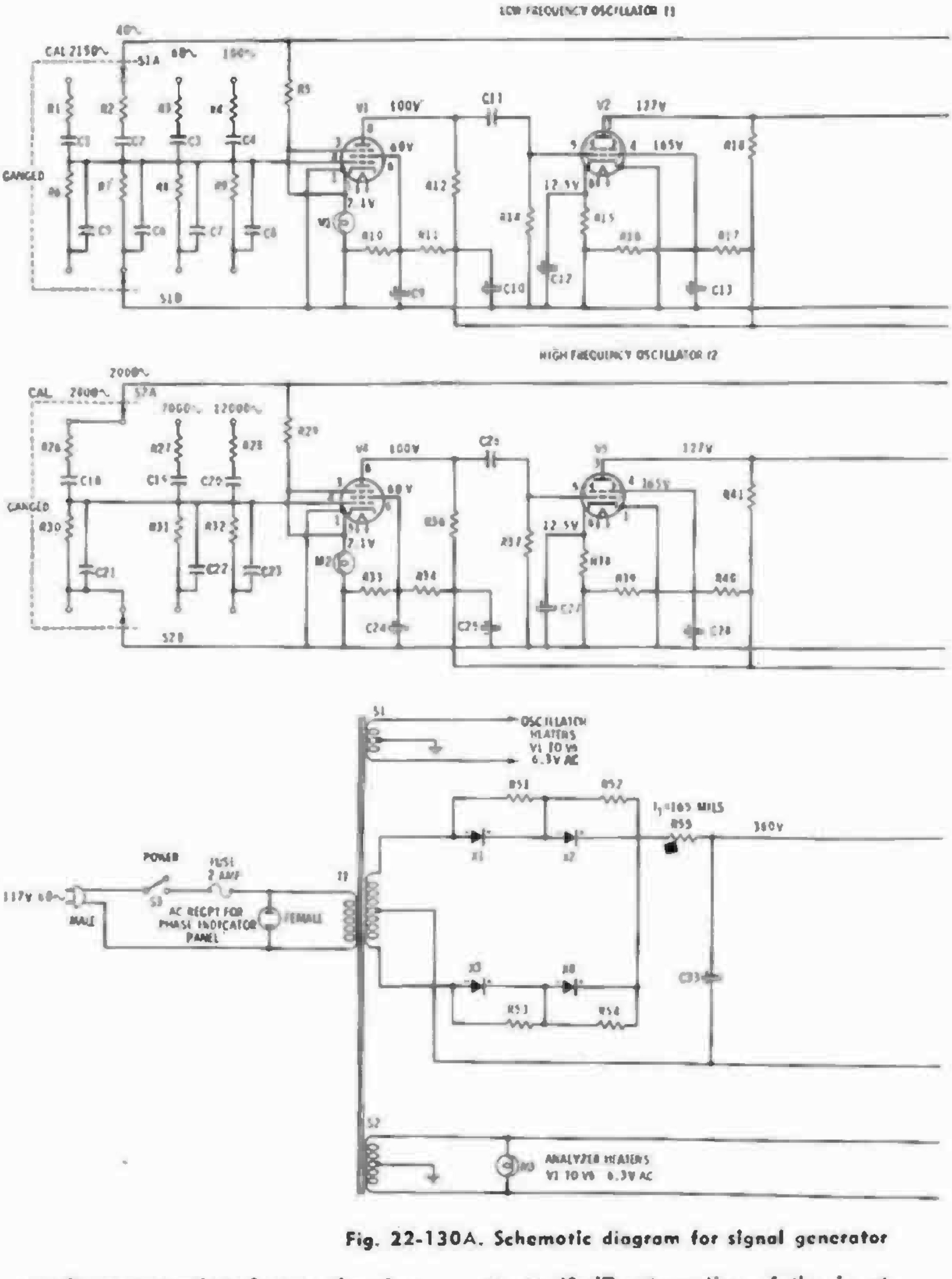
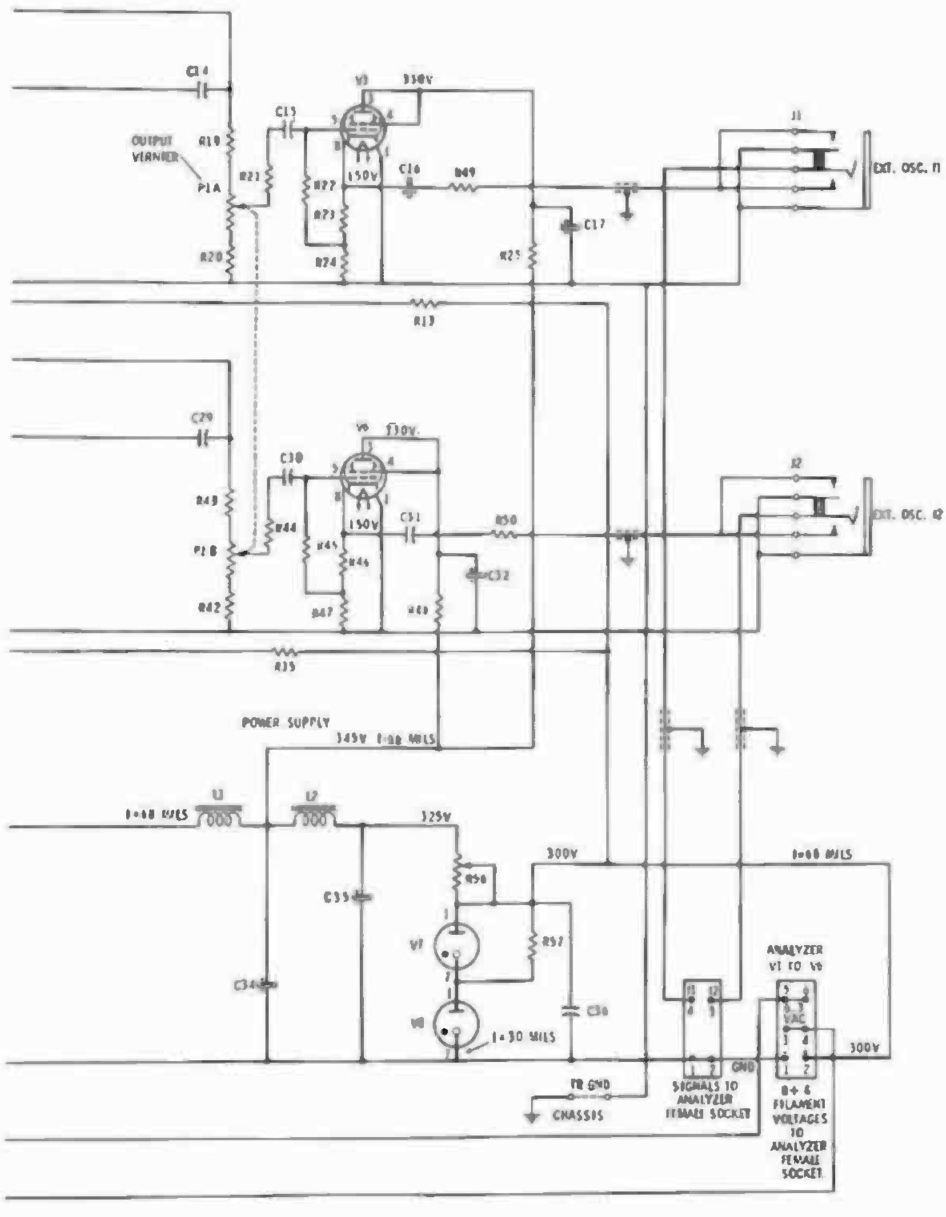


Fig. 22-130A. Schematic diagram for signal generator

a voltage-measuring device; therefore the output power of the device under test must be dissipated in the load resistance to prevent overloading the input of the analyzer.

The input signal is taken from the input plus/minus terminal through 7500-ohm resistor R1 to section A of a six-deck, eleven-circuit function switch, S1. This switch selects the proper circuitry for calibration and analyzing the input signal. This switch also provides up to 40-dB attenuation of the input signal in steps of 10 dB each, for testing devices with high output level. The final position connects directly to the vacuum-tube voltmeter for measuring the internal noise of the device being tested.

The first four positions of function switch S1 are for calibrating the signal generator section and the analyzer section and are indicated: f1 10% Cal, f2 10% Cal, f1 4:1 Cal, and f2 4:1 Cal. The first two positions adjust the level of the



and power-supply section of intermodulation analyzer.

10-percent intermodulation signals (Fig. 22-129L) during the initial calibration. The 10-percent signal is set to a ratio of 10:1 by a 20-dB attenuator. A 12-dB attenuator is inserted for the 4:1 ratio (both these attenuators are in the signal analyzer section). With the function switch set to zero attenuation (position 6), the received signal is connected to the upper terminals of a 1500-Hz high-pass filter. This filter is of composite design, having extremely sharp cutoff characteristics (Fig. 22.131B) and consists of a group of toroidal coils. Resistors R2 through R6 compose an input attenuator network. For intermodulation testing the function switch is set to one of the five loss positions to reduce the input signal and to prevent overloading of the high-pass filter and to present a solid resistive input for the device under test.

The purpose of the 1500-Hz high-pass filter is to remove the low-fre-

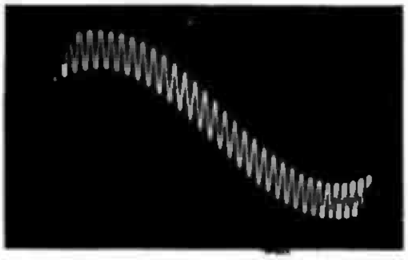


Fig. 22-130B. Fraction of a cycle of the intermodulation signal at "send" terminals of the send section. The waveform has been expanded to show the high-frequency signal on the low-frequency carrier (f_1 equals 40 Hz, and f_2 equals 2000 Hz).

quency component of the intermodulation signal being analyzed. Thus, only the high-frequency component is permitted to pass to the analyzer circuits. The high-frequency signal is fed to transformer T1, which has an impedance ratio of 600/50,000 ohms and is terminated on the secondary side by a 50,000-ohm, Use Calibrate control. This potentiometer adjusts the signal level applied to two-stage resistance-coupled amplifier V1 and V2, tubes which are heavily bypassed at the cathodes and screen grids. The output of V2 is parallel plate-coupled to interstage transformer T2, having an impedance ratio of 15,000/50,000 ohms.

Fed by the secondary of T2 is a full-wave bridge rectifier circuit, consisting of two 6AL5/5726 twin diodes V3 and V4. It might be well to mention at this point that semiconductor diodes cannot be used in this position due to their higher distortion characteristics. Also, a full-wave bridge circuit is necessary because the distortion occurs on both half-cycles of the waveform being analyzed. If a half-wave rectifier circuit is employed, the rectifier ignores the distortion on the unrectified half of the cycle.

The 600-Hz low-pass filter following the bridge circuit removes the high-frequency component of the signal being analyzed, leaving only a pulsating dc or ripple frequency (Fig. 22-129I). This pulsating dc signal results from modulation of the high-frequency component by the lower frequency of the test signal. The coils for this filter are also of toroidal design. All coils must be encased in nested shields of at least 90-dB attenuation.

A resistive network composed of resistors R19 to R21 and potentiometer P4 is connected at the output of the filter to provide an adjustment for setting the carrier-current meter, M1, to a predetermined calibration point on its scale.

The metering section, a vacuum-tube voltmeter, consists of two tubes, V5 and V6. The ripple voltage at the output of

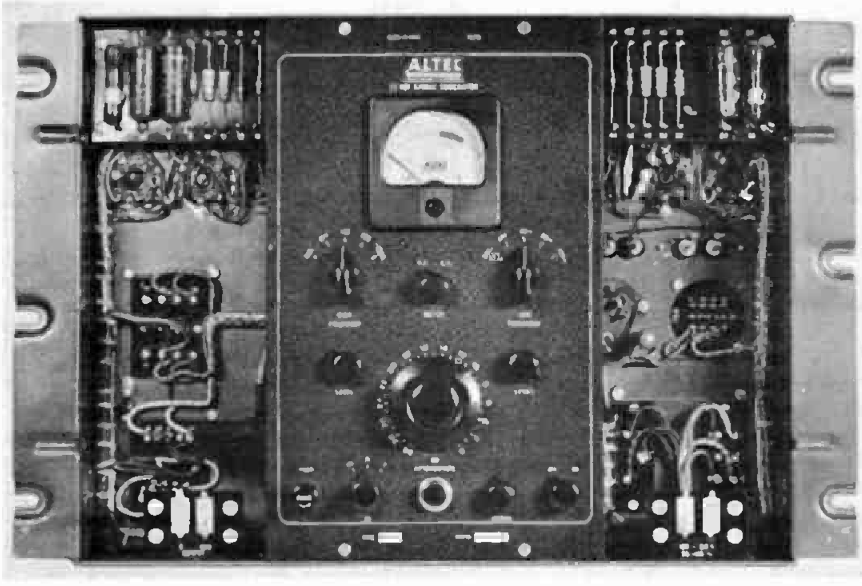


Fig. 22-130C. Exterior view of signal generator section of the Altec Lansing Corp. intermodulation analyzer.

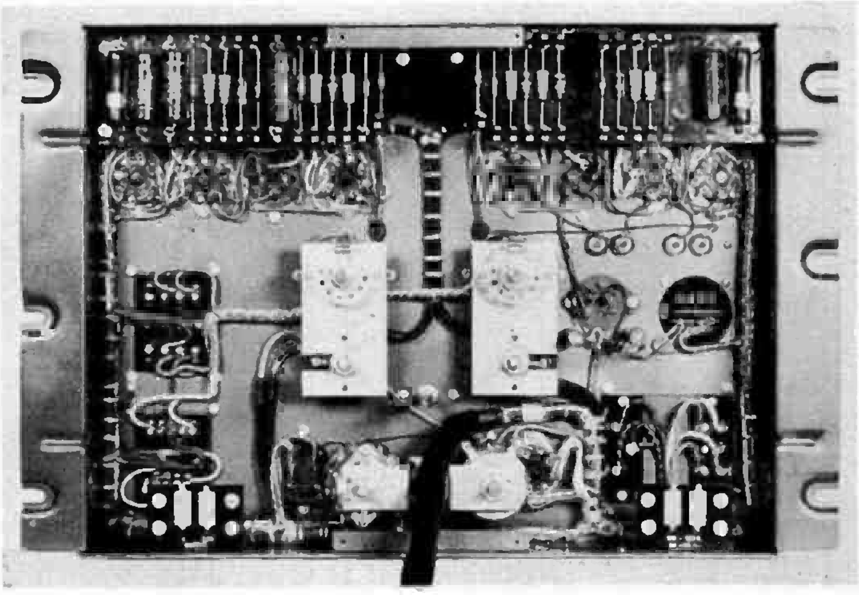


Fig. 22-130D. Interior view of the signal generator section of the Altec Lansing Corp. intermodulation analyzer.

the low-pass filter is read directly in percent intermodulation distortion on meter M2. The input to the metering circuit is automatically switched to the output of the low-pass filter when function switch S1 is in any of the input attenuation positions. The range of the metering circuit is controlled by an input attenuator network, resistors R22 to R27. After amplification the ripple voltage is rectified by four diodes D1 to D4, which are connected in a full-wave bridge circuit. An 80-μF capacitor C26 is connected across the meter movement to remove low-frequency pulsations. The lower end of the bridge circuit is returned to ground through potentiometer P6 in the cathode circuit of V5, a negative-feedback loop which serves to adjust the gain of the metering circuit and flatten the frequency characteristics.

The attenuator network in the metering circuit employs 1-percent precision resistors and provides six full-scale deflections of 0.3, 1.0, 3.0, 10.0, 30, and 100 percent. These same steps provide a full-scale noise-measuring range from minus 60 dBm, to plus 10 dBm. The meter scale is calibrated to read in both percent modulation and decibels at the same time.

Three 100,000-ohm potentiometers P1, P3, and P7 are included in the circuitry for adjusting the amplitude of the signal fed to the oscilloscope switch, S2. With S2 in the off position, the oscilloscope is grounded; in the in position the oscilloscope is connected to the output of a hybrid coil in the signal section and displays the waveform of the signal being sent into the device under test (see Fig. 22-129E). With switch S2 set to the in position, the output of the waveform of the device being tested is displayed. The modulation position displays the modulated waveform at the output of the high-pass filter, and in the IM position, the generated intermodulation products as seen by the metering circuits.

At the right of sections D, E and F of S1 is a hybrid-coil signal-mixing circuit. Here the test frequencies from the signal generator are applied to a resistive network for controlling their amplitudes and setting the calibration and testing ratios. Hybrid coil T3 has an impedance ratio of 600-600/600 ohms. When the coil is balanced by potentiometer P10, the hybrid coil provides a minimum balance of 45 dB at all frequencies between 40 and 12,000 Hz. Test frequencies arriving from the signal generator are fed to test signal switch S3, which permits the selection of the test frequency either singly or in combinations. Potentiometers P8 and P9 adjust the level of the input signals before they are passed to the ratio atten-

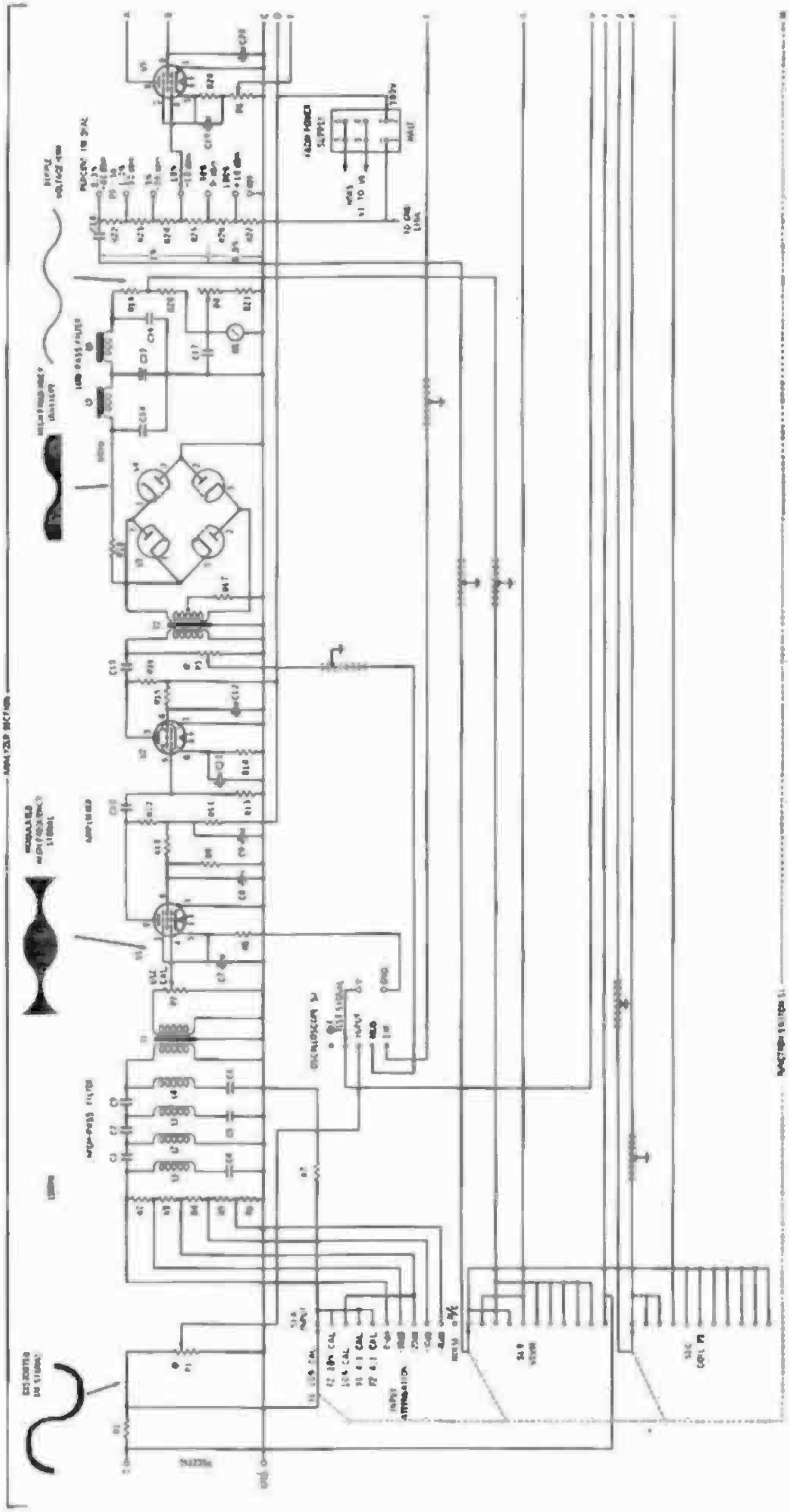
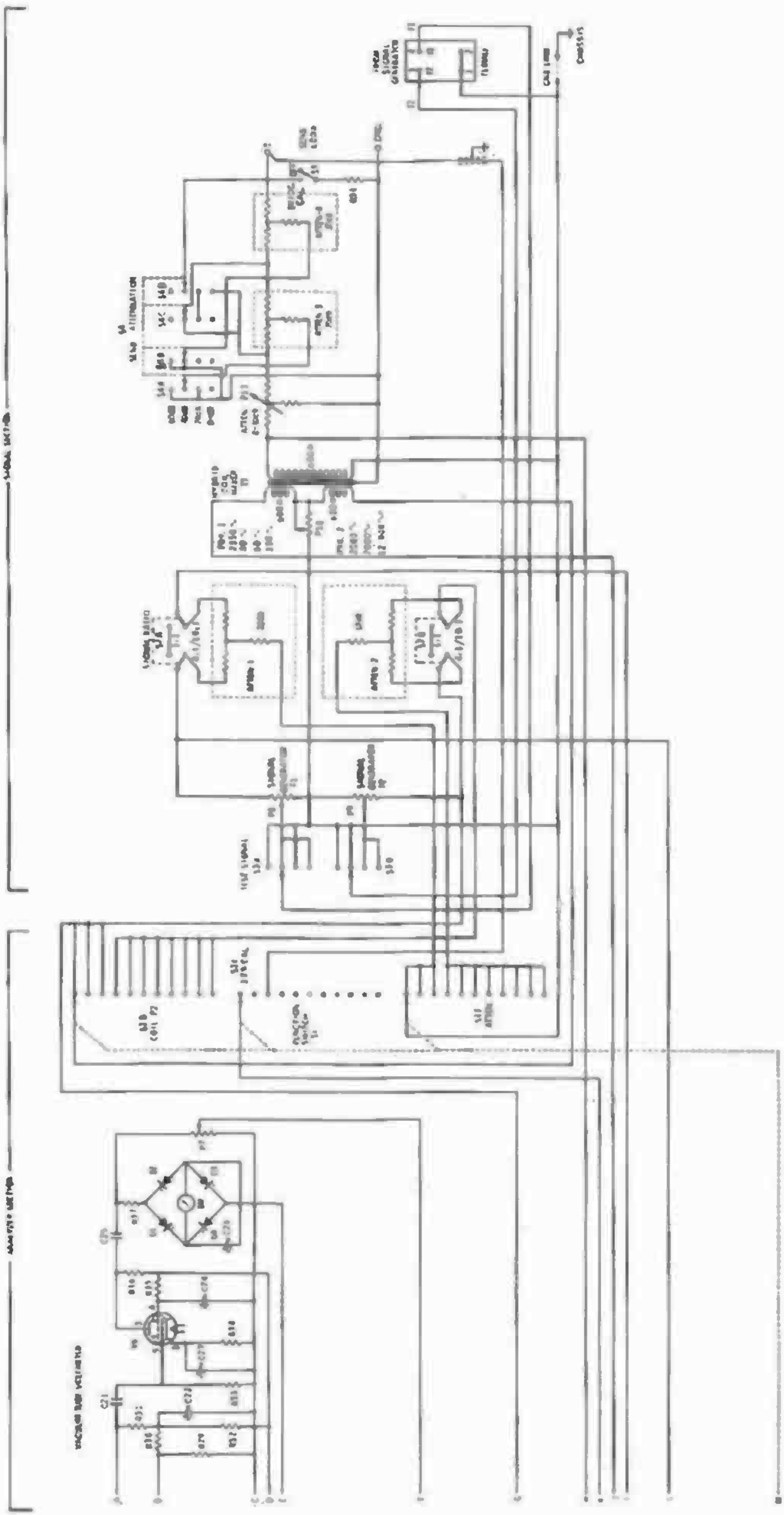


Fig. 22-131A. Schematic diagram of analyzer



section of an intermodulation analyzer.

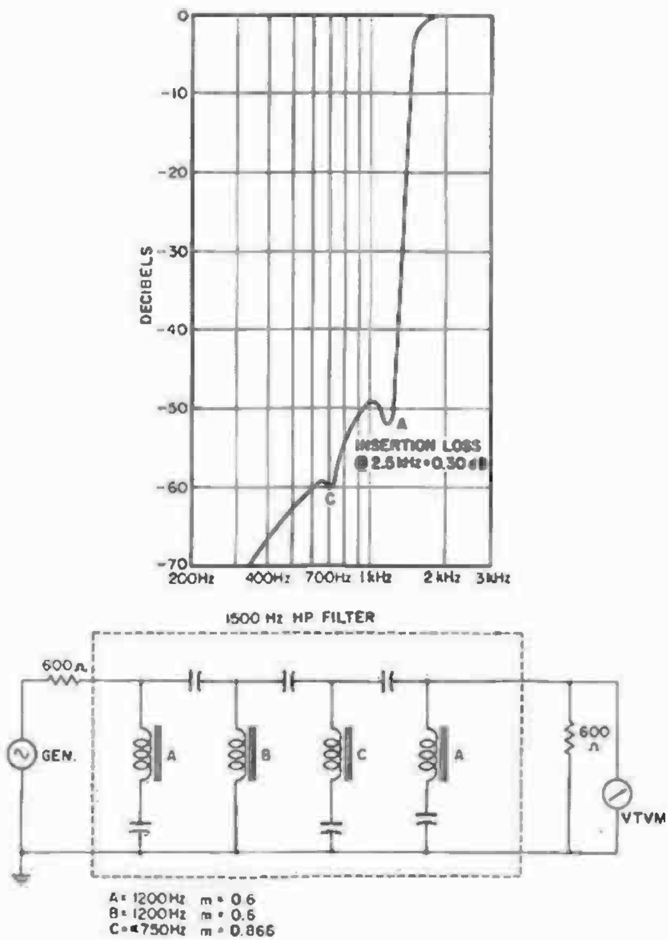


Fig. 22-131B. Configuration and frequency response of a 1500-Hz, high-pass filter.

uators and hybrid-coil circuit. Hybrid coils are discussed in Question 8.66.

Two attenuators, one of 20-dB loss and one of 12-dB loss, follow potentiometers P8 and P9. The 20-dB attenuator establishes the 10:1 signal ratio for the initial calibration (see Fig. 22-129M) and the 12-dB attenuator sets the 4:1 ratio for intermodulation testing. These attenuators are automatically switched in and out of the circuit by function switch S1.

The secondary of the hybrid coil is terminated by an attenuator network somewhat like the attenuator section of a gain set. Switch S4 permits four 20-dB steps to be selected while attenuator P11 provides attenuation from 0 to 30 dB, in steps of 1 dB. Terminating resistor R38 is used for devices with a bridging input. Operating voltages are supplied from the signal generator section discussed in Question 22.130. The residual intermodulation is less than 0.05 percent, due to slight unbalances in the hybrid-coil mixing circuits.

22.132 What is the difference between the SMPTE and CCIF methods of measuring intermodulation?—In the SMPTE method, two signals are applied to the input of the device being tested, one frequency lying between 40 and 100 Hz, and the other between 1000 and 12,000 Hz. The high-frequency signal is set 12 dB lower in amplitude than the low frequency; the ratio is 4:1.

Because of the nonlinearity of the device being tested, sum and difference frequencies will be generated and exist as sidebands. This is illustrated in Fig. 22-129D. The distortion may be defined:

$$\frac{\text{Nth harmonic of sidebands}}{\text{High-frequency amplitude}}$$

The CCIF method also uses two frequencies, but of equal amplitude, with

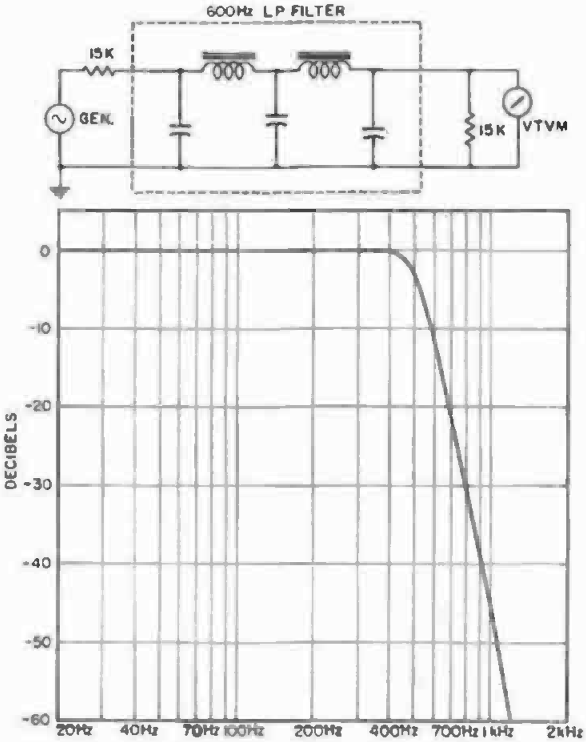


Fig. 22-131C. Configuration and frequency response of a 600-Hz, low-pass filter.

a difference in frequency somewhere between 30 and 1000 Hz. These frequencies are kept at a fixed difference and moved up and down the spectrum, producing a constant beat frequency, while $f_2 - f_1$ remains fixed. Intermodulation products are generated between each high frequency and the second harmonic of the other high frequency. The frequency spectrum for the CCIF system is shown in Fig. 22-132.

The intermodulation products are measured by means of a harmonic wave analyzer similar to those described in Question 22.65. The distortion may be defined:

$$\frac{\text{Difference-frequency amplitude}}{\text{Sum of high-frequency test signals}}$$

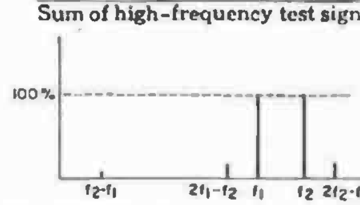


Fig. 22-132. Frequency spectrum seen using the CCIF method of measuring intermodulation. Compare this with the spectrum of the SMPTE system shown in Fig. 22-129D.

22.133 What is a difference-frequency intermodulation test?—It is another name for the CCIF method described in Question 22.132.

22.134 What is a decade amplifier and what is its purpose?—It is a precision amplifier with fixed steps of gain, generally 20 and 40 dB (amplification of 10 and 100). It is used for increasing the sensitivity of oscilloscopes, transistor and vacuum-tube voltmeters, and similar instruments. As a rule, such amplifiers have broad frequency characteristics, reaching into the megahertz region for bandwidth.

The schematic diagram for a transistor-type decade amplifier, manufactnred by Hewlett-Packard, is shown in Fig. 22-134A. The circuit consists essentially of a number of cascaded common-emitter amplifier stages. Additional stages have been added preceding and following the amplifier to give the instrument the desirable impedance characteristics of high and low output impedance with additional circuitry added to change the gain from 40 to 20 dB. An input clipping circuit, consisting of two diodes and a resistor R4, protects the amplifier from accidental overloads. Special biasing

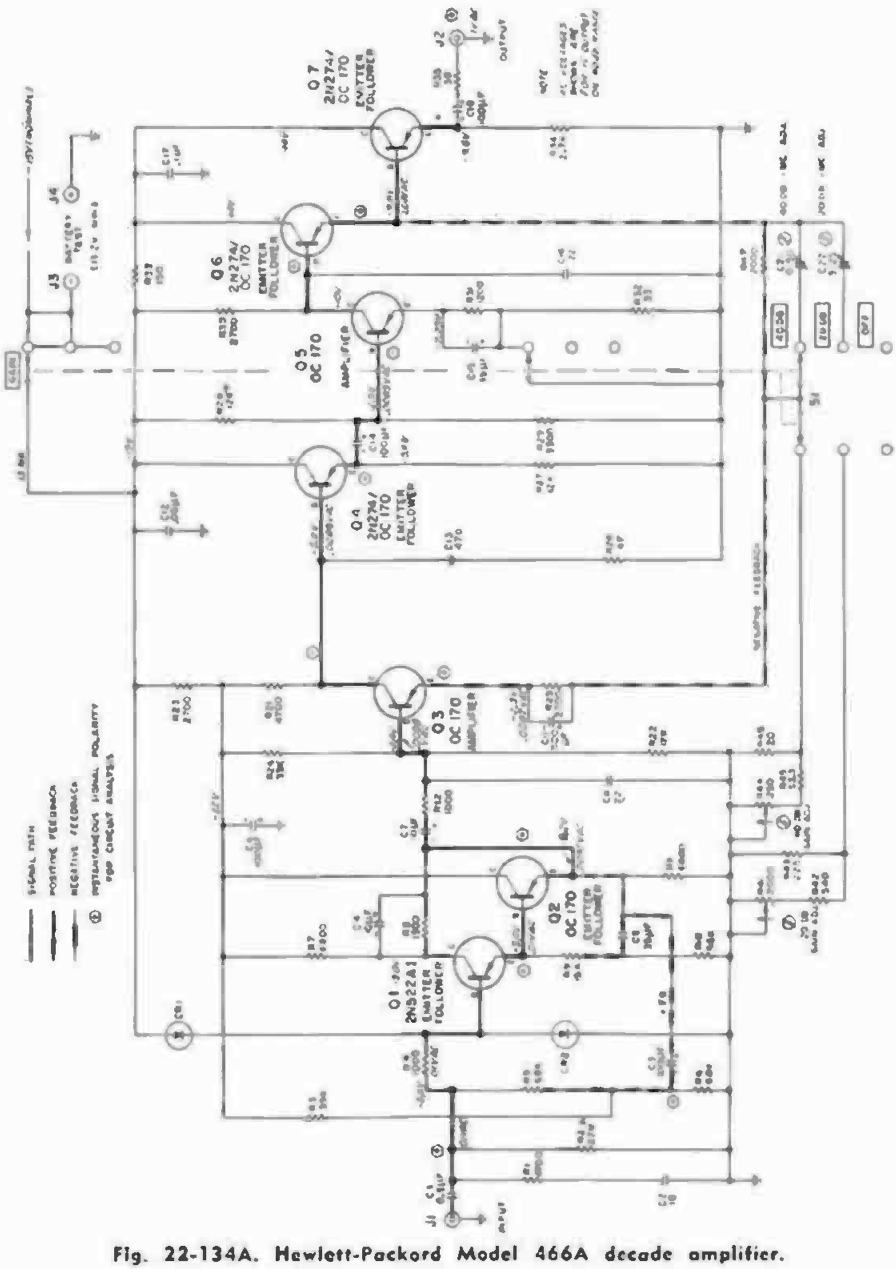


Fig. 22-134A. Hewlett-Packard Model 466A decade amplifier.

circuits are used to compensate for temperature and for varying characteristics between transistors of the same type.

The frequency response of this instrument is plus or minus 0.5 dB, 10 Hz to 1 MHz, dropping off to less than 3 dB at 2 MHz. The input impedance is 1 megohm shunted by 25 pF, with an internal output impedance of 50 ohms in series with 100 μF. Although the internal output impedance is 50 ohms, the external load must not be less than 1500 ohms. The internal noise is 75 microvolts for a 100,000-ohm source resistance. Distortion is less than 1 percent, 10 to 100,000 Hz, with less than 5 percent at 1 MHz. The device may be operated either from batteries or an ac power supply (Fig. 22-134B).

22.135 *What is a step generator?*— A device for testing the linearity of an amplifier by applying a step wave to the input of the amplifier and observing the stepped waveform at output of amplifier by means of an oscilloscope.

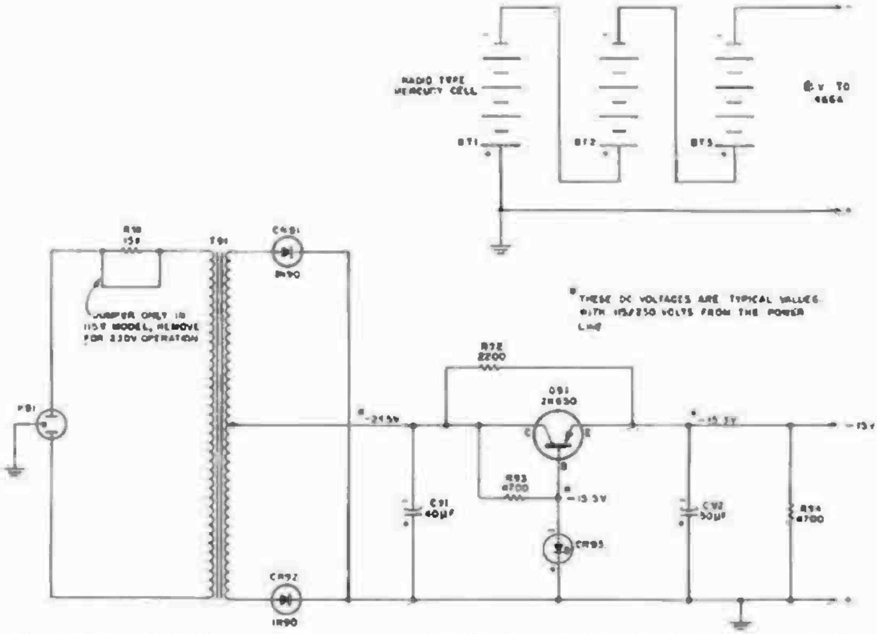


Fig. 22-134B. Battery and ac power supply for Hewlett-Packard decade amplifier Model 466A.

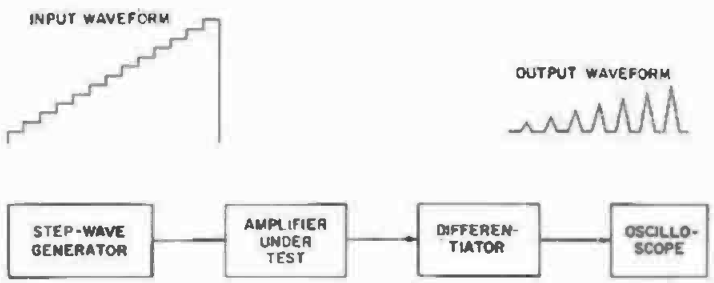


Fig. 22-135. Step generator for testing linearity of amplifiers by increasing input level in stepped values as shown by the input waveform.

If the amplifier is linear, the observed pulses on the oscilloscope will increase in height in a linear manner. If curved upward or downward, it is an indication of nonlinearity. The block diagram for such an instrument is shown in Fig. 22-135.

22.136 What is an octave-band noise analyzer?—An octave-band noise analyzer (Fig. 22-136A) is designed for simple and rapid analysis of noises having complex spectra. It may be operated directly from a microphone or fed from the output of a sound-level meter. It can be used for acoustic measurements of sound recording stages, auditoriums, offices, vibration studies, and all types of frequency analysis, except those requiring a detailed knowledge of the frequency spectra. In this latter in-

Fig. 22-136A. General Radio Co., Model 1558A octave-band noise analyzer.

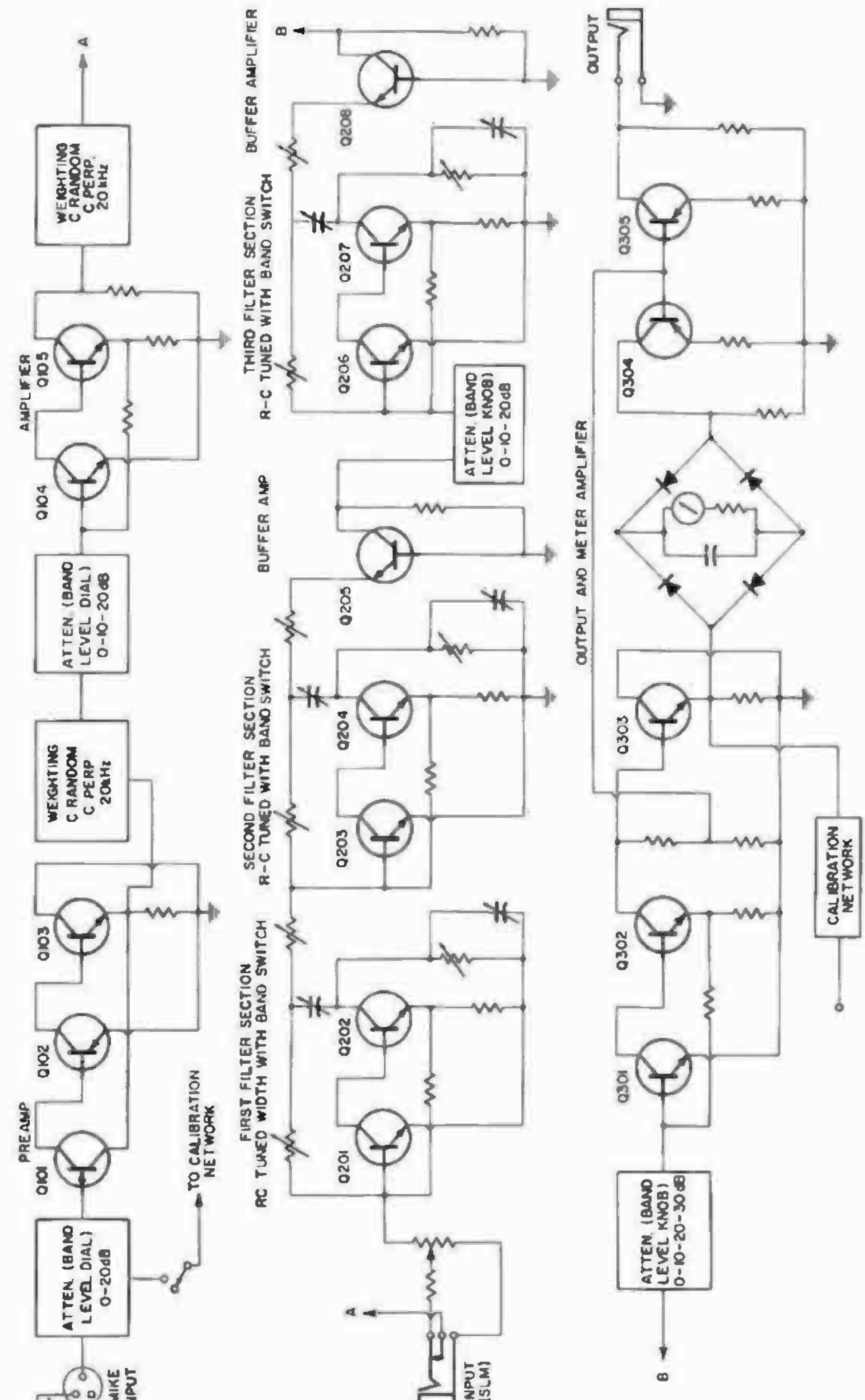


Fig. 22-136B. Elementary schematic diagram for General Radio Co., Model 1558A octave-band noise analyzer.

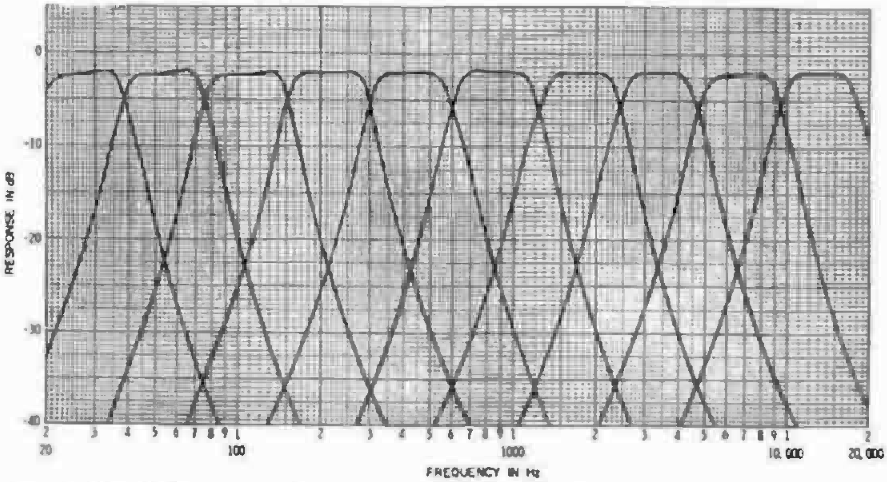


Fig. 22-136C. Frequency characteristics of octave-band filters in General Radio Co. Model 1558A octave-band noise filter.

stance, a harmonic wave analyer similar to those described in Question 22.65 is employed.

The basic components of an octave-band analyzer manufactured by the General Radio Co. are given in Fig. 22-136B. At the upper portion of the diagram is the circuitry for use with a microphone, which includes an attenuator, high-impedance microphone preamplifier, and weighting networks. If the instrument is to be driven from the output of a sound-level meter, the input signal is applied to jack SLM. The signal for either type of input is passed through a group of 1-octave bandpass filters with center frequencies ranging from 37.5 to 9500 Hz, as given in Fig. 22-136C.

At the center frequency, the response is uniform within 1 dB, with a maximum of 1 dB from the all-pass level (Fig. 22-136C). For the bands from 37.5 to 9600 Hz, the response at the nominal cutoff frequency is 3.5 dB below the response at the center frequency. For the all-octave bands, the attenuation is at least 30 dB at half the lower cutoff frequency and at twice the upper frequency. The attenuation is at least 50 dB at one-fourth the lower cutoff frequency and at four times the upper frequency. The 75-Hz low-pass filter has at least 30-dB attenuation at 200 Hz and at least 50-dB attenuation at 400 Hz. These characteristics are given in Fig. 22-136D. On leaving the filter section, the signal is again attenuated and fed to a metering circuit.

LOWER CUTOFF FREQUENCY Hz	UPPER CUTOFF FREQUENCY Hz	CENTER FREQUENCY Hz *
18.75	37.5	26.5
37.5	75	53
75	150	106
150	300	212
300	600	424
600	1200	848
1200	2400	1696
2400	4800	3392
4800	9600	6784
9600	19,200	13,570
LOW PASS	75	
ALL PASS		

* GEOMETRIC MEAN

Fig. 22-136D. Octave-filter frequency characteristics.

In using this meter the sound source is measured in octave bands which for acoustical measurements eliminates many difficulties due to backgronnd noise or reflections, as the spectrum is measured in bands rather than as a whole. Such instruments are widely used with white- and pink-noise measurements. The use of this instrument is not confined to acoustical measurements, but whenever octave-band measurements are necessary. (See Question 22.56.)

The normal sensitivity range of the instrument is 44 to 150 dB (re: 0.0002 μbar). However, the sensitivity may be extended by the use of a special plug-in microphone and preamplifier to 24 to 150 dB. Two frequency characteristics are provided: one essentially flat, and a weighted network characteristic (Fig. 22-136E). An output circuit is also provided for driving a magnetic tape re-

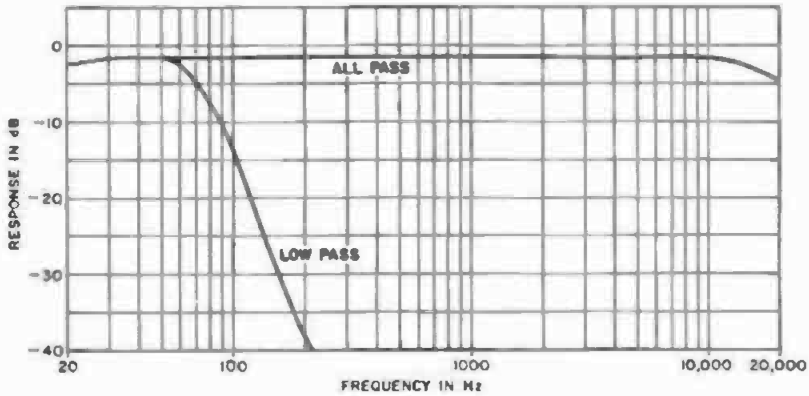


Fig. 22-136E. Low-pass and all-pass characteristics.


Fig. 22-137A. General Radio Co., Model 1559B microphone reciprocity calibrator.

corder, oscilloscope, graphic level recorder, or wave analyzer.

22.137 *Describe a reciprocity microphone calibrator.*—A reciprocity microphone calibrator is used to calibrate microphone sensitivity, and it is the means used by the United States National Bureau of Standards for such calibrations. Fig. 22-137A shows such a device, manufactured by the General Radio Co. The instrument includes a circuit and structure required for the close-coupler (cylindrical cavity) reciprocity technique, which is the preferred method of calibration for standard microphones. The instrument shown also includes an analog computer which performs the calculations necessary to determine the microphone sensitivity. The reversible transducer is a PZT ceramic microphone, and the auxiliary transducer a PZT cylinder which forms the cylindrical wall of the coupler (Figs. 22-137B and C). The insert voltage (test signal) is varied in 10-dB steps to extend the calibration range while maintaining a high resolution. The coupler has a volume of 17.74 cubic centimeters and is designed to yield the random incidence response of the microphone over an extended range without the use of helium.

The PZT ceramic cylinder also serves as a stable acoustic source when the instrument is used as a souud-level calibrator. A meter calibrated in terms of the sound-pressure level produced indicates the absolute value of the signal applied to the PZT cylinder. An exter-

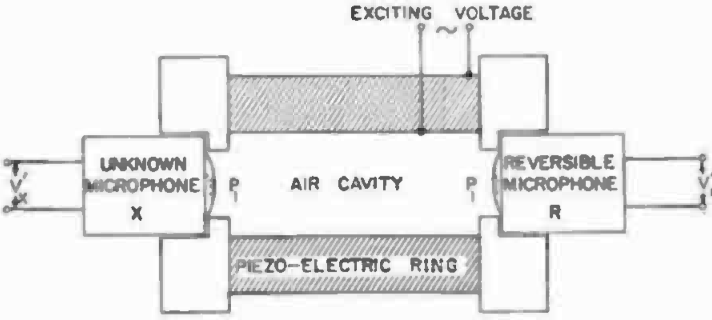


Fig. 22-137B. Conditions for determining the ratio of microphone sensitivities.

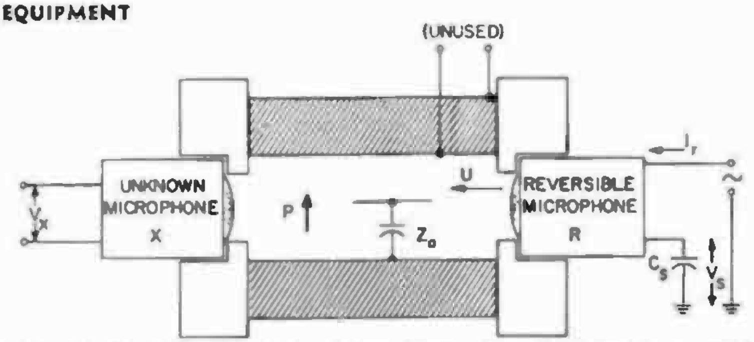
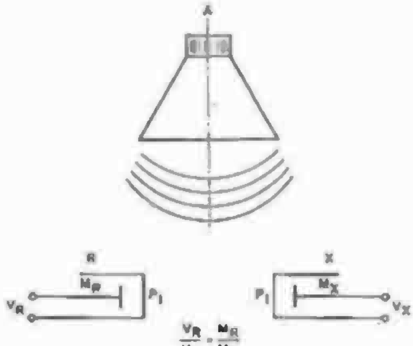


Fig. 22-137C. Conditions for determining the product of microphone sensitivities.



$$\frac{V_R}{V_X} = \frac{M_R}{M_X}$$

Fig. 22-137D. Relationship of three transducers in reciprocity calibration.

nal source of audio frequencies capable of developing a power of 1 watt at an impedance of 600 ohms is required. A detector is required to measure the output signal of the calibrator and should have an impedance of at least 5 megohms and be capable of measuring a signal level of several millivolts with a signal-to-noise ratio of 20 dB or greater. It should also have a meter capable of indicating 1-percent change in signal level. The detector may be a sound-level meter or an octave-band analyzer (usually the instrument under calibration). In making such calibrations, corrections for altitude and barometric conditions must be considered. A nomograph for this purpose is given in Section 25.

The cross-sectional view of the coupler (Fig. 22-137B) shows how the microphone to be calibrated is placed in the measuring cavity. The accuracy of the measurement depends on the (1) resistance values of the attenuator, (2) mechanical dimensions of the coupler cavity, and (3) the value of a current-sampling capacitor.

Transducer A (Fig. 22-137D) is used as a speaker which equally excites the unknown microphone X and a reciprocal microphone R with sound pressure. The ratio of the open-circuit voltages of the two microphones R and X equals the ratio of the microphone sensitivities. The ratio of the open-circuit voltage to the exciting pressure is the definition of the microphone sensitivity: $M = V/P$. The sensitivity is commonly expressed 20 Log_{10} M (re: 1 V/μbar). If the two microphones are coupled together by a known acoustical impedance (cavity) and the reciprocal microphone R is driven as a speaker, the ratio of the open-circuit voltage of microphone X to the driving current of reciprocal microphone R can be related theoretically to the product of the microphone sensitivities. By solving two relationships—the ratio of sensitivities of microphones R and X, and the product of these sensitivities—the sensitivity of either microphone can be determined.

22.138 ***Give the principles of reciprocity testing.***—The principle of reciprocity is stated as follows: The ratio to excitation is unchanged if the points of excitation and observation are interchanged, provided the terminal conditions remain the same. The equation in Fig. 22-138 illustrates the relationship that follows from the principle of reciprocity for a two-port, four-terminal electrical network. The design of the General Radio Co. Model 1559B microphone reciprocity calibrator of Fig. 22-137A is based on the forward current transfer equalling the reverse voltage transfer.

The previously described instrument can be used to calibrate microphones in the range from minus 75 dB to minus 35 dB (re: 1 V/μbar.) Accuracy is plus or minus 0.2 dB.

Reciprocity is not restricted to linear and passive networks: it applies to any linear, bilateral, or passive network. The reciprocity technique for microphone is

$$\frac{I_2}{I_1}\Big|_{V_2=0} = \frac{V_1}{V_2}\Big|_{I_1=0}$$

FOR A TWO-PORT, FOUR-TERMINAL NETWORK

$$V_1 = Z_{11}I_1 + Z_{12}I_2 \qquad V_2 = Z_{21}I_1 + Z_{22}I_2$$

BY RECIPROCITY,

$$Z_{12} = Z_{21} = 2$$

$$\frac{I_2}{I_1}\Big|_{V_2=0} = \frac{V_1}{V_2}\Big|_{I_1=0} \qquad \frac{I_1}{I_2}\Big|_{V_1=0} = \frac{V_2}{V_1}\Big|_{I_2=0} = \frac{2}{3}$$

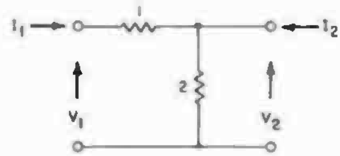


Fig. 22-138. The principle of reciprocity applied to a two-port four-terminal network.

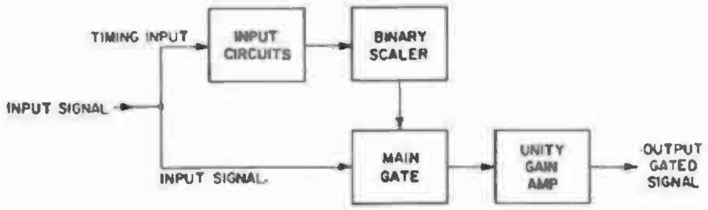


Fig. 22-139A. Basic block diagram for General Radio Co. tone-burst generator Model 1936A.

Fig.22-139B. General Radio Co. tone-burst generator, Model 1936A.

a primary method of calibrations in which acoustic standards are required. This method of calibration is set forth in USASI (ASA) Standard Z24.4-1949.

22.139 ***Describe a tone-burst generator.***—Tone-burst generators are used for a variety of transient tests and fall between continuous tone testing and step-function or pulse testing. Tone-burst generators find diverse applications. One of the most useful features of such instruments is the output signal can be coherent; that is, the gate can be made to open and close at the same point in a signal voltage cycle for each tone burst. Coherence is of value for two reasons. First, the coherent tone burst is much more easily observed or analyzed by oscillographic or sampling methods and second, the frequency content of the tone burst depends on the phase of the gate compared to the signal. If the signals are not coherent, the phase, and hence the frequency content, is drifting, and test results are less reproducible or less clear than with a coherent tone burst. Another useful feature of these devices is the control of the number of cycles in the tone burst, which permits the production of tone bursts with consistent and controllable frequency content.

Some of the uses to which this instrument can be put are the testing of acoustic enclosures, speakers, rectifying circuits, recovery time of various devices, amplifier music-power testing, speaker-distortion measurements, gen-

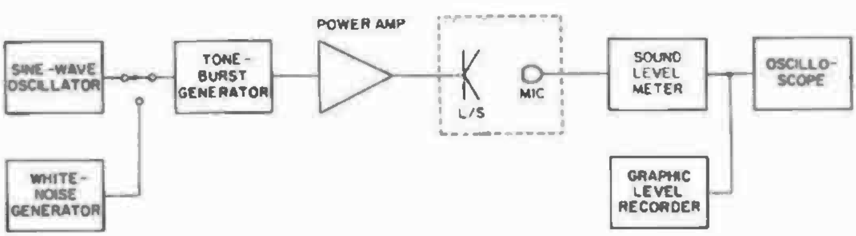


Fig. 22-139C. Tone-burst generator connected in a test setup for measuring the acoustical characteristics of an enclosure.

Fig. 22-140A. Voltage-controlled oscillator (VCO) Model 610B manufactured by Waveforms Inc.

Fig. 22-140C. Logarithmic ac volt and frequency meter Model 620A manufactured by Waveforms Inc. An external frequency source is required with this instrument.

eration of power-line transients, and filter and equalizer testing, to name but a few.

An elementary block diagram of a tone-burst generator, manufactured by the General Radio Co., is shown in Fig. 22-139A. The tone-burst generator requires an external source of audio frequency and may be supplied by any audio oscillator. The audio signal is gated by the tone-burst generator, which has provisions for timing the opening and closing of the gating circuitry. As will be noted, the audio signal and the timing signal are the same.

The main gate is a transmission gate of the shunt type which accurately reproduces the input signal at the gated signal output terminals when the gate is open and permits no output signal when the gate is closed. The opening and closing of the main gate is controlled by a gating signal from a binary scaler. The scaler counts once per cycle of the timing signal, and can be controlled to set the gate-open and gate-closed intervals.

A front-panel view of a tone-burst generator is shown in Fig. 22-139B. A typical test setup for making an acoustical measurement of an enclosure is shown in Fig. 22-139C.

22.140 ***Describe an automatic frequency-plotting system.***—In the making of daily routine measurements in large recording plants, many measurements of frequency response are made. If these are made manually, several hours a day can be consumed. The solution to this problem is making measurements automatically and recording a graph that may be stored away to be referred to later as an analysis of the stability of such equipment. Automatic plotting is not confined entirely to the making of amplifier measurements, but can be used for almost any equipment. Tests may include the measurement of impedance, amplifier frequency response, acoustical enclosures, amplitude and phase balance, comparative tests, constant percentage frequency modulation, phonograph pickups, tape recorders, and many others. The response is plotted on an X-Y plotter, which is explained in Question 22.112.

For making such measurements a device pictured in Fig. 22-140A and

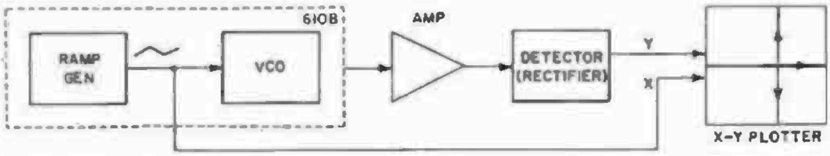


Fig. 22-140B. Waveforms Inc. Model 610B voltage-controlled oscillator (VCO) connected for making automatic frequency response measurements on an amplifier using an X-Y plotter.

manufactured by Waveforms, Inc., may be used. This instrument is a voltage-controlled oscillator (VCO) with its frequency controlled by an internal or external ramp voltage. A 100:1 tuning range may be selected between 2 and 2000 Hz and 20 to 20,000 Hz; thus the flexibility of an electrical nonmoving part can be achieved. The voltage may be one of several, viz, the internal sawtooth generator and internal potentiometer for single-range manual sweep, the potentiometer of an X-Y plotter to slave the VCO to the plotter, or other sources of control voltage.

An internal sweep-frequency control generator produces a positive ramp output voltage (straight-line voltage with a rising characteristic) which varies between 1 and 10 volts. A 1-volt value corresponds to the VCO lowest frequency limit and 10 volts to its highest (1000 times the lowest). Since the characteristics of the VCO are approximately logarithmic frequency versus voltage, the sweep is logarithmic, which is the plotting characteristic of most audio-frequency equipment. The ramp control voltage is also available on the front panel and may be used to drive the X axis of a plotter or an oscilloscope. The sweep time may be varied from 6 to 60 seconds. A typical setup for measuring the frequency response of an amplifier is given in Fig. 22-140B, using an X-Y plotter. The ramp control voltage determines the VCO frequency and controls the X position of the plotter pen. The horizontal plot (X) is in three logarithmic frequency decades. The vertical position (Y) is determined by the magnitude of the voltage at the output of the device under test. The output voltage is rectified so the plotter Y drive is a dc voltage directly proportional to the amplitude of the output signal. The result is a plot of amplitude versus frequency. The device shown includes a logarithmic frequency meter and an attenuator system with 100-dB loss, which may be used externally if desired. The reset time is about 2 seconds and always resets to 20 Hz. The response is plus or minus 0.5 dB. The power output is 2.5 volts in a 600-ohm line, with THD of 1 percent at 1000 Hz.

A somewhat similar instrument is shown in Fig. 22-140C, containing a logarithmic voltmeter and frequency meter and the necessary rectifying circuits for directly driving an X-Y plotter. However, an external source of frequency is required, which may be a prerecorded magnetic tape, record, film, or oscillator.

22.141 ***Describe the basic circuitry for a transistor test set.***—Transistor test sets are manufactured in two types: in-circuit, and for general bench testing. The first instrument to be discussed is the Model T.T.164 in-circuit transistor tester, manufactured by AVO, Ltd. (England). A front-pauel view of the instrument is given in Fig. 22-141A. This instrument is portable and may be used for in-circuit testing for both pnp and npn transistors. In addition, two low-current ranges are provided for the measurement of leakage current in silicon transistors, which is considerably lower than in other types. These two ranges are amplified by a single transistor amplifier.

Current gain (beta) may be measured with the transistor in or out of

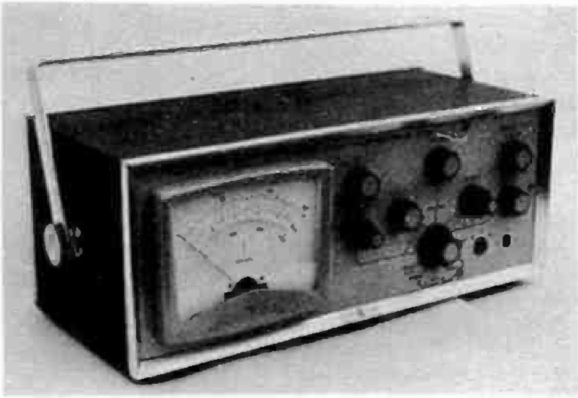


Fig. 22-141A. AVO Ltd (England) in-circuit transistor tester.

the circuit over a range of 0 to 150 and to 300, provided the base to the emitter loading is over 400 ohms and the collector impedance is greater than 25 ohms. Collector current is continuously variable from 0 to 10 μA in steps of 10 mA to 30 mA, with continuously variable collector voltage of from 0 to 10 volts.

A dc bridge circuit (Fig. 22-141B) enables in-circuit components to the transistor to be balanced out before the dc operating conditions are set. Measurement of beta is carried out at frequency of 1000 Hz by means of an ac bridge, after the effect of impedance in the base circuit is balanced out. The ac collector current is monitored by measuring the voltage across a 1-ohm resistor in the collector circuit.

The indicating meter is calibrated to read 0 to 150 for measurements of beta, 0 to 10 for measurement of collector current and two current ranges of 100 μA and 1 mA for leakage measurements, and 0 to 300 for leakage currents below 3 μA. There is also a battery check scale. Overload protection is also provided. For in-circuit testing, special probes are required. Power is supplied from four 9-volt batteries regulated to supply a 20-volt low-impedance output.

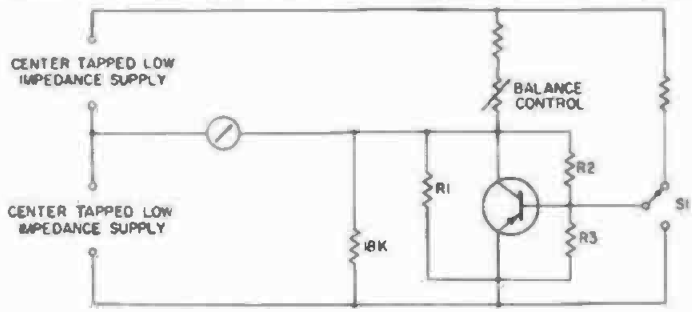


Fig. 22-141B. Basic dc balance circuit for AVO in-circuit transistor tester.

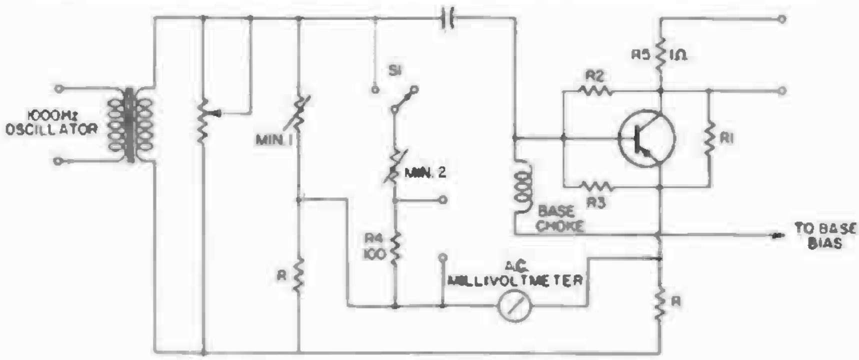


Fig. 22-141C. Basic ac balance circuit for AVO in-circuit transistor tester.

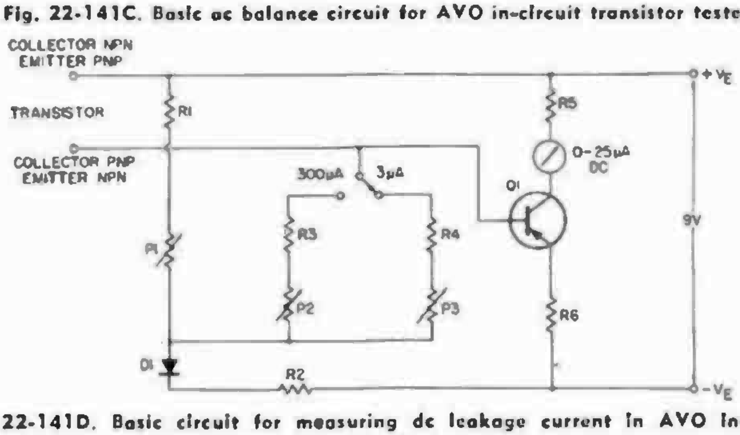


Fig. 22-141D. Basic circuit for measuring dc leakage current in AVO in-circuit transistor tester.

Fig. 22-141E. Heathkit Model IM-30 transistor tester.

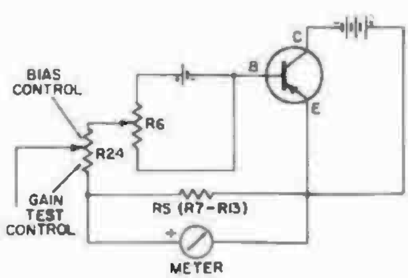


Fig. 22-141F. Base current measurement.

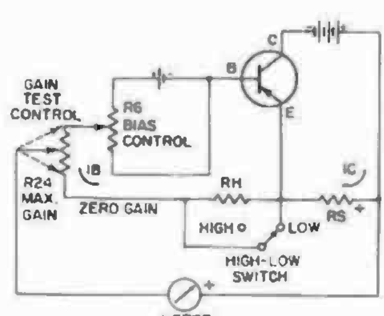


Fig. 22-141G. Gain test (dc Beta and dc Alpha).

center tapped to provide 10 volts on either side.

With switch S1 in position 1, the shunt paths are represented by resistors R1, R2, and R3, which are balanced out with the transistor not functioning. After balancing, the switch is placed in position 2 and the collector current is adjusted.

Referring to the ac bridge circuit in Fig. 22-141C, the circuit is balanced first with S1 open, using control minimum-1, and the transistor inoperative. This is to nullify the effect of shunt paths around the transistor. With switch S1 closed, the bridge is again balanced, using the minimum-2 control and with the transistor operating normally. Ac base current is balanced out in this operation by measuring the voltage drop across the 100-ohm resistor R4. The ratio of the voltage across the 100-ohm resistor to the voltage across 1-ohm resistor R5 in the collector circuit is a measure of current gain.

Leakage-current measurements employ a simple amplifier, as shown in Fig. 22-141D. Germanium diode D1 together with resistor R5 applies a forward bias voltage of approximately 0.5 volt to the base of high-gain transistor Q1. This ensures not only the linearity of the instrument calibration, but also that the calibration starts at zero. The zero setting is a factory adjustment; however, an internal potentiometer provides a zero adjustment should readjustment be necessary. Preset potentiometers P1 and P2 provide adjustment for the leakage-current ranges.

A second transistor tester, manufactured by Heathkit, is shown in Fig. 22-141E. This instrument tests transistors and diodes under conditions corresponding to actual operating conditions. Ac operating conditions are readily found by testing the transistor for two different values of bias, and then calculating the ac operating conditions. The

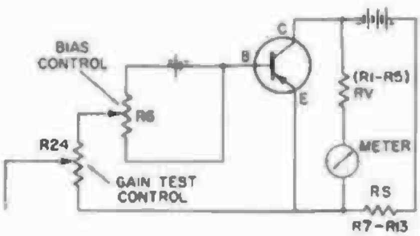


Fig. 22-141H. Collector voltage test.

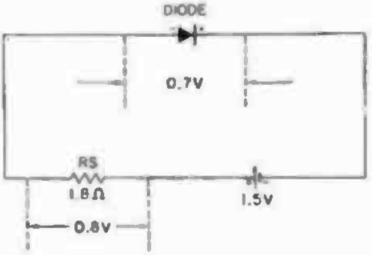


Fig. 22-141I. Diode test.

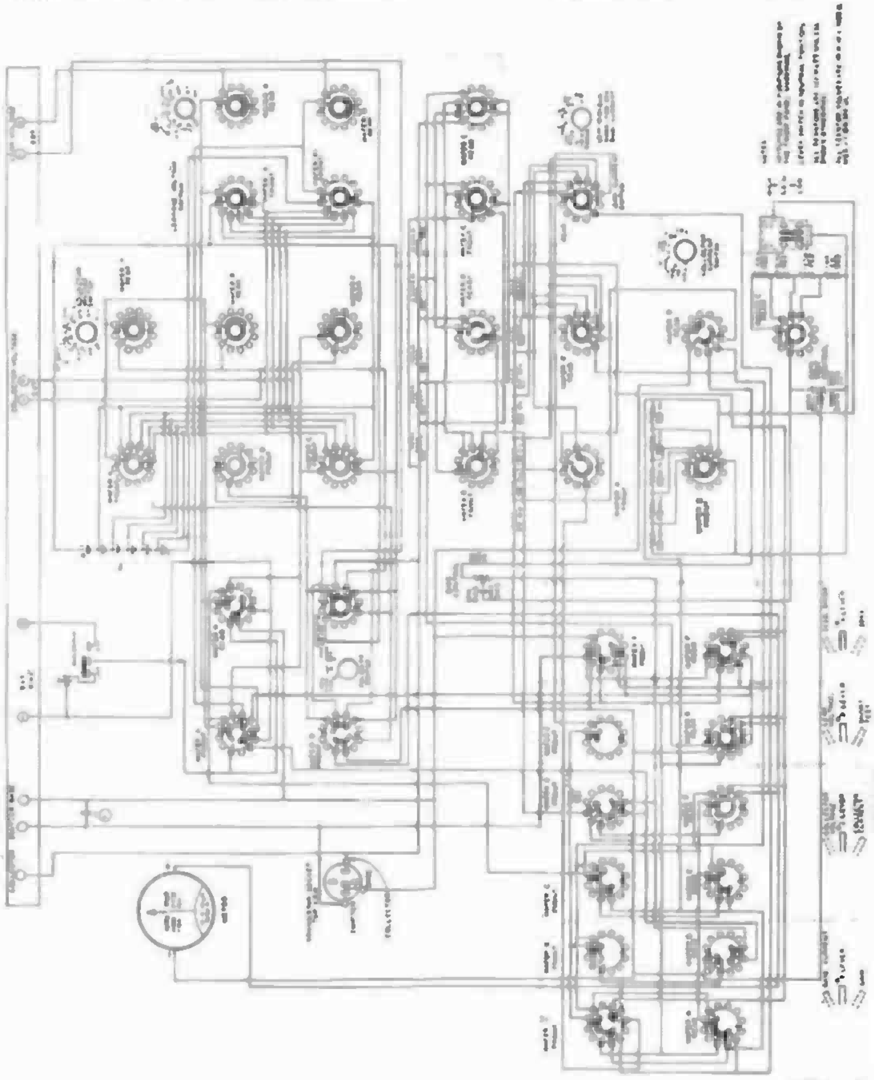


Fig. 22-141J. Schematic diagram of Heathkit Model IM-30 transistor tester.

various diagrams show the basic circuits which are automatically set up by the operation of certain designated switches.

Fig. 22-141F is the basic circuit for measuring base current (I_b) and is indicated by the meter connected directly in the base circuit. Gain (dc beta and dc alpha) is measured by the circuitry in Fig. 22-141G. Here the meter measures the voltage drop across resistor R_s and that across gain test control R24. By adjustment of R24 for a null indication on the meter, the voltage drop across Rs equals the voltage drop across R24. Therefore I_c (collector current) times R_s equals I_b (base current) times R24. Gain and beta are directly proportional to the value of R24, which is calibrated to read values of alpha or beta directly. Alpha equals beta divided by 1 plus beta.

Collector voltage E_c (Fig. 22-141H) is the voltage between collector and emitter. This voltage is selected by the collector voltage switch. Resistors R1 to R5 represent the meter multipliers. This voltage is also measured under operating conditions. Other tests include collector current (I_c), collector-emitter leakage (I_{ceo}), diode leakage tests, collector-to-base leakage (I_{cbo}), and internal short tests.

The various operating conditions may be calculated:

Dc current gain $h_{FE} = I_c/I_b = h_{fe} = \beta_{dc}$ $\quad \alpha = \beta_{dc}/(\beta_{dc} + 1)$

Ac current gain $= \dfrac{\Delta I_c}{\Delta I_b}$ (E_c held constant) or $\dfrac{I_{c1} - I_{c2}}{I_{b1} - I_{b2}}$ (at the same value of E_c)

Dc transconductance $G_{fe} = I_c/E_b$

Ac transconductance $G_{fe} = \dfrac{\Delta I_c}{\Delta E_b}$ (E_c held constant) or $\dfrac{I_{c1} - I_{c2}}{E_{b1} - E_{b2}}$

Dc base resistance $= E_b/I_b$

Ac base resistance $= E_b/I_b$ (E_c held constant) or $\dfrac{E_{b1} - E_{b2}}{I_{b1} - I_{b2}}$

Dc collector resistance $= E_c/I_c$

Ac collector resistance $= E_c/I_c$ (I_b held constant) or $\dfrac{E_{c1} - E_{c2}}{I_{c1} - I_{c2}}$

Symbols for transistor characteristics are given in Question 11.115.

The circuitry for testing diodes is given in Fig. 22-141I. Resistance R_s is a current-limiting resistance and must always be connected externally in series with the diode. Its value depends on the current rating of the diode. Silicon diodes generally have an internal voltage drop of about 0.7 volt. Using a battery voltage of 1.5 volts and using 0.8-volt drop, $R = E/I$. Thus, for a diode rated 500 mA, the series resistance would be 0.8/0.5, which equals 1.6 ohms. The resistor may be left in the circuit when measuring reverse currents as its value compared to the reverse current resistance is negligible. The schematic diagram for this tester is given in Fig. 22-141J.

22.142 Describe the basic principles of a solid-state electronic counter.—Frequency counters are extremely useful in the laboratory and for general calibration work. Heretofore such instruments have been quite bulky because of the large number of vacuum tubes or transistors required for their operation. The counter pictured in Fig. 22-142 is a Model 5215A, 12.5-MHz solid-state general-purpose counter manufactured by Hewlett-Packard. The circuitry of this instrument employs integrated circuits (IC) for its principal components, thereby reducing the physical size, cost, weight and power consumption. The total power required for this instrument is 20 watts, or less than one-tenth that previously required for vacuum-tube models. The weight of this instrument is 7 pounds as compared with 50 to 150 pounds for older models.

The counter shown displays the readout using seven digits made visual by seven cold-cathode Nixie display tubes manufactured by the Burroughs Corp. When the unit is in operation a blanking circuit suppresses the display of unwanted zeros to the left of the most significant digit. An internal 10-MHz crystal controlled self-checking oscillator is included as a time base to ensure that the decade counters, gates, function selector, amplifier, and time base are always operating satisfactorily. The nor-

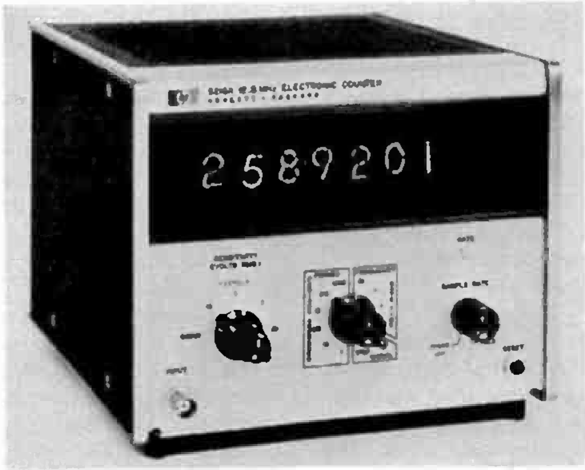


Fig. 22-142. Front panel view of Hewlett-Packard Model 5215A 12.5-MHz electronic counter.

mal display time ranges from 50 milliseconds to 5 seconds and may be held until reset. The accuracy is plus or minus 1 count, plus or minus the time-base accuracy. Gate times range from 0.01 to 10 seconds. Input impedance is approximately 1 megohm shunted by 50 pF. Sensitivity is 10 millivolts, rms sine wave. The input signal may be attenuated in decade steps ranging from 0.01 to 10 volts, with the triggering voltage continuously variable.

In addition to the counting operation, time intervals of 10 μs to 10 seconds may be measured, or periods averaged. By the substitution of an external signal source (f_1) for the internal time base, and connecting a second source (f_2) to the counter input, the frequency ratio f_1/f_2 can be measured. Terminals are provided on the rear apron for a 1-MHz, 3-volt output frequency as well as a four-line BCD code as a standard, with assigned weights of 1, 2, 4, 8 (1 state positive respect to 0 state).

In operation the instrument is connected directly across the circuit of which the frequency is to be measured and read in either kilohertz or megahertz. A reset push button on the front panel resets the display to zero. The counter is then ready for counting a new cycle.

Also included on the rear apron is a switch for disabling the storage feature. The display storage provides a continuous visual display while the instrument is totalizing a count. Only if the count differs from the previous count will the display change. The storage switch also disables the zero blanking circuits.

22.143 ***Describe the circuitry of a solid-state digital multimeter.***—Fig. 22-143A is a front-panel view of a Fairchild Instrument Division Model 7050 solid-state digital multimeter, with an interior view of its construction given in Fig. 22-143B. The instrument consists of a single chassis housing a power supply, counting circuits, and readout tubes. It may be operated from a line voltage of 117 Vac, 50 to 400 Hz. The instrument measures both dc voltages and resistance. For current measurement, external shunt resistances are required.

Four voltage ranges are provided, ranging from 1.500, 15.00, 150.0, and 1000 volts full-scale. The 1.5-volt scale has a resolution of 1 mV. Resistance measurements range from 1.500K, 15.00K, 150.0K, and 15.00 megohms. The 1.500 kilohm scale has a resolution of 1 ohm. For normal room temperatures,

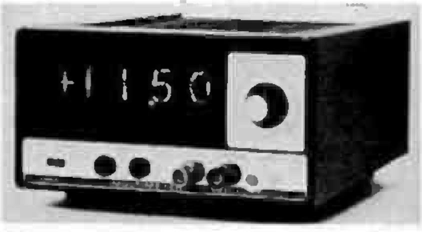


Fig. 22-143A. Front panel view of Fairchild Instrumentation Model 7050 solid-state digital multimeter, using dual-slope integration techniques with 0.1-percent accuracy.

the scale accuracy is plus or minus 0.1 percent of the reading plus or minus 1 digit; for the kilohm scale, plus or minus 0.2 percent of reading plus or minus 1 digit. Three digital readout indicators are provided with one overrange digit, decimal-point indicators, and a plus or minus polarity indicator. The readout display is steady without blinking. The average instrument of this type requires about 60 solid-state devices for its operation.

Each measurement cycle consists of two parts. During the first part, current is drained from an integrating capacitor by the unknown current, while clock pluses feed continuously into the counter. At 4000 counts from the clock the counter overspills to 0000 and the second part of the measurement cycle begins. At overspill (t_1 on the timing waveforms of Fig. 22-143C) a full-scale pulse is emitted. This pulse triggers flip-flop FF1 which in turn closes a reference-current solid-state switch. The integrating capacitor now recharges at a rate proportional to the reference current minus the unknown current (t_1 to t_2). When the voltage across the integrating capacitor reaches zero (t_2), the level detector changes the state of FF1, then turns off the reference-current switch and activates a strobe one-shot multivibrator. These actions temporarily disable the clock circuit and gate the buffer storage units, permitting the count stored at that particular instant to be applied to the readout tubes for display. The input, measurement, and clock circuitry are calibrated to result in a direct visual readout of the unknown voltage. The clock circuit delayed just long enough to prevent a count during the strobe pulse. By placing a short between the hold terminals, the readout voltage may be held until the short is removed.

Referring to both the block and schematic diagrams of Figs. 22-143C and D, the voltage to be measured is applied between the volts and common terminals on the front panel. Input signals are routed to a function/range switch consisting of six-ganged wafers. Wafer S1C switches the input common when the function switch is actuated. Wafer S1F sets the range indication by grounding the cathode element of the proper decimal-point indicator in the readout tubes. Section S1D grounds the cathodes of the polarity indicating lamps when the function/range switch is in any voltage range. The remaining wafers A, B, and E route the input signal voltage through a group of attenuating resistors to the input amplifier. Potentiometer R45 is for calibrating the 1000-volt range; R36, the 150-volt range; and R28, the 15-volt range.

The attenuated input signal is applied to a chopper-stabilized amplifier. The chopper is driven at 120 Hz from solid-state switch Q13. The drain element of Q16, a 2N4142 MOSFET is tied to the input of the ac carrier amplifier, with the source element connected to

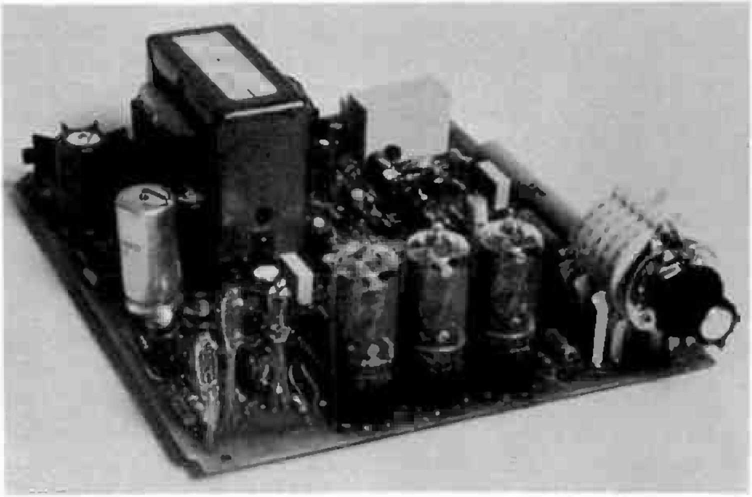


Fig. 22-143B. Interior view showing the construction of Fairchild Instrument Model 7050 solid-state digital multimeter. The three tubes at the front are the digital readout tubes.

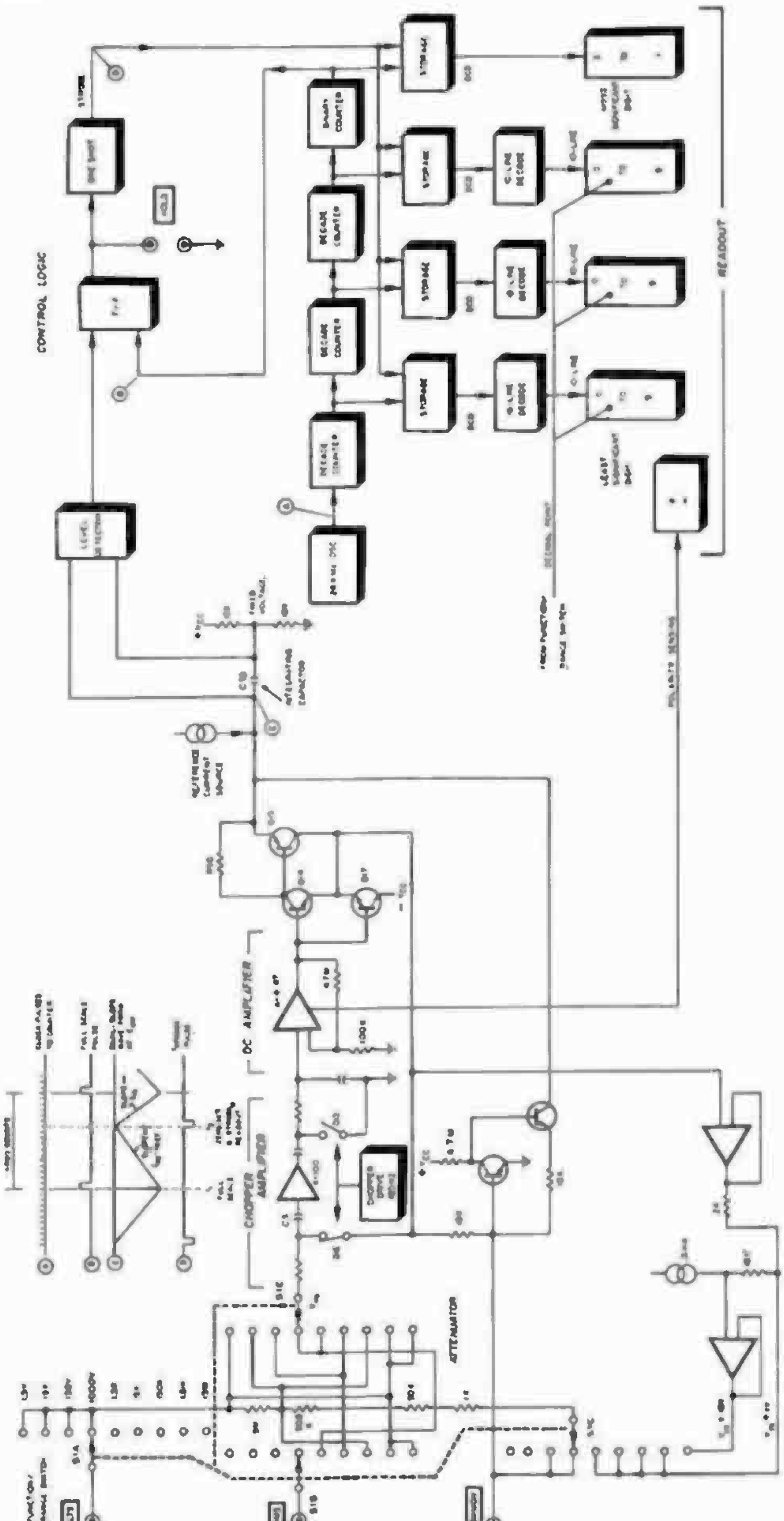


Fig. 22-143C. Block diagram for Fairchild Instrumentation Model 7050 solid-state digital multimeter.

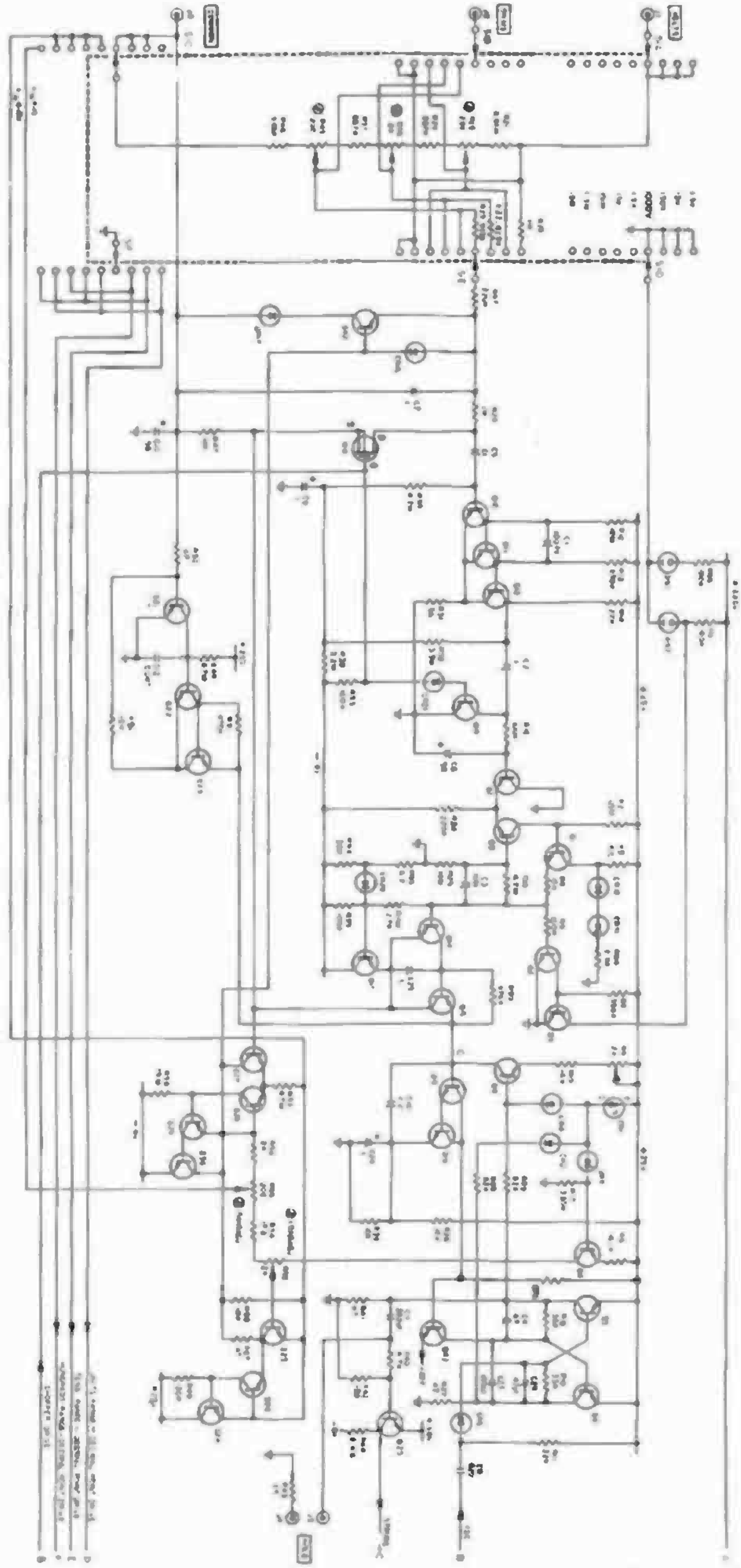
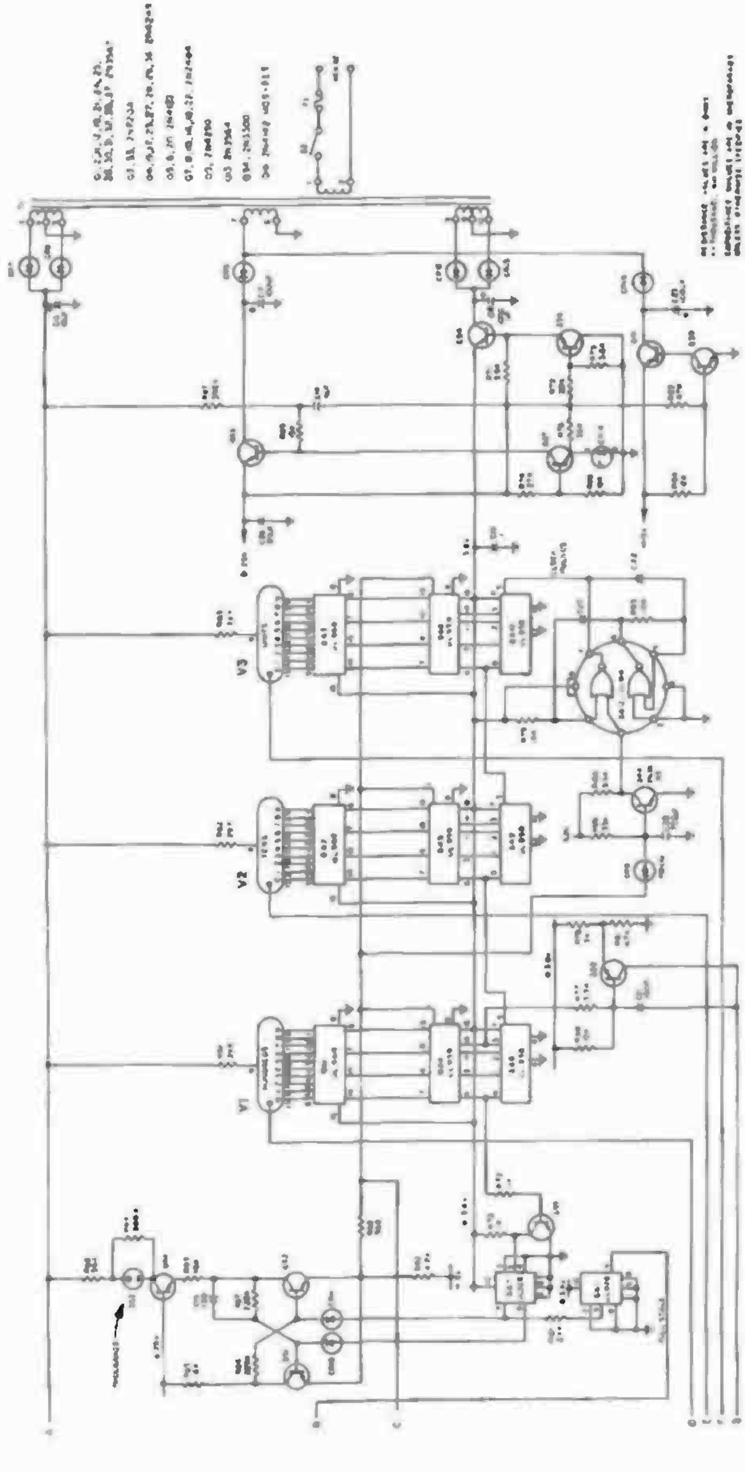


Fig. 22-143D. Schematic diagram for Fairchild Instrumentation, Model 7050 solid-



state digital multimeter.

the output of the dc amplifier for a closed loop. When the chopper is operating, any potential difference that exists between the amplifier input and output is coupled through capacitor C5 to the amplifier input as an error-signal voltage. This voltage is amplified and fed back in phase to the source element of Q16. When the voltages at the dc amplifier output and ac carrier amplifier input reach the same level, the amplifier is stabilized and the total equivalent unknown input voltage is impressed across 10,000-ohm resistor R47. The current through R47 is drawn from the charge on integrating capacitor C10.

When chopper transistor Q16 is in the on position, the voltage at its drain element (V_{in}) is chopped to the level of its source. The voltage step (positive or

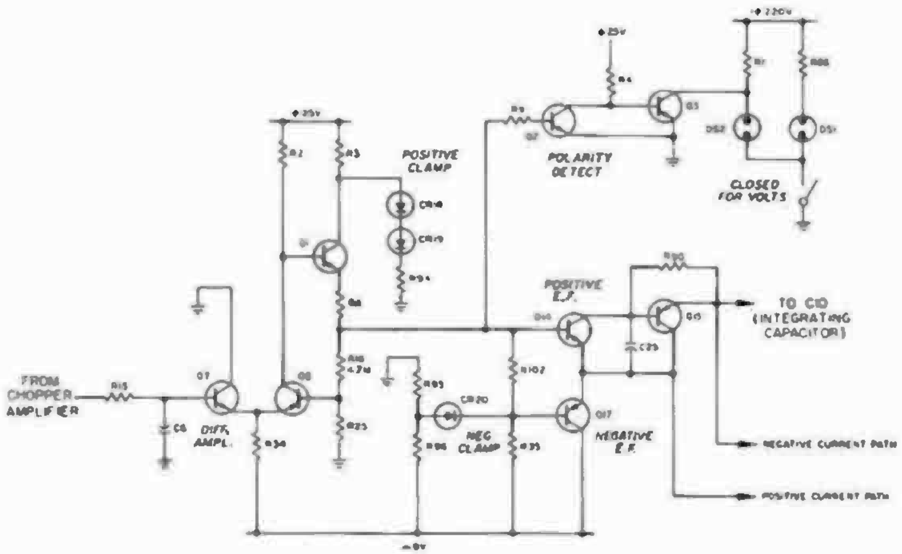
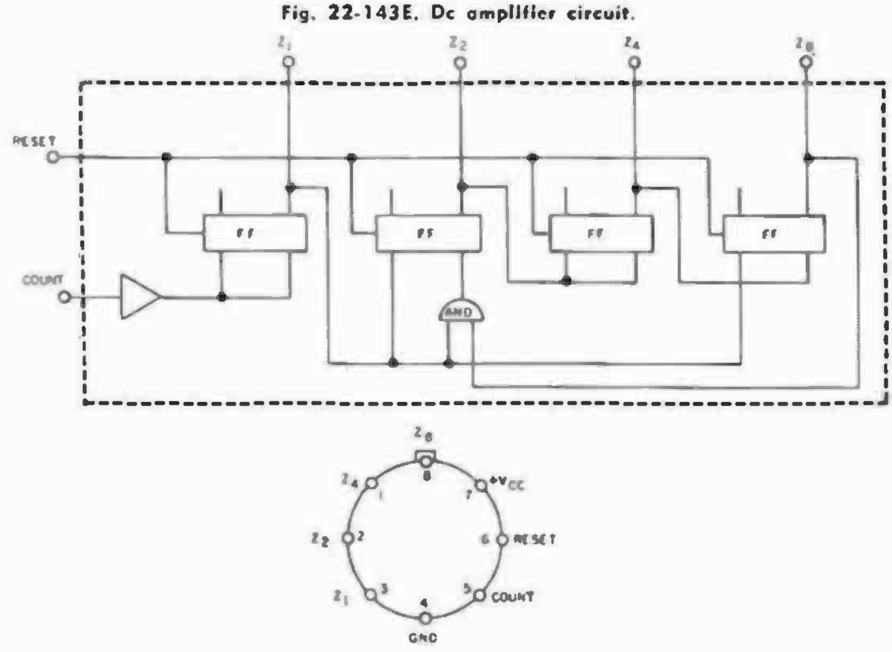


Fig. 22-143E. Dc amplifier circuit.



BASE CONNECTIONS

Fig. 22-143F. FS CUL 958 counter and base connection.

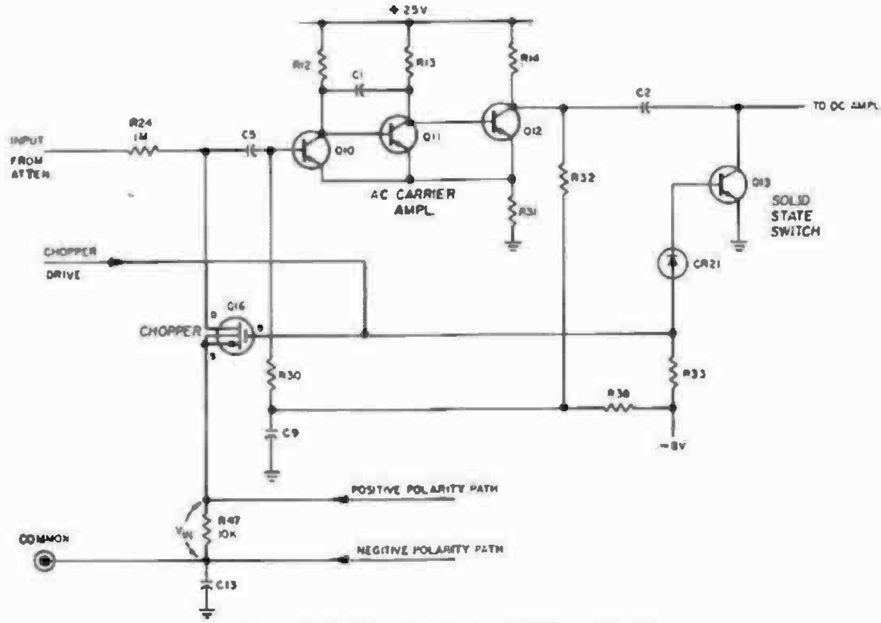


Fig. 22-143G. Chopper-amplifier circuit.

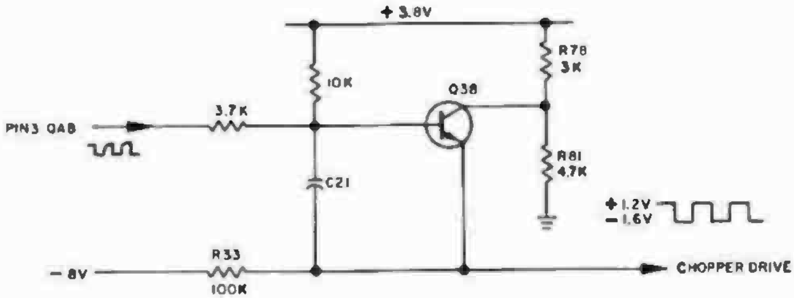


Fig. 22-143H. Chopper-drive circuit.

negative, depending on the polarity of the input signal) is coupled through capacitor C5, amplified and inverted in Q10, Q11, Q12, and then coupled through C2 to the collector of Q13. At this time Q13 is off, so the voltage step is integrated (filtered) by R15 and C6 and applied to the input of the dc amplifier. When Q16 goes from on to off, any voltage change at the drain element is amplified and appears inverted at the collector of Q13. However, at this time Q13 is turned on, and, therefore, the signal is grounded.

Transistors Q7 and Q8 (Fig. 22-143E) constitute a differential amplifier. The voltage at the base of Q8 follows the voltage at the base of Q7. Emitter followers Q14 and Q17 conduct (Q14 conducts for positive input signals; Q17 for those of negative polarity) and the voltage step is fed back to the source element of Q16. The step is in phase with the input signal; therefore, the voltage across the chopper approaches zero. Diodes CR18 and CR19 compose a clamping circuit for positive-polarity input signals. Resistor R95 and diode CR20 form a clamp for negative-polarity input signals.

The polarity indicators consist of lamps DS1 and DS2. When ionized, DS1 presents a horizontal bar of light at the readout panel, and DS2 presents a vertical bar. The two lamps function together to produce a positive-polarity image, or DS1 on and DS2 off to indicate a negative polarity. When a voltage range is selected, the cathodes of DS1 and DS2 are grounded through switch S1D. The anodes are connected through resistors R1 and R86 to plus 200 volts. In addition the anode of DS2 is connected to the collector of Q3 in the dc

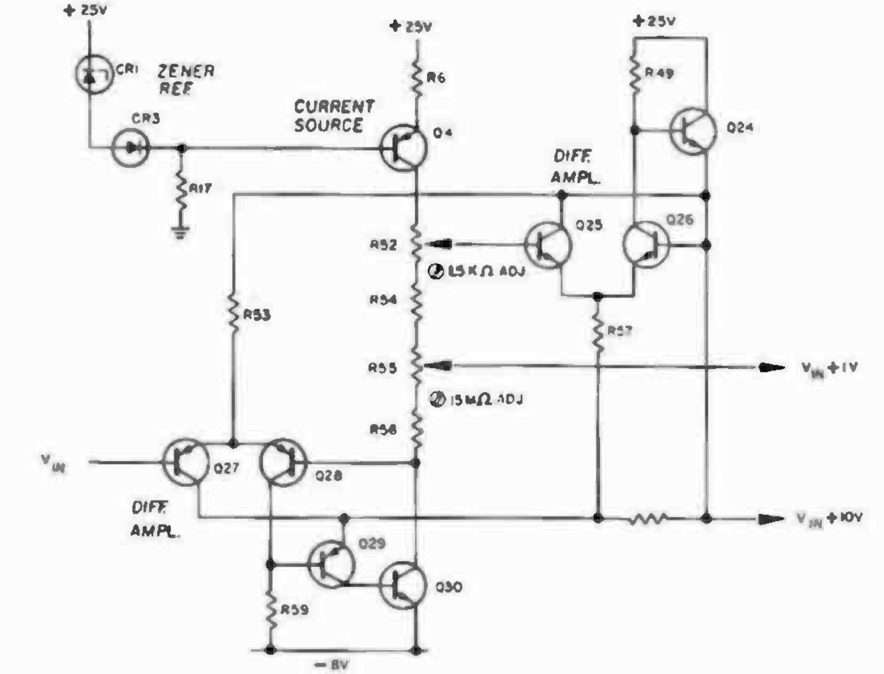


Fig. 22-143I. Ohms-converter circuit.

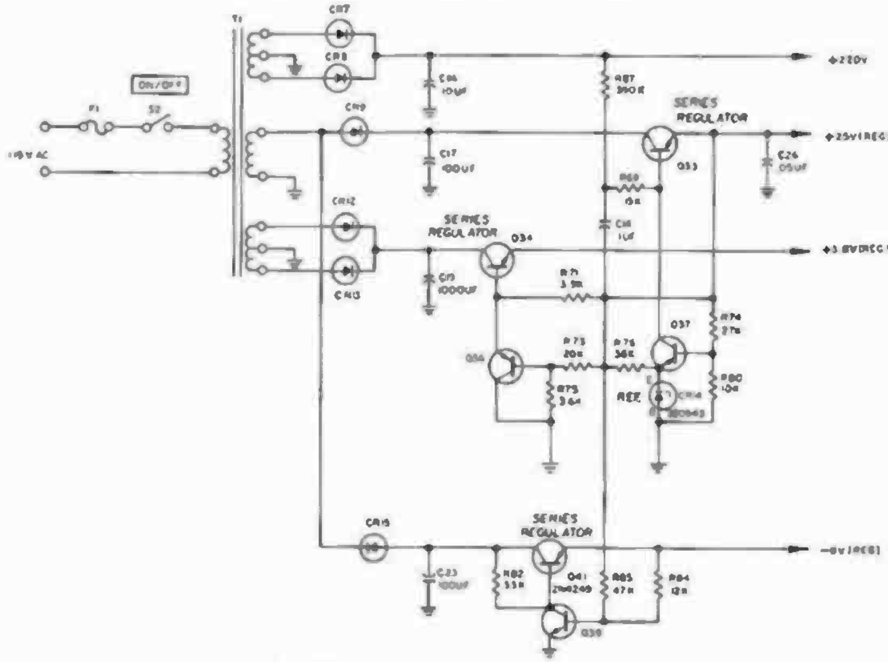


Fig. 22-143J. Power-supply circuit.

amplifier. For dc voltage inputs of negative polarity, Q3 is turned on and DS2 held off. Thus, DS1 remains ionized and a negative indication results. For positive indication, Q3 is turned off by solid-state switch Q2. The voltage at the anode of DS2 rises and allows the lamp to light, indicating the vertical bar portion of the plus indication.

The recharging of integrating capacitor C10 is provided by transistor Q9 (Fig. 22-143D). This current generator is switched off when the zero detector is activated, and held off until a full-scale pulse is emitted. Reference for the current source is established by zener diode CR1. When Q9 is off, current through CR1 is drained through CR2 to ground. With Q9 in the on position, the 6.8-volt zener diode across CR1 breaks down and fixes the voltage at the base of Q9. Transistor Q9 then conducts to produce a stable reference current. Potentiometer R5 provides a means of adjusting the current amplitude.

Note that one side of integrating capacitor C10 is connected to a voltage divider (R26 and R39) that fixes the voltage at a value of plus 12 volts. The voltage on the opposite side of C10 is discharged during the first part of a measurement. During the second portion of the cycle, it is recharged by a current equal to the reference current minus the input current. When the voltage returns to plus 12 volts(plus a small off-set voltage), the level detector is triggered.

The level detector consists of transistors Q18 and Q19 connected in a Darlington configuration. When the voltage across the integrating capacitor reaches 0 volts (12.5 volts each side) plus a small off-set voltage, the base-emitter junctions of Q18 and Q19 become forward biased. The voltage at the collector of Q19 drops, and the flip-flop steering transistor Q43 is turned on.

Flip-flop transistors Q5 and Q6 control the action of the current-reference switch and the strobe generator. Each time the counter reaches full scale, the last counting unit emits a positive spike (full-scale pulse). This pulse triggers the flip-flop via steering diode CR6. The positive pulse turns Q5 off, and Q6 is turned on through regeneration coupling. With Q5 off, current-reference switch Q9 conducts, and current from the reference source flows to the integrating capacitor. When C10 is recharged, the zero detector is activated and produces a negative-going output pulse which turns on switch Q43. This action turns Q6 off. With Q6 off, Q5 turns on through regenerative coupling. With Q5 on, the voltage at its collector is high and the positive spike turns current-reference switch Q9 off and triggers the strobe one-shot multivibrator.

The strobe multivibrator consists of transistor Q20 and RC network R40 to R42 and C11. At time t_f (Fig. 22-143C) in the measurement cycle, Q5 in the flip-flop is turned on and produces a positive pulse output. This pulse is coupled through C11 to turn Q20 off. Values of C11 and R40 to R42 are chosen to hold Q20 in the off position for 5 μs. Hence the strobe pulse (A) is a pulse of minus 0.5 volt of 5-μs duration. Strobe pulse A turns off the clock temporarily and B gates the buffer storage units, displaying the counting units.

The time base (QA12) is a 24,000-Hz oscillator, consisting of a Fairchild UL 914 dual two-input gate whose outputs have been cross connected to form a bistable oscillator. Resistors R79 and R85 and capacitors C20 and C22 are chosen to produce an output frequency of 24,000 Hz.

The clock output drives three cascaded decade counters (CUL 958 elements) and one binary counter UL 926. Each UL 958 element consists of four binary triggered flip-flops modified by feedback loop to count 1, 2, 4, 8 in binary-coded decimal. A block diagram of this micrologic component is given in Fig. 22-143F. The BCD count (Fig. 22-143C) produced in the decade counters is stored in CUL 959 buffer storage elements until a strobe pulse occurs. Strobe pulses trigger the storage elements, permitting each to transfer its BCD data to a CUL 960 decimal decoder/driver.

Buffer storage units CUL 959 consist of four gated latch circuits and a common gate driver. The input characteristics of the latch circuitry match the CUL 958 decade counter output characteristics. The buffer stage permits the state of the counter outputs to be sampled and held for an indefinite time. Each time a strobe pulse occurs, the CUL 958 counter outputs are transferred into the buffer stages.

The decimal decoder/driver units accept the 1, 2, 4, 8 BCD output from their respective storage elements and transform the BCD count into ten mutually exclusive outputs which directly control the ionizing potentials of gas-filled cold-cathode display tubes V1, V2, and V3. Ten individual lines between each driver-tube pair provide for a one-digit display of any number from zero to nine.

When the count exceeds 999, the eight output from the last CUL 958 element is high and drives Q35. The collector is connected to pin 3 of QA7.

With Q35 on, the voltage at its collector is low and QA7 is triggered. Element QA7 is a flip-flop and controls bistable multivibrator Q31 and Q32, which operates DS3. Transistors Q31 and Q32 normally are held off and CR10 and CR11 are back biased. During the strobe pulse, the voltage at the emitters of Q31 and Q32 drops, thus permitting the output state of Q47 to be sampled. If pin 7 is high and pin 9 is low, CR11 becomes forward biased; Q31 is turned on, and Q32 off. With Q32 in the on state, Q40 conducts, creating a voltage drop across DS3 and the number 1 appears as the most significant digit in the readout.

The chopper (Fig. 22-143G) is a F1100 MOSFET. The gate must be biased to approximately 4 volts negative with respect to its source element before the device will turn on. The chopper drive circuit (Fig. 22-143H) produces a square-wave signal that alternates between plus 1.2 and minus 6 volts. The positive 1.2-volt signal is necessary because of the voltage drop across diode CR2 and transistor Q13.

Element QA8 (Fig. 22-143D) is the third CUL 958 element in the series decade counters. The first two counter elements divide the 24,000-Hz clock signal by 100. Flip-flop circuit QA8 divides the 240-Hz input by two. The resulting 120-Hz signal at pin 3 of QA8 triggers driver transistor Q38. The square waveform appearing at the collector of Q38 alternates between plus 1.2 and minus 6 volts. This output signal drives chopper transistor Q16 and solid-state switch Q13.

The ohms converter (Fig. 22-143I) is switched into the circuitry when an ohms range is selected by the function switch. This circuit consists of a bootstrap amplifier that produces two outputs. One output is used for a 15-megohm range only. This output adds 1 volt to the voltage at the amplifier input. The second output is used for all other ranges except 15 megohms. This second output adds 10 volts to the amplifier input. The appropriate output voltage is connected to the low end of the attenuator for a given range. The attenuator network programs a test current through the unknown resistance. Test currents are: 0.1 μA for the 15-megohm range, 1 μA for the 1.5-megohm range, 0.01 mA for the 150,000-ohms range, 0.1 mA for the 15,000-ohm range, and 1 mA for 1500-ohm range. The voltage drop across the unknown resistance is measured with the dc voltage-measuring portion of the instrument.

Transistor Q4 is a current source for the ohms converter. The zener breakdown potential of CR1 provides a reference for the current source. Current from Q4 flows through R52, R54, R55, and Q30. Elements Q27 and Q28 compose a differential amplifier. The base voltage of Q27 is always directly proportional to the value of the unknown resistance. An increase in V_{in} causes Q27 to conduct less and Q28 to conduct more, thus causing Q25 to conduct harder because of the increased voltage appearing at R52. The base voltage of Q26 rises to follow the voltage at the base of Q25 and hence to remain 10 volts above the voltage at the input to the ohms converter (Q27). Therefore, the output at the base of Q26 is always equal to the voltage at the base of Q27 plus 10 volts. The output for the 15-megohm range appears on the arm of R55. Because of the bootstrap action, the output voltage is always 1 volt above the voltage that is present at the base of Q27.

A solid-state power supply (Fig. 22-143J) provides voltage levels of plus 220 volts, plus 25 volts, plus 3.8 volts, and minus 8 volts. The 220 volts supplies the anode potential for the display tubes and lamps. The plus 25-volt and minus 8-volt supplies are used with the control logic and the analog-to-digital (A to D) converter. The plus 3.8-volt supply is used in connection with the micrologic elements.

Reference for the plus 25-volt, 3.8-volt, and minus 8-volt outputs is established by the zener breakdown of CR14, a transistor that has its collector lead open circuited and its base and emitter terminals connected as shown in detail in Fig. 22-143H. The emitter-to-base breakdown potential of CR14 establishes reference voltage for the emitter of Q37. The base lead is connected to the junction of sensing resistors R74 and R80. If the output voltage decreases because of a load change, the decrease causes Q37 to conduct less. The voltage at the base of series regulator Q33 rises, and Q33 conducts harder to compensate for the drop in output voltage. Capacitor C26 is a filter capacitor for the plus 25-volt supply.

Reference voltage for the minus 8-volt supply is taken from the plus 25-volt supply. Sensing resistors R84 and R85 are connected between the plus 25-volt output and the minus 8-volt terminals. Assuming that the plus 25-volt output is absolutely stable, the slightest variation in the minus 8-volt output will be sensed across R84 and R85. This variation is sensed at the base of Q39. For a drop in output voltage (less negative) the base voltage of Q39 will rise and Q39 conduct less. The voltage at the base of Q41 drops and Q41 conducts harder to accomplish regulation.

Diode CR15 rectifies the transformer output and C23 filters the rectified input signal. Resistors R73 and R75 form a voltage divider that fixes the voltage at the base of Q36. Q36 acts as a constant-current source and fixes the voltage at the base of Q34. When the voltage at the emitter of Q34 tends to drop, Q34 conducts harder and compensates for the voltage drop. The plus 3.8-volt output is also dependent on the plus 25-volt supply.

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Section 23

Audio-Frequency Measurements

Lord Kelvin once said, "I often say that when you can measure what you are speaking about, and can express it in numbers, you know something about it; but when you cannot measure it, when you cannot express it in numbers, your knowledge is of meager and unsatisfactory kind." A measurement, however, is only as good as the equipment and method used in extracting this knowledge. The USASI (ASA) has specified the characteristics of certain test instruments, and set forth standards to be met by equipment under test, but it often fails to specify the manner in which the measurement is to be made. The generally accepted methods of such measurements are delineated herein. Although a given device has been selected to facilitate discussion of certain types of measurements, the information is usually applicable to other such instruments and similar measurements. Methods of calculating and plotting the results, correction factors to be observed where indicated, and proper load terminations are included in the discussions.

23.1 How are the frequency characteristics of an audio amplifier measured? —Four circuits commonly used for the measurement of amplifier characteristics are shown in Figs. 23-1A to D.

The simplest form of measuring the gain or frequency response of an amplifier is to connect an audio oscillator to the input of the amplifier through a suitable attenuator network and send frequencies of constant amplitude into the amplifier, measuring the current at the input with a sensitive current-indicating meter (Fig. 23-1A). A second meter is used in the output or load circuit. The ratio of the input to output currents is then used to compute the gain of the amplifier at the frequencies of interest.

This method of measurement is not without its difficulties, however, due primarily to the fact that most audio-frequency, current-indicating instruments are generally insensitive to very small values of current. Therefore, it is customary to employ instruments indicating voltage rather than current. In Fig. 23-1B is shown a similar circuit using a single vacuum-tube voltmeter for measuring the voltage applied to the input and the voltage developed by the amplifier at its output. The ratio of the input voltage to the output voltage is then used to compute the frequency characteristics.

The only disadvantage of this circuit is that when the meter is switched from input to output, the multiplier of the meter must be changed to a higher scale to prevent damage to the meter movement.

One advantage of the preceding method of measurement is that frequency errors in the meter are cancelled

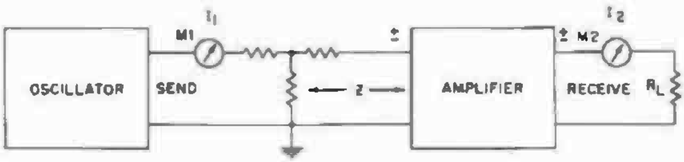


Fig. 23-1A. Measuring the frequency response of an amplifier by using two sensitive milliammeters.

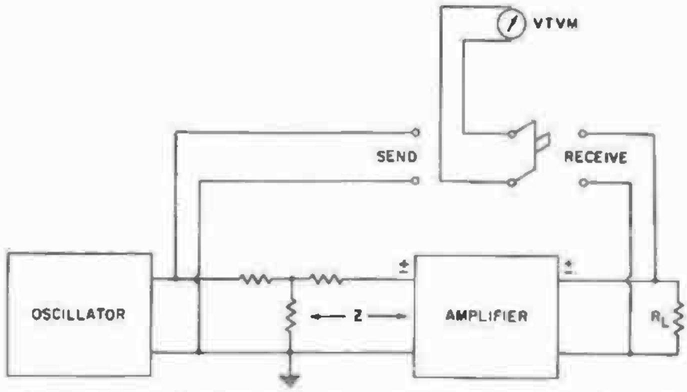


Fig. 23-1B. Measuring the frequency response of an amplifier by using a single vacuum-tube voltmeter.

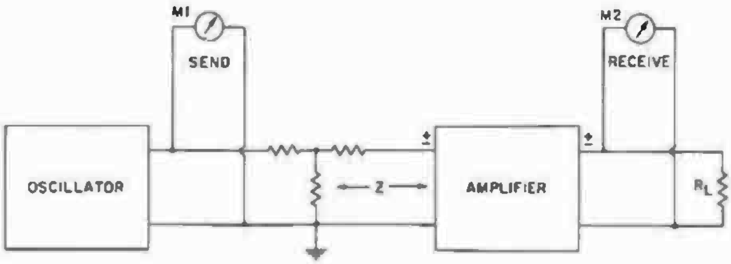


Fig. 23-1C. Using two voltmeters to measure the frequency response of an amplifier.

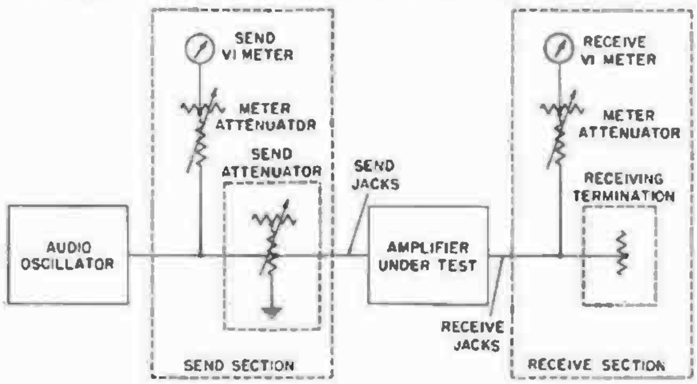


Fig. 23-1D. Basic circuit of a two-meter gain set using an unbalanced configuration.

out because the same meter is used to measure both the input and output circuits. (See Question 23.25.)

The meter at the input is always connected at the oscillator output ahead of the network to provide a high output voltage to facilitate reading of the meter, because the voltage at the input of the average amplifier will be rather low. As an example, if the voltage at the oscillator output is 1 volt and the attenuator has a loss of 40 dB, the voltage at the input of the amplifier is 0.01 volt, or a ratio of 100/1. Reading the input voltage at the output of the oscillator increases the signal-to-noise ratio of the readings.

Fig. 23-1C shows a similar circuit but with two meters, one connected at the input and the other across the load resistance. In this instance the input meter is held to a given voltage for all frequencies and the difference is read on the output meter. The disadvantage of this circuit is that the two meters must have similar frequency characteristics over the frequency range to be measured. If the frequency characteristics of the meters differ, the true fre-

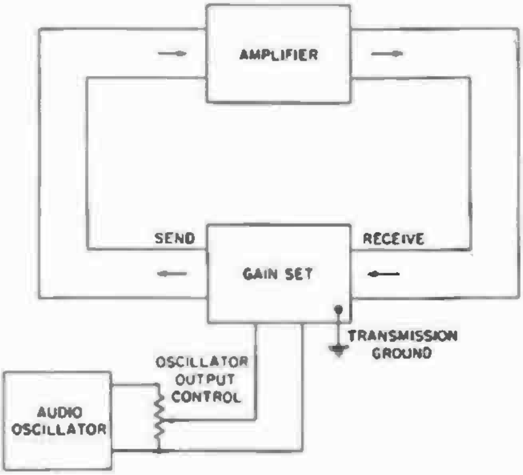


Fig. 23-1E. Block diagram for a typical gain set (also called a transmission set).

quency response will not be obtained. (See Question 23.25.)

The fourth method of measuring the frequency response of an amplifier is by the use of a gain set, as described in Questions 22.122 to 22.124. The basic block diagram of a gain set is shown in Fig. 23-1D. Here, two meters are again used; however, in this instance, they are standard VU meters as described in Question 10.3.

The configuration of the attenuator network used at the input of the amplifier will depend on the configuration of the amplifier input. The network may be balanced or unbalanced, variable or fixed in loss; however, it must supply an impedance match between the oscillator output and the input of the amplifier.

The insertion of an attenuator network between the output of the oscillator and the input of the amplifier serves a triple purpose: (1) It attenuates the signal voltage to a level suitable for the input of the amplifier; (2) it provides an impedance match between the oscillator and the amplifier; and (3) it increases the signal-to-noise ratio of the oscillator signal by attenuating the internal noise and hum of the oscillator, which is constant.

The basic circuits shown in this question may also be used for measurement of overload, power, and distortion. To simplify the block diagrams of the various measurements to be discussed in this section, the basic block diagram of a gain set will be used, as it presupposes that an oscillator, send attenuators, input and output meters, and load termination are available and that all impedance matches are satisfied.

Fig. 23-1E is a basic block diagram of a typical gain set and is the one which will be used for illustrating the various methods of measurement.

23.2 ***At what level output should a frequency-response measurement be made on an amplifier?***—If the normal characteristics of the amplifier are reasonably flat, measurements are made 8 to 10 dB below the maximum output level. If the characteristics are not flat, the procedure as outlined in Question 23.13 is followed.

23.3 ***Describe the procedure for making a frequency-response measurement using a gain set.*** — To illustrate the procedure for making a gain-versus-frequency measurement using a gain set, it will be assumed that the measurement is to be made on a simple amplifier having symmetrical (ungrounded) 600-ohm input and output impedances. The gain set to be used will be similar to that described in Question 22.122.

The amplifier input, being symmetrical, may be directly connected to the gain-set send terminals, as shown in Fig. 23-15A. To prevent possible damage to the gain set and to the equipment under measurement, the maximum loss of the gain-set attenuators is inserted before the measurement is started.

Next, the send and receive impedance switches of the gain set are set to 600 ohms and the amplifier chassis is grounded to the transmission ground. (See Question 23.24.) The amplifier may now be turned on and permitted to warm up to bring the tubes to their normal operating temperatures before proceeding with the measurements.

If the audio oscillator is of the beat-frequency type (see Question 22.51), it will also require a warm-up period to prevent it from drifting off frequency during the measurement. If the oscillator is of the Wien-bridge type (see Question 22.50, it will require only a few minutes of warm-up time.

The send VU meter attenuator on the gain set is set to its plus 4-dBm position. The receive VU meter attenuator is set to its highest position for protection until the actual measurement is started. Set the oscillator to 1000 Hz and adjust its output for an indication of plus 4 dBm on the send VU meter. Turn the attenuator of the receive VU meter to plus 4 dBm and start the measurement by removing loss from the gain-set attenuators in small steps until a reading of plus 4 dBm is obtained on the receive VU meter. With the send and receive meters both reading plus 4 dBm, the gain of the amplifier is equal to the loss in the gain-set attenuators.

If the receive meter indicates a fractional part of a decibel below the zero calibration mark, remove 1 dB from the attenuators and read the fractional part on the plus side of the zero calibration mark. The gain of the amplifier is now read as the total loss of the attenuators plus the fractional part on the receive VU meter.

To illustrate: If the gain-set attenuator loss adds up to 43 dB and the receive meter indicates 0.5 dB above the zero mark, the gain of the amplifier is 43.5 dB.

When a fractional part of a dB on the plus side of the zero mark is read, it is added to the loss of the attenuators, and when read on the minus side of the zero mark, it is subtracted from the attenuator loss.

In some gain sets, the send impedance switch induces a resistive network which causes an additional loss. This loss must be added to the loss of the variable attenuators to arrive at the true gain of the amplifier. The loss induced by the input network is indicated on the dial of the selector switch. If an external attenuator is being used, its loss must also be included in the final summing up.

The same procedure is used for measuring each frequency of interest. It is important that the send signal be maintained at the same level for all frequencies.

The basic principles of gain-set operation may be summed up as follows: A known signal level is sent into the input of the gain-set attenuator network and sufficient loss is induced by the gain-set attenuators to obtain a level at the receive VU meter equal to that of the send VU meter (assuming all impedance matches have been satisfied). Therefore, under these conditions *and these conditions alone* will the gain of the amplifier be equal to the loss of the gain-set attenuators in decibels. (See Question 23.4.)

23.4 ***How are the dials and meters of a gain set read?***—To illustrate the procedure for reading the attenuator loss and meter indications of a gain set, a typical two-meter, two-dial gain set is shown in Fig. 23-4. Suppose that the dials indicate 40-dB loss on the first dial and 3-dB loss on the second dial. This is a total loss of 43 dB. The receive VU meter is shown indicating plus 0.5 dB above the zero calibration mark. How is this fractional part added to the dial readings?

Referring to the illustration in Fig. 23-4, an amplifier under these conditions would have a gain of 43.5 dB. Because the attenuators read in even steps of 1 dB and the amplifier has a gain of 43.5 dB, the fractional part of the gain will be indicated on the receive meter.

If the fractional part appears on the minus side of the receive-meter zero, 1 dB is removed from the attenuators and the fractional part read on the plus side of the zero mark as shown. If the fractional part is read on the minus side of zero, it is subtracted from the loss of the attenuators (44 minus 0.5 equals 43.5 dB). In either instance the total gain is the same; however, it is more convenient to read fractional parts on the plus side when adding up the dials.

To illustrate how the attenuators function to indicate gain, the signal from the oscillator is reduced by the attenuators by exactly the same amount as the amplifier has amplified it, except for the fractional parts read on the receive VU meter. This is illustrated by the heavy line in Fig. 23-4. In reading gain, the loss of the attenuators is con-

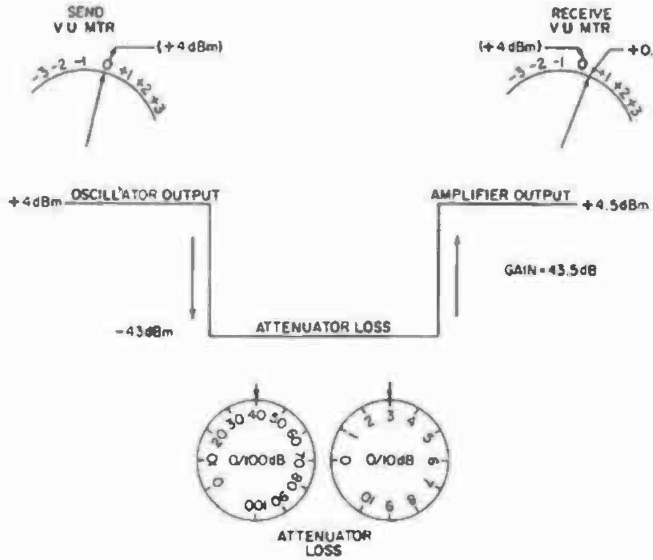


Fig. 23-4. Voltage reduction by the attenuators in a gain set for an amplifier of 43.5 dB gain.

sidered to be a positive value. If the gain set includes an attenuator calibrated in 0.10-dB steps, it is adjusted to bring the receive VU meter to the zero mark. The reading of the fractional attenuator dial is then added to that of the other dials.

The same procedure is followed for the measurement of each frequency of interest. The second (oscillator) signal level is maintained at the same value for all frequencies. A gain set may also be used to read insertion loss and power-output levels and to indicate the overload point of an amplifier. These subjects are discussed elsewhere in this section.

23.5 ***If a gain set includes a 0.10-dB attenuator, how is it read?***—It is adjusted to bring the receive meter to zero. Its loss is then added to the other attenuators.

23.6 ***How is a frequency-response measurement tabulated when using a gain set?***—If a large number of frequency runs are to be made daily, a tabulation sheet similar to that shown in Fig. 23-6 is prepared. The frequency column starts at 30 Hz and progresses to 15,000 Hz. Other frequencies may be used. However, the ones shown will generally cover most measurements involving recording and reproducing equipment. Space is provided at the bottom of the sheet for entering the noise, distortion, and power output of the various devices measured. The values of loss, as read from the gain-set attenuators, are entered in the *Run* column. Later these values will be used to convert the measurements to a reference frequency which is entered in the *Delta* (Δ) column.

A typical gain run has been entered in the Run column to illustrate how the settings of the gain-set attenuators are tabulated. The values in the Delta column have been converted to a reference frequency of 1000 Hz. It is customary to enter on the reverse side of the sheet a block diagram of how the measurement was made, indicating the various pieces of test equipment, operating levels, impedance values, terminations, and any other pertinent data that would assist in duplicating the setup at a later date. To illustrate how the gain run shown is converted to a reference frequency, assume that 1000 Hz is to be used as the reference frequency. The attenuator loss at 1000 Hz is 40 dB and remains at this value up to 6000 Hz. At 7000 Hz the attenuator loss is 39.9 dB, and at 10,000 Hz it is 37 dB. Going downward in frequency, the attenuator loss remains 40 dB until 60 Hz is reached, at which point the loss increases to 40.5 dB; at 30 Hz, it increases to 43 dB.

Inspection of the attenuator loss indicates that at frequencies above 6000 Hz, the gain of the amplifier must fall

TRANSMISSION MEASUREMENT DATA

Date ____________ Equipment ____________ Engineer ____________

Frequency	*Run*	Δ	*Run*	Δ	*Remarks*
30 Hz	43.0	+3.0			
40 Hz	41.8	+1.8			
50 Hz	41.0	+1.0			
60 Hz	40.5	+0.5			
80 Hz	40.0	0.0			
100 Hz	40.0	0.0			
150 Hz	40.0	0.0			
200 Hz	40.0	0.0			
300 Hz	40.0	0.0			
400 Hz	40.0	0.0			
800 Hz	40.0	0.0			
1 kHz	40.0	0.0			
1.5 kHz	40.0	0.0			
2 kHz	40.0	0.0			
2.5 kHz	40.0	0.0			
3 kHz	40.0	0.0			
4 kHz	40.0	0.0			
5 kHz	40.0	0.0			
6 kHz	40.0	0.0			
7 kHz	39.9	−0.1			
8 kHz	39.6	−0.4			
9 kHz	38.5	−1.5			
10 kHz	37.0	−3.0			
11 kHz					
12 kHz					
13 kHz					
14 kHz					
15 kHz					
NOISE	−60 dBm.				
DISTORTION	0.5% at +4 dBm.				
MAX. POWER	+10.0 dBm.				

Note: Attach a complete block diagram of the test setup used. Show all impedance values, operating levels, gain, and equipment type numbers.

Fig. 23-6. Typical transmission-run data sheet.

off, because less attenuator loss is required to maintain a constant level at the receive meter. At the low frequencies, the amplifier gain must rise, because more attenuation is required to maintain a constant output level.

Keeping the above facts in mind, the attenuator loss may now be converted to a reference frequency to show the rise and fall of the frequency characteristic above and below the reference frequency.

The attenuator loss, being the same for frequencies between 80 and 6000 Hz, indicates the amplifier frequency response over this range is flat. This fact is entered in the Delta column as 0.0 dB. Moving upward in frequency to 7000 Hz, the attenuator loss at this frequency is subtracted from the attenuator loss at 1000 Hz, and the difference is entered in Delta column as minus 0.10 dB. At 10,000 Hz the loss is 37 dB, or 3 dB less than the value at 1000 Hz. Therefore, the amplifier has a loss of 3 dB at 10,000 Hz. This is noted as minus 3 dB in the Delta column. The fractional decibel values are read on the receive VU meter, as described in Question 23.3, if the gain set does not include an attenuator calibrated in fractional parts of a decibel.

At the lower frequencies the reverse takes place. At 30 Hz, 43 dB of loss is required; therefore, the amplifier has a rise of 3 dB at 30 Hz. The rule to remember when converting attenuator loss to a reference frequency is: If the attenuator loss is less than that at the reference frequency, the device under measurement shows a *loss* in frequency response, and if the loss is greater than that at the reference frequency, the device shows a *rise* in frequency response.

A typical plot using a reference frequency of 1000 Hz is shown in Fig. 23-7A. In some instances it may be desirable to plot the frequency characteristic in gain rather than by referring to a reference frequency. In this latter type of plot, the actual values of loss are entered on the graph (which represents gain) as shown in Fig. 23-7B.

23.7 State the general rules pertaining to the measurement and plotting of amplifier characteristics.—In accordance with IHF Standard A-200-1966, amplifiers are to be tested at a line voltage of 120 volts plus or minus 1 percent, at 60 Hz, or within 2 percent of the lowest supply frequency given by the manufacturer. The waveform of the line voltage is to be sinusoidal, with a THD of less than 2 percent. The ac line input is to be connected to the power transformer primary tap that results in minimum gain on the highest gain input, and the connection is not to be changed during any other test performed on the amplifier. One side of the ac line is to be at ground potential.

Before the tests are started, the amplifier is to be preconditioned by operating all channels at one-tenth the highest reference power into its stated

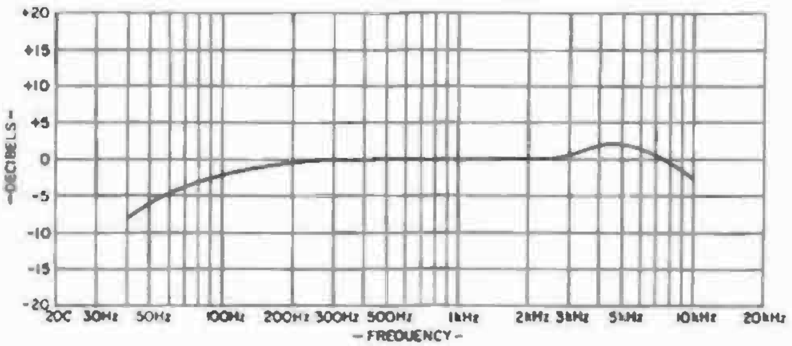


Fig. 23-7A. Frequency response of typical amplifier plotted in decibels with reference to 1000 Hz.

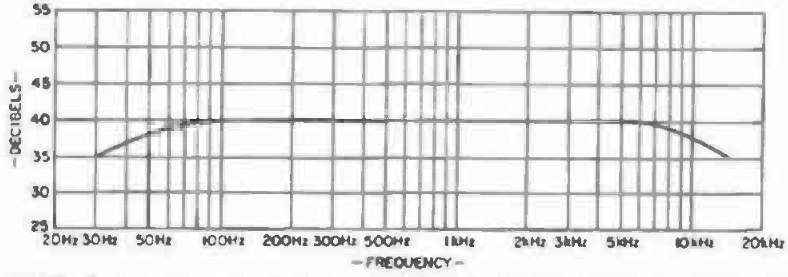


Fig. 23-7B. Frequency response of an amplifier plotted as frequency versus actual gain in decibels with reference to 1000 Hz.

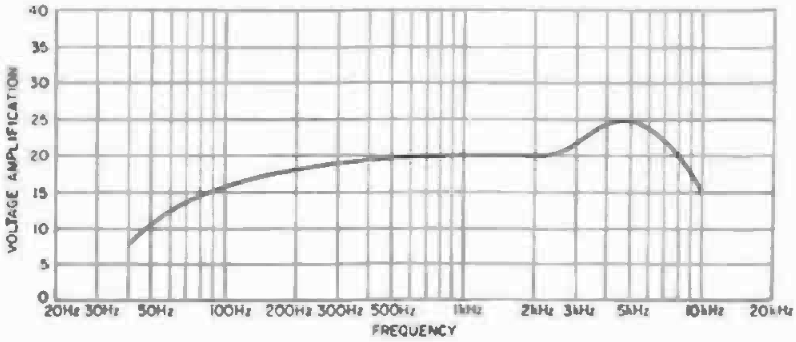


Fig. 23-7C. Frequency response of the amplifier considered in Fig. 23-7A plotted in voltage gain, with reference to 1000 Hz.

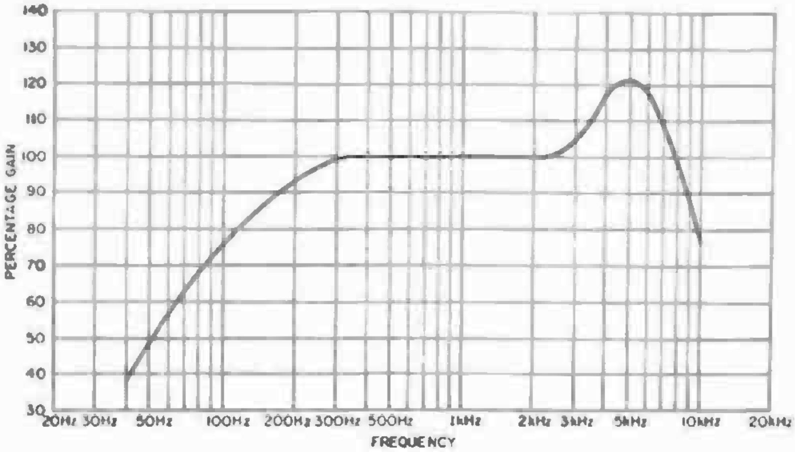


Fig. 23-7D. Frequency response of amplifier considered in Fig. 23-7A plotted in percent gain of 1000 Hz.

load resistance for at least 1 hour at an ambient temperature of 25°C plus or minus 3°C in still air, in its normal operating position, with all shields and bottom plates in position.

Gain or other controls affecting the output level are to be preset to their maximum gain positions. Tone and any other controls affecting the frequency response are to be set for the most uniform response. Stereophonic balance controls are to be set to the normal position as specified by the manufacturer. Automatic control circuits actuated by the input signal are to be disconnected.

The harmonic distortion of the input signal (oscillator) shall not be greater than 20 percent of the measured distortion of the amplifier to be tested. This means, for an amplifier of 1-percent total harmonic distortion, the THD of the oscillator can not exceed 0.20 percent. Test frequencies are to be plus or minus 2 percent of the specified value.

The output of the amplifier is terminated in a resistive load with not more than a 10-percent reactive component at any frequency up to five times the highest test frequency and capable of dissipating the full load of the amplifier while maintaining its resistive value within 1 percent of the rated value. Unless otherwise specified, the amplifier is to be terminated in an 8-ohm resistive load. Preamplifier outputs are to be terminated in 0.1 megohm shunted by a 1000-pF (0.001-μF) capacitor, plus or minus 5 percent unless otherwise specified by the manufacturer.

To show the complete characteristics of an amplifier, several different measurements are required. Among these are frequency response, continuous sine-wave power output, harmonic and intermodulation distortion, music power

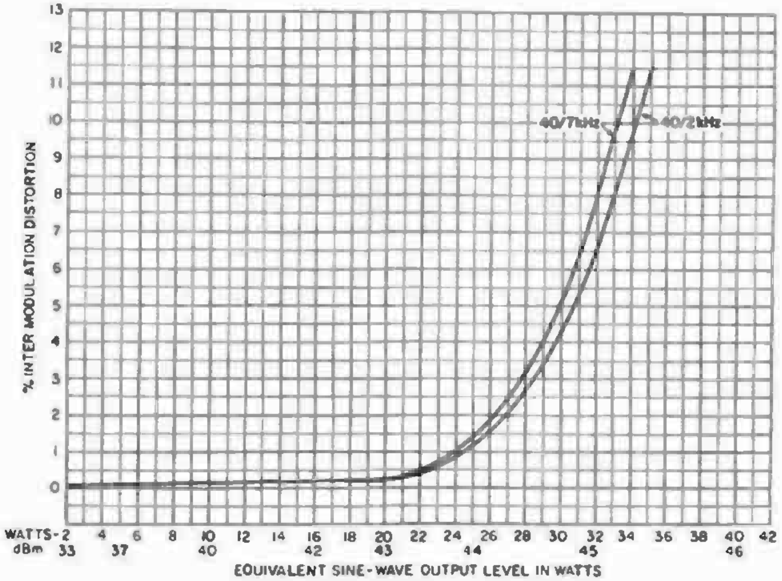


Fig. 23-7E. Intermodulation distortion curve of a 20-watt amplifier.

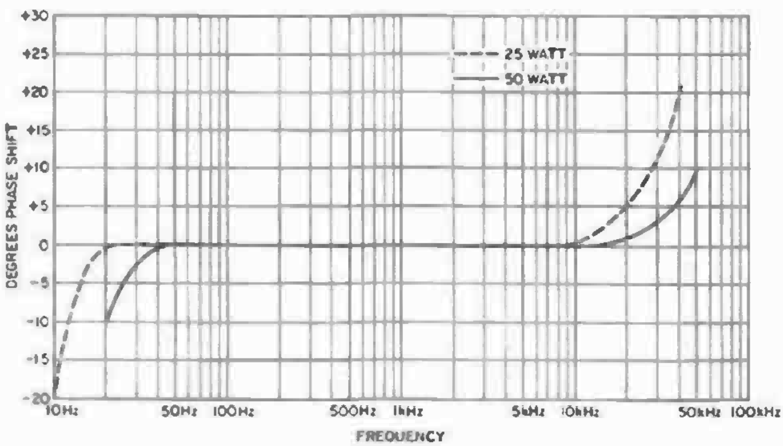


Fig. 23-7F. Phase-shift characteristics of two high-quality amplifiers.

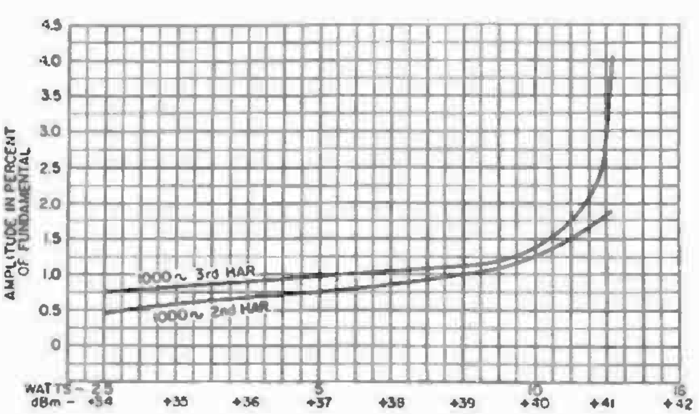


Fig. 23-7G. Harmonic amplitude versus power output.

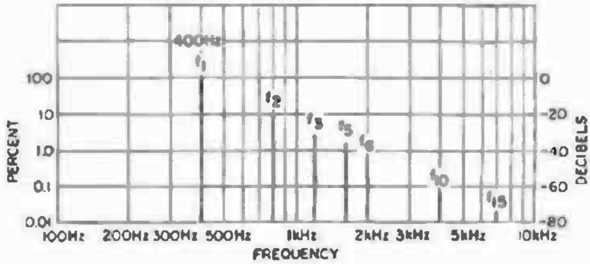


Fig. 23-7H. Amplitudes of a fundamental frequency f_1, plotted in decibels below the amplitude of f_1, using a wave analyzer.

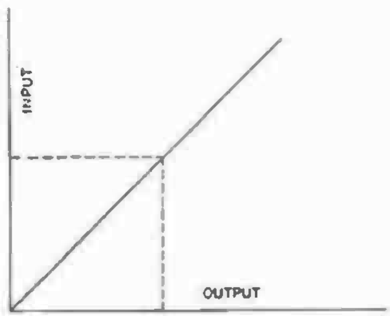


Fig. 23-7I. Linearity characteristics of an ideal amplifier.

output, transient distortion, linearity, hum and noise, internal output impedance, damping, phase shift, sensitivity, and stability. If the amplifier is for stereo, additional tests for separation (crosstalk) and tracking error are to be included. These measurements are made with the input signal applied to both sides simultaneously. Plotting of these characteristics are given in Figs. 23-7A through K.

The frequency response of an amplifier may be plotted several different ways. Fig. 23-7A shows the most common method, using semilog graph paper. Frequency response may also be plotted as in Fig. 23-7B, showing the actual gain in decibels at each frequency. The plot in Fig. 23-7C is that of the amplifier in Fig. 23-7A, except that it is plotted voltage versus frequency, and again in Fig. 23-7D as percent gain versus frequency.

Harmonic or intermodulation distortion is plotted as given in Fig. 23-7E, and with phase distortion it is plotted as shown in Fig. 23-7F. Figs. 23-7G and H illustrate two methods of plotting the amplitude of harmonics, using a harmonic wave analyzer. Linearity characteristics are plotted in Fig. 23-7I. The power bandwidth of an amplifier is determined by plotting a power curve (Fig. 23-7J), and spotting the points where the response is down 3 dB. The plot in Fig. 23-7K, using 3 × 5 log-log paper, is that given in IHF Standard A-201-1966. (See Questions 23.207 and 23.208.)

Step-generator measurements are made by photographing the input and output signals simultaneously on a dual-trace oscilloscope. The amplifier departs from a linear device when the output step pattern departs from the waveshape of the input signal. The internal output impedance is plotted as shown in Fig. 23-138B.

23.8 In what position is the send meter generally set for a normal gain run on an amplifier?—The send-meter attenuator is, as a rule, set to a plus 4-dBm position. The same signal level is then maintained for all frequencies. The relationship of the zero calibration

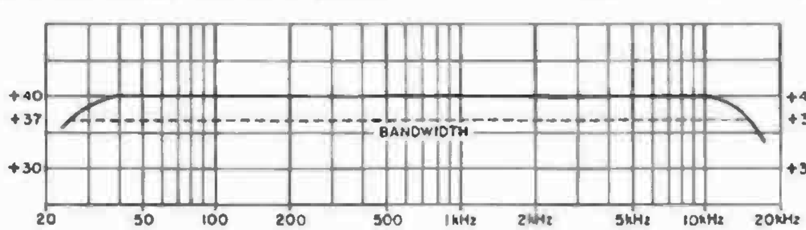


Fig. 23-7J. Power bandwidth curve for a typical 10-watt amplifier showing the half-power points versus frequency.

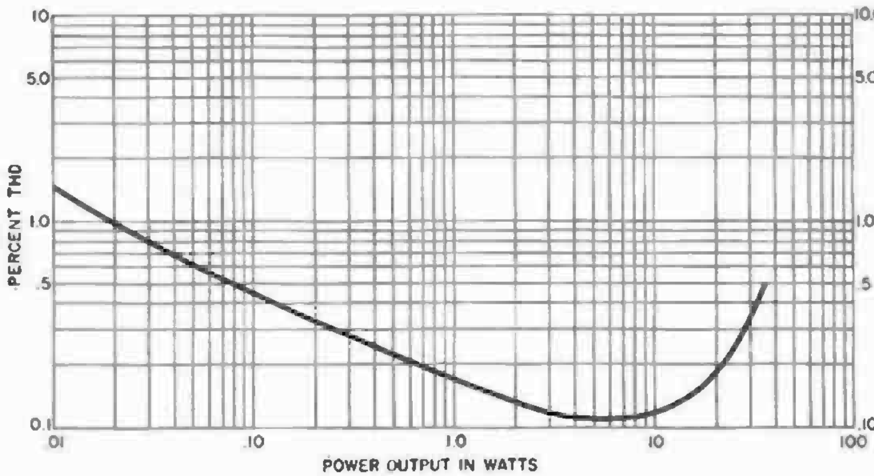


Fig. 23-7K. Distortion characteristics for a typical amplifier plotted using log-log graph paper. Plotting the distortion in this manner permits the distortion at the lower levels to be more easily observed.

to the actual signal level indicated by a VU meter is fully described in Questions 10.17 and 10.24.

23.9 ***Is it permissible to set the receive meter at a higher level than a plus 4 dBm or zero?***—Yes, the receiving meter may be set to any level. However, for normal frequency-response measurements, it is generally set to the *plus 4-dBm level*. If so desired, the measurement may be made at a higher output level such as a plus 40 dBm (10 watts), in which case, it is made in exactly the same manner as if it were run at a lower level.

Precautions must be taken that the input and output leads to the amplifier are properly separated to prevent feedback and that the amplifier chassis is grounded to the transmission-ground system. If this is not observed, oscillation may take place and damage the receive meter of the gain set.

23.10 ***Can the signal level at the send VU meter be set to a lower or higher level than plus 4 dBm?***—Yes. When the meter attenuator is set to plus 4 dBm, the signal level at the input of the attenuator network of the gain set is a plus 4 dBm and then is reduced to a lower level by the attenuator network. At times it may be desirable to increase the send signal level or, in some instances, to reduce it below a plus 4 dBm.

If a bridging amplifier of low gain is being measured, it may be necessary to increase the send level by several decibels to permit a measurement to be made. If this is necessary, the increased level at the send end must be taken into consideration when computing the final gain of the amplifier. As an example, if the send meter is set for a plus 10 dBm, the receive meter is set to a plus 4 dBm, and the attenuator loss is 20 dB, the actual gain of the amplifier is 14 dB. (See Question 23.16.)

To analyze this statement, note that the send level is plus 10 dBm. The attenuator loss is 20 dB. Therefore the signal at the amplifier input has a level of minus 10 dBm. If the output meter indicates a level of plus 4 dBm, the gain of the amplifier must be 14 dB, because the input signal was raised from a minus 10 dBm to a plus 4 dBm, for a total increase of 14 dB. To further illustrate how the attenuators and meters are used, assume the send VU meter reads a minus 2 dBm, the attenuator loss is 34 dB, and the receive meter reads minus 2 dBm. What is the amplifier gain under those conditions? With a minus 2-dBm signal at the input of the gain set, the level at the amplifier input will be minus 36 dBm. To raise a minus 36-dBm signal to a minus 2-dBm level will require an amplifier gain of 34 dB. The rule for such measurements is: Whenever the send VU meter reads higher than the receive meter, the difference between the two levels is subtracted from the attenuator dial readings. If the level of the send meter is lower than the receive meter, the difference is added to the loss of the attenuator dials. If the send and receive meter levels are the same, only the loss of the attenuators, which is equal to the amplifier gain, is considered.

23.11 If the send meter is set to a level below plus 4 dBm, how is the gain of an amplifier computed?—Assume the send meter is set to minus 4 dBm (see Questions 10.17 and 10.24) and the receive meter reads a plus 4 dBm, with 20-dB loss in the gain-set attenuators. What is the gain of the amplifier? The true gain is 28 dB.

This statement may be analyzed as follows: The send signal is a minus 4 dBm. The loss of the attenuators is 20 dB. This makes the signal level at the input of the amplifier minus 24 dBm. The receive meter reads plus 4 dBm. To increase the signal level from a minus 24 dBm to a plus 4 dBm requires a gain of 28 dB. Therefore, the amplifier gain must be 28 dB. It is assumed that all impedance matches are satisfied and no correction factors are required for either meter.

23.12 What relationship does the attenuator loss in a gain set bear to the input signal of a device under measurement?—If the send VU meter is set to indicate a true 1-milliwatt reference level, or 0 dBm, the input level is the level indicated by the loss of the attenuators. As an example, if the level indicated by the send VU meter is zero dBm and 40 dB of loss is inserted in the attenuators, the signal level at the gain-set send terminals will be minus 40 dBm. However, this will only be true if the gain-set send terminals are terminated in their characteristic impedance. If the send VU meter is indicating a plus 4 dBm, then the level at the send terminals (properly terminated) for 40-dB loss would be minus 36 dBm. (See Question 10.17.)

23.13 What precautions should be taken when measuring the frequency response of an amplifier having a considerable amount of equalization? —The frequency of maximum rise should be determined by sweeping the oscillator across the frequency band and noting the frequency of maximum rise. This frequency is then used for a reference frequency.

To illustrate the procedure, suppose that an amplifier indicates a 20-dB rise at 80 Hz with respect to 1000 Hz (a typical magnetic-tape playback amplifier response). If a frequency of 1000 Hz is used for the reference frequency, the amplifier output must rise 20 dB, which, depending on the amplifier design, might cause serious overload and result in an erroneous frequency response.

If a frequency of 80 Hz is taken for the reference frequency, the frequency response will fall off as 1000 Hz is approached. Thus, the amplifier is prevented from being driven into overload. After the frequency response has been obtained in the preceding manner, it is replotted with reference to 1000 Hz.

The foregoing statement is particularly true for amplifiers used in magnetic-tape recorders where the equalization at the high and low ends may rise 15 to 22 dB with reference to 1000 Hz. (See Fig. 23-52A.)

23.14 Show the different types of input circuits used with balanced gain set.—A balanced gain set consists of a group of balanced attenuators using either a T, bridged-T, or H-type configuration. As explained in Question 22-123, the center tap of each of these configurations is grounded. Therefore, if a balanced gain set is used with an unbalanced input circuit, it is possible to short out a portion of the attenuators or a portion of the input circuit of the device under test. To prevent this, a repeat coil is connected between the send terminals of the gain set and the input terminals of the device being tested. (See Fig. 23-14A.) Three types of input circuits commonly used with

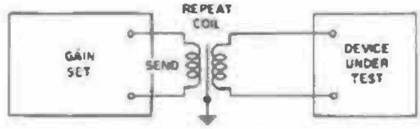


Fig. 23-14A. Repeat coil connected between the send terminals of a gain set and the input of a device under test.

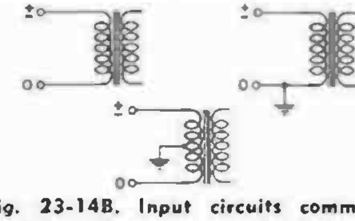
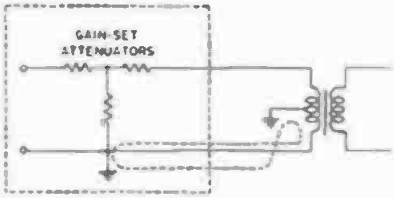
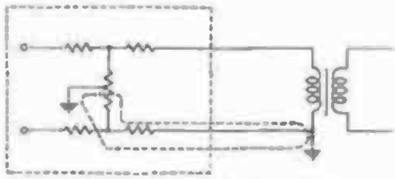


Fig. 23-14B. Input circuits commonly used in audio equipment. (a) Symmetrical input. (b) Grounded input. (c) Balanced to ground input.



(*a*) *Unbalanced to balanced input.*



(*b*) *Balanced to unbalanced input.*

Fig. 23-14C. How an improperly isolated input circuit is shorted out by the gain-set ground.

audio-frequency equipment are shown in Fig. 23-14B.

Fig. 23-14C illustrates how the input of a device being tested could be shorted out by improper isolation of the two circuits. The proper terminology and use of input circuits is discussed in Question 8.25.

In (a) to (h) of Fig. 23-14D are shown circuits suitable for use between the send terminals of a balanced gain set and the input circuit of a device with one of the three circuits illustrated in Fig. 23-14B. It will be noted that many of the circuits make use of a repeat coil for isolation. When sending into the circuits in (a) and (e) of Fig. 23-14D, where two ground connections are involved, it might be necessary to impose a repeat coil in the circuit to prevent ground noise, particularly if the physical separation of the two ground points covers a considerable distance. An excellent rule to remember is: When in doubt, use a repeat coil. Repeat coils, as a rule, have wide frequency ranges and produce little or no effect on the circuits in which they are connected if the impedance match is satisfied. The insertion loss is generally between 0.10 and 0.50 dB. An electrostatic shield is interposed between the primary and the secondary windings and brought out to a separate terminal for grounding. This electrostatic shield permits transmission only by inductive coupling between windings. Electrostatic noises such as

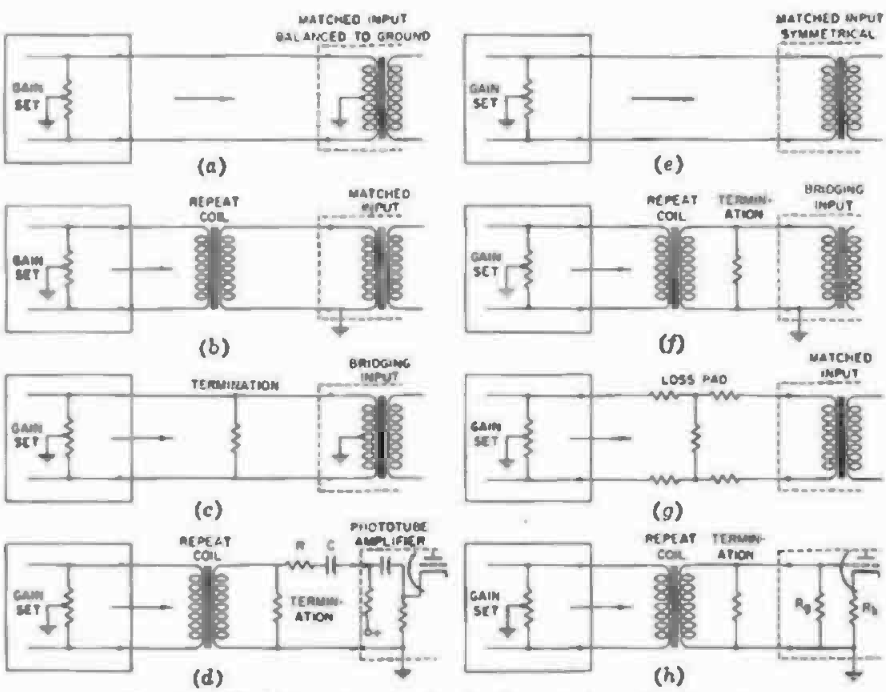


Fig. 23-14D. External send circuits for a balanced gain set.

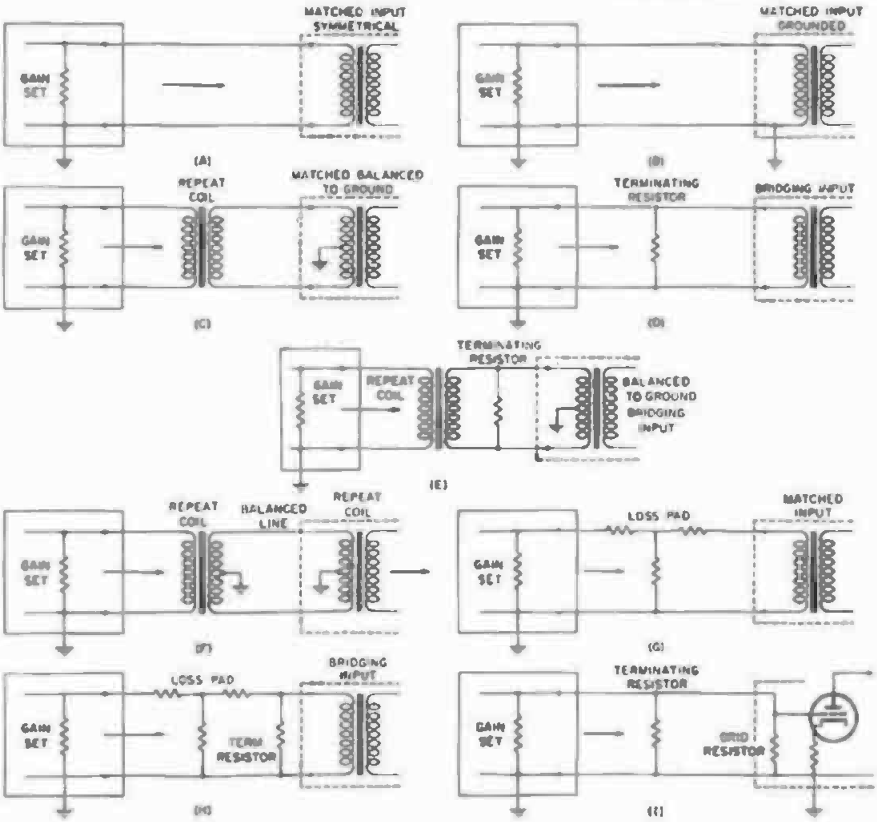


Fig. 23-15. External send circuits for an unbalanced gain set.

the breaking of power circuits and radio signals are shunted to ground by the shield. (See Question 23.199.)

23.15 Show the different types of input circuits used with an unbalanced gain set.—In Fig. 23-15 are shown several circuits suitable for use with an unbalanced gain set when sending into input circuits such as shown in Fig. 23-14B. Other input circuits are also shown in Fig. 23-15.

23.16 How is a device with a bridging input impedance measured?—When a device which employs a bridging input impedance is measured, the send section of the gain set must be terminated in its normal send impedance with a terminating resistor, as shown in (e) of Fig. 23-14D and in (D) of Fig. 23-15, to provide the proper load for the gain-set attenuators.

As a rule, a terminating resistor is provided in the gain-set send section and may be switched in as required. If the send section is not terminated when measuring a bridging input impedance, the gain set sees only the high impedance of the bridging input. The loss or actual level indicated by the attenuator dials will not be correct; also, the loss of the attenuators will not increase or decrease in exact 1-dB steps when the loss is less than 6 dB.

The terminating resistor also provides the terminating impedance for the bridging circuit. Bridging impedances, as a rule, have values between 10,000 and 30,000 ohms and are designed to operate from a circuit impedance of 500 or 600 ohms. Thus, with the correct terminating resistor at the send terminals of the gain set, the bridging impedance is measured under its normal operating conditions and a true measurement of the gain (if it is an amplifier) is obtained.

At times the bridging impedance of an amplifier may be specified to be 150 ohms. If so, the gain set is set for 150 ohms and then terminated in 150 ohms. The device is then measured in the normal manner. Many times bridging amplifiers are used across different circuit impedances. The expected gain for each value of terminating impedance may be measured by the previous method. As the terminating impedance is lowered in value, the measured gain of the amplifier is also lowered. A terminating impedance of extremely low value should not be employed without consulting the manufacturer's data sheet for the device, as the frequency response may be materially affected by a low value of terminating impedance.

23.17 ***What are the different terms used to express the gain of an amplifier?***—The term "gain" as used in electronics is rather a loose term. It is usually understood to mean voltage gain, unless specifically stated otherwise. Other terms are power gain and current gain. If the term "gain" is generalized to "transfer," it can then include voltage-to-current gain (transadmittance) and current-to-voltage gain (transimpedance). The following symbols could then be used: T_v, voltage transfer or e_o/e_{in}; T_i, current transfer i_o/i_{in}; and T_p, power gain. Thus an amplifier whose T_z equals 10^6 is a transimpedance-type amplifier. It requires a current input and gives a voltage output, the values of which are determined by multiplying the input current by 10^6. Given in the table below are four possible ways to arrange voltages and currents as inputs and outputs for devices.

23.18 ***Can an impedance-matching transformer be connected in the send section of a gain set?***—Yes, provided its frequency characteristics are such that it does not induce a frequency characteristic of its own. A coil used with a gain set should have a frequency response wider than the bandwidth to be measured. Very often an impedance-matching transformer is permanently connected at the send terminals of the gain set for convenience of operation. If the absolute input signal level must be known, the insertion loss of the coil must be included in the summing up of the attenuator losses. Gain sets using unbalanced configurations often include a repeat coil in the design to eliminate the need for an external coil.

As a rule, repeat coils are not used in the output circuit of a device developing any amount of power. However, it is not uncommon to use a repeat coil in the output of an equalizer or filter, when it is followed by an amplifier or attenuator to isolate an unbalanced circuit from a balanced circuit or to eliminate a ground loop. If it is necessary to use a coil in the output of an amplifier, precautions must be taken to ensure that it is not overloaded, because this would affect the frequency response of the device being measured. The Daven gain set, described in Question 22.122, has a built-in repeat coil as a part of its normal configuration.

23.19 ***If a repeat coil is used with a bridging input, where is the terminating resistor connected?***—The terminating resistor is always connected on the side of the coil nearest the device being tested, as shown in (d) of Fig. 23-14D, and Fig. 23-15E. This ensures that the frequency response of the coil will not be affected by improper coil loading.

23.20 ***If an external repeat coil is used with a gain set, how is its insertion loss accounted for?***—The insertion loss is added to the loss of the attenuators. As a rule, the average high-quality repeat coil has an insertion loss of 0.25 dB at 1000 Hz.

23.21 ***If an external attenuator is used with a gain set, how is its insertion loss accounted for?***—The insertion loss of the attenuator is added to that of the gain-set attenuators. This is assuming that the far end of the attenuator is terminated in its correct load impedance. If the attenuator is not loaded properly, its loss will not be the stated loss. (See Fig. 23-21.)

23.22 ***If the zero calibration mark on a VU meter represents a plus 4 dBm, where is 0 dBm on the scale?***—At the minus 4-dBm calibration mark. The reason for this is described in detail in Question 10.17.

23.23 ***When equipment is measured at a distant point, how is the formation***

INPUT		*OUTPUT*		*Equation*
Signal	*Impedance*	*Signal*	*Impedance*	*& Symbols*
Voltage	High	Voltage	Low	Voltage Gain $e_o/e_{in} = T_v$
Current	Low	Current	High	Current Gain $i_o/i_{in} = T_i$
Voltage	High	Current	High	Trans-admittance $i_o/i_{in} = T_y$
Current	Low	Voltage	Low	Trans-impedance $e_o/i_{in} = T_z$

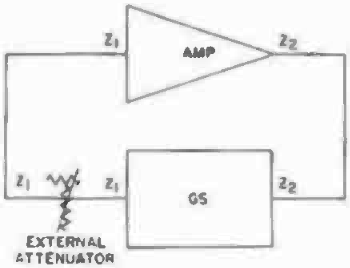


Fig. 23-21. Use of an external attenuator with a gain set. The loss of the external network is added to the loss of the gain-set attenuators.

of ground loops prevented?—By the use of a repeat coil in the transmission line at the send end. This separates the physical grounds at the two ends of the measurement, thus preventing a return circuit via the ground. (See Fig. 22-23.)

In some instances a second repeat coil is connected at the receive end. In this latter type of connection, the center taps of the repeat coils are grounded to prevent the flow of longitudinal currents, as explained in Question 8.42.

23.24 ***What does the term "transmission ground" refer to?***—To the ground connection of the transmission-measuring equipment. Equipment to be tested is connected to this ground to ensure that the ground potential will be the same for the equipment under test as for the transmission-measuring equipment.

The transmission ground is connected to a water main or to a special ground consisting of a copper plate buried in moist ground. No other equipment is connected to this ground except the transmission equipment and that under test. The installation of transmission and system grounds is discussed in Question 24.33.

23.25 ***Why is it necessary that the meters in a two-meter gain set have identical frequency characteristics?***—Suppose an amplifier is measured and that it has a frequency response known to be uniform over a frequency range of 20 to 20,000 Hz. It will be further assumed that the send meter has uniform frequency characteristics, but the receive meter connected across the output of the amplifier (Fig. 23-25) has a loss of 2 dB at 15 kHz compared to a reference frequency of 1000 Hz. If frequencies are applied to the input of the amplifier at a constant amplitude as read on the send meter at the output of the oscillator, the frequency response indicated by the receive meter will show a loss of 2 dB at 15 kHz.

Now, if the positions of the two meters are reversed and the same measurement is again made, it will be noted the amplifier frequency response will

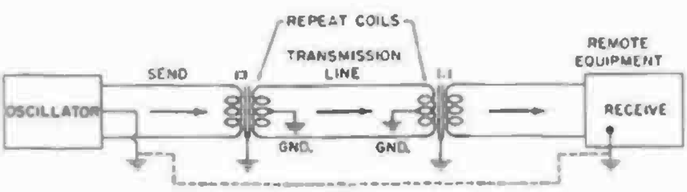


Fig. 23-23. The use of repeat coils in a transmission line to prevent the formation of a ground loop via the ground.

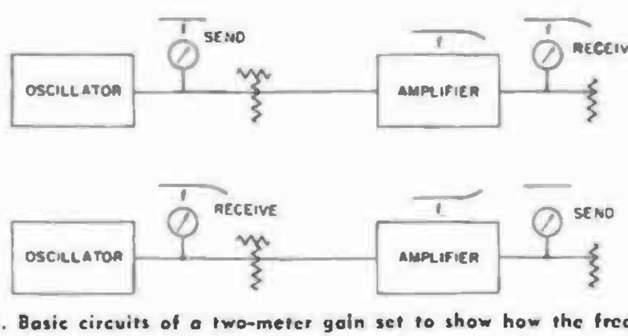


Fig. 23-25. Basic circuits of a two-meter gain set to show how the frequency characteristics of the meters will affect the measured response. The frequency response of each meter is indicated by the line above each meter.

show a rise of 2 dB at 15 kHz. The discrepancy between the two measurements is due to the fact that in the first measurement the send meter had a uniform frequency characteristic and all frequencies at the output of the oscillator were sent at exactly the same amplitude; however, the receiving meter had a loss of 2 dB at 15 kHz, so the amplifier response appeared to be down 2 dB at 15 kHz.

In the second measurement the receive meter is connected across the oscillator output and the send meter across the amplifier output. Because of the 2-dB loss at 15 kHz which is now in the send meter, the amplitude at the higher frequencies is increased 2 dB at 15 kHz. Under these conditions the amplifier will show a rising characteristic as the higher frequencies are approached.

If both meters have exactly the same loss at the higher frequencies, the error is cancelled; however, the actual signal voltage at the input to the amplifier will be higher at 15 kHz than at the lower frequencies. This is not too important, unless the increase of voltage at the amplifier input overloads the first stage. It should also be pointed out that if a rectifier-type meter is used with a vacuum-tube voltmeter, there can be a wide frequency discrepancy between the two meters, because of the limited frequency range of the rectifier-type meter compared to that of the vacuum-tube voltmeter. Also, rectifier-type meters are affected by the magnitude of harmonic frequencies contained in the signal under measurement and, in many instances, will not read the correct voltage. (See Question 22.103.)

23.26 ***How is the tracking of a two-meter gain set checked?***—The gain-set 600-ohm send terminals are patched to the 600-ohm receive terminals, both VU-meter attenuators are set to plus 4 dBm, and all loss is removed from the gain-set attenuators. Now, adjust the level at the send VU meter to plus 4 dBm. This should bring the receive meter to a plus 4 dBm. The difference in level between the two meters should not exceed 0.2 dB at any frequency between 20 and 15,000 Hz. Usually commercially made gain sets provide an adjustment for setting the meters to absolute zero dBm and for tracking the meters with respect to each other.

23.27 ***How are the attenuators in a gain set checked for different combinations of loss?***—By connecting an amplifier with a variable-gain control between the send and receive sections, as for a normal gain versus frequency measurement. The attenuators are set for different combinations of loss for a given setting of the amplifier gain control.

If the attenuators are operating properly, the gain will be measured the same regardless of the combination of loss setting for the attenuators. This test should be made at frequencies of 20 and 15,000 Hz (or the maximum frequency of measurement specified by the manufacturer). It is to be expected that some variation will be indicated, the amount depending on the accuracy of the attenuators. As a rule, the variation in a well-designed gain set is not more than 0.10 dB.

23.28 ***How is the frequency response of a gain set checked?***—In the manner described in Question 23.27. The attenuators are set for different values of loss, and the frequency response is checked at 20 and 15,000 Hz. If the gain set is operating properly, the variation will be on the order of 0.10 dB. However, it should be pointed out that difficulties may arise when making such measurements because of unbalance in the amplifier circuit, ground loops, leakage, and other sources. All these possibilities must be explored before assuming that the difficulty is in the gain set.

Discrepancies are likely to show up above 5000 Hz, if any of the previously mentioned troubles are apparent. The first test should be a 10,000-Hz turnover test at medium gain as described in Question 23.58.

23.29 ***What frequency is generally used for a reference frequency when audio-frequency measurement data are plotted?***—For most work, 1000 Hz; however, 400 and 800 Hz are often used for equalizer plotting.

23.30 ***What are the standard impedance values used for input circuits?***—50, 150, 250, and 600 ohms. In some older equipment, 30, 200 and 500 ohms were also used.

23.31 ***What are the standard impedance values used for output circuits?***—4, 8, 16, 150, and 600 ohms. For older equipment, 10, 250, and 500 ohms may also be included.

23.32 If an amplifier input circuit is rated at 600 ohms, can it be assumed that it will supply a 600-ohm termination to the source feeding it?—No, because many input circuits, though rated to work from a 600-ohm source, do not necessarily present 600 ohms to the source impedance. This may be checked by noting whether the secondary of the input transformer is terminated by a resistance. If it is terminated, it may more than likely present a 600-ohm impedance to the source. If the transformer is not terminated on the secondary side, it is treated as an open-circuit input, as described in Questions 4.75, 8.30, 8.48, and 23.50.

23.33 If the secondary of an input transformer is unterminated, how does the primary impedance appear?—It appears as the reactance of the primary winding, paralleled by the reflected loading of the tube input capacitance and the reflected characteristics of the secondary winding.

23.34 How are output circuits treated with respect to grounding when using a gain set?—Four output circuits, the most commonly encountered in audio-frequency equipment, are shown in (a) to (d) of Fig. 23-34. At (a) of Fig. 23-34 is shown a symmetrical or ungrounded output circuit. It may be measured with or without a ground connection. However, if the amplifier gain is 60 dB or more, it might be advisable to ground the lower end of the output winding to insure stability and reduce leakage at the high frequencies. The circuit in (b) of Fig. 23-34 is a grounded and unbalanced one. As the ground shown is generally connected internally in the amplifier, only the normal chassis-ground connection is required when measuring this circuit.

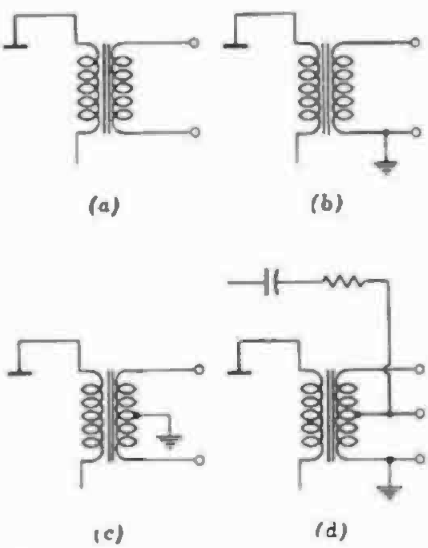


Fig. 23-34. Output circuits most commonly encountered in audio-frequency equipment. (a) Symmetrical. (b) Grounded (unbalanced). (c) Center tap to ground (balanced). (d) Grounded with negative-feedback loop taken from output tap (unbalanced).

In (c) of Fig. 23-34, the output transformer winding is center-tapped to ground and is a balanced circuit. Only the normal chassis ground is used. In (d) of Fig. 23-34 is shown the output circuit most commonly used in negative-feedback amplifiers. The feedback loop is connected to one of the output taps. In this type of connection, the circuit is unbalanced and can only be measured using a chassis ground as shown, because a ground connection on the gain-set receive terminals may ground out the output signal.

The best rule to follow is: Leave the gain-set receive section ungrounded and apply a ground as required. When the output circuit of the device under test is grounded, use only the chassis ground. Operating in this manner ensures that ground loops will not be formed and that the output will not be grounded out by a second ground connection.

23.35 How can the gain of an amplifier be measured without an external pad, if the gain exceeds the maximum attenuator loss of the gain set?—If the total loss of the attenuators in the gain set is only 100 dB and the amplifier gain is 130 dB with a rated power output of 10 watts (plus 40 dBm), the gain could be measured by setting the receive-meter attenuator to a plus 30 dBm. This would require a loss of 100 dB in the gain-set attenuators. The level at the receive meter would read a plus 30 dBm. Adding the loss of the attenuators to the indication on the receive meter results in a total of 130-dB gain.

Very often, when making measurements as previously described, difficulties arise because of internal feedback in the gain set due to coupling between the send and receive sections. However, most gain sets are designed to measure at least 110 dB with an output power level of 20 watts (plus 43 dBm).

A better way of measuring the preceding amplifier is to connect an external pad of 20- to 40-dB loss between the send terminals of the gain set and the input of the amplifier; thus, the signal level is reduced. Using a 40-dB pad with a 0-dBm signal from the gain-set send section would require only 86 dB of loss in the attenuators for an output level at the receive section of plus 4 dBm.

If the amplifier output power exceeds that specified for the gain set, an external termination is used and the gain is computed on the basis of the loss of the gain-set attenuators plus the output level of the amplifier.

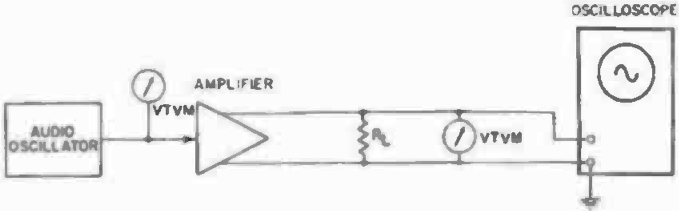
23.36 What is a power gain versus frequency measurement? — To properly evaluate the frequency and power output capabilities of an amplifier, it is necessary to make three frequency measurements: one near the maximum power output, a second at 0 dBm, and a third at 20 dB below 0 dBm. Two methods may be used to determine the point of maximum power output.

In the first method the amplifier is terminated in its normal load impedance, an oscilloscope is connected across the load termination, and a 400-Hz signal is applied to the input (Fig. 23-36A). The output waveform is observed as the input signal level is increased. At the first indication of a departure from a sine wave, the power developed is noted and taken as the maximum power-output level.

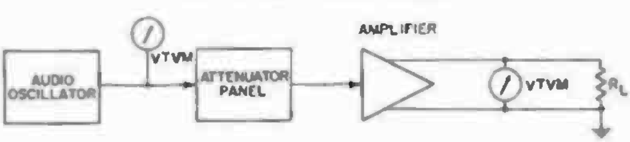
The second method of determining the maximum power output requires the use of an attenuator, calibrated in decibels, connected in the input (Fig. 23-36B). Assume the amplifier is capable of producing a power output of 10 watts (plus 40 dBm). Set the meter across the output termination to read a plus 40 dBm. Remove loss from the attenuators until a plus 40 dBm is obtained. Insert 3 dB of loss in the attenuators. The amplifier output should drop 3 dB (half power), or to plus 37 dBm. Now remove 1 dB of loss from the attenuators. The output level should increase exactly 1 dB. Remove an additional 1 dB of loss. The output should again increase 1 dB. A third 1-dB step is removed. If the amplifier is not being driven into overload, the output will increase exactly 1 dB. However, as this is near the maximum power output of the amplifier, it may be driven into overload. Overloading will be indicated by an increase of slightly less than 1 dB at the output for an increase of 1 dB at the input. (See Question 23.14.)

In the average amplifier the maximum power output will increase about 0.8 dB. If the amplifier is driven very far into overload, the output will show an increase of only 0.5 dB or less. The harder the amplifier is driven into overload, the less will the output level increase. (This takes place because the amplifier is incapable of producing any more power.)

As a rule, when determining the maximum power point as described, the maximum power output is taken at the



(a) *Using an oscilloscope.*



(b) *Using an attenuator panel.*

Fig. 23-36. Test circuits that are used for measuring the overload characteristics of a power amplifier.

output level where the amplifier increases 0.8 dB for an increase of 1 dB at the input. A frequency measurement is then made holding the input voltage constant over the frequencies of interest. Distortion is now measured, and the maximum power rating is then set from the results of these two measurements, stating that the amplifier has a maximum power output of 10 watts with a total rms harmonic distortion at 400 Hz of a given percentage. If the maximum power output is to be rated in percent intermodulation, the rating must be made in accordance with the data given in Question 23.113.

If an external load resistance is used, the power across the load resistance R_L may be computed:

$$\text{Watts} = E^2/R_L$$

where,

E is the voltage developed across load resistor R_L.

If a VU meter is used rather than a vacuum-tube voltmeter across the load resistance, and the value of R_L is different from that for which the VU meter is calibrated, a correction factor must be applied to the meter to obtain the true value of output power. The correction factor may be computed:

$$\text{dB} = 10 \text{ Log}_{10} (Z_1/Z_2)$$

where,

Z_1 is the meter impedance,
Z_2 is the load impedance.

The use of the correction factors and values most commonly employed are given in Question 10.32. (See Question 23.38.)

23.37 ***How is the gain of a recording system computed?***—Starting at a given input of the system, the losses induced by the mixer controls and networks, coils, filters, and other devices are added. The gain, including the preamplifiers, booster, and line, and recording amplifiers, is then added. The total loss is then subtracted from the total gain. The remainder is the new gain of the system. As an example, if the total gain of the system is 170 dB and the loss used during normal operation is 86 dB, the net gain of the system is 84 dB. Thus, if a bridging bus level of plus 14 dBm is required, the lowest-level signal that can be applied to the input for a bridging bus level of plus 14 dBm is minus 70 dBm.

23.38 ***How is the power output of a high-power amplifier measured?***—By using a resistive voltage divider across the output of the amplifier as shown in Fig. 23-38.

Suppose that an amplifier rated at 100 watts of output and having a 16-ohm output winding is to be measured. Four 16-ohm, 50-watt resistors are connected in series-parallel across the 16-ohm output winding as shown. A vacuum-tube voltmeter is connected across one of the 16-ohm resistors. The measured power is multiplied by four to obtain the true power. Thus, for an amplifier with a 100-watt output, a power of 25 watts will be indicated. Using the circuit shown, the power output of the amplifier is carried by the terminating resistors and only the voltage drop across the resistor is measured by the meter. The accuracy of the measurement will depend on the accuracy of the resistors. The circuit shown may also be used for making a power versus frequency run as described in Question 23.36.

23.39 ***At what power-output level should transistor amplifiers be measured?***—The measurement of harmonic and intermodulation distortion in transistor amplifiers is treated somewhat differently than that for vacuum-tube amplifiers. It is customary when measuring a tube amplifier to terminate it in a specified resistive load and then measure the distortion at the half-power and full-power points. The amplifier is then rated at these two points. The same procedure is measured in equivalent sine-wave power. (See Question 26.36.)

Transistor amplifiers are also measured using a resistive termination. However, since transistor amplifiers generally employ class-B push-pull output stages, they are measured at several different power-output levels. This procedure is necessary because of the crossover or notch distortion induced by operating the output stage class-B. (See Question 12.227.) Therefore, distortion measurements are made

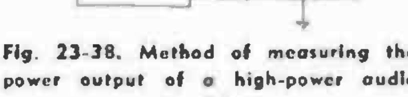


Fig. 23-38. Method of measuring the power output of a high-power audio amplifier.

at 3, 7, 10, 13, 17, and 20 dB below the reference power output, for frequencies between 20 and 20,000 Hz.

If the input stage does not include a capacitor, one must be connected in series with the test circuit to prevent the low dc resistance of the test circuit from shorting out the input stage bias voltage. Similar precautions should be taken with the output stage as the output transistors may be damaged.

Measurements are not required above five times the reference distortion or lower than 30 dB above the residual hum and noise. Reference power and distortion are defined as the values stated by the manufacturers of the equipment. Power bandwidth is defined by the two frequencies where the curve of distortion versus frequency taken 3 dB below the reference power output crosses the line of reference distortion. (See Question 23.7.)

23.40 How is a resistance-coupled amplifier input measured using a gain set?—As shown in (h) of Fig. 23-14D, and Fig. 23-15I. If the input of the amplifier circuit includes a bias battery, a capacitor is connected in series with the grid side to prevent the shorting of the battery by the dc resistance of the repeat coil.

A resistance-coupled amplifier stage measured in the manner just explained will not show the actual gain of the stage, because the high impedance of the input is considered to be a bridging input fed from a low-impedance source. Therefore a correction factor is applied to the apparent gain to obtain the measure of true gain. The correction factor may be calculated:

$$dB = 10 \, Log_{10} \frac{Z_2}{Z_1}$$

where,
Z_1 is the gain-set send impedance,
Z_2 is the input resistance of the amplifier (grid-resistor value).

A typical example would be a gain-set send impedance of 500 ohms bridged by an amplifier with an input resistance of 500,000 ohms:

$$10 \, Log_{10} \frac{500{,}000}{500} = 1000$$

$$10 \, Log_{10} 1000 = 10 \times 3 = 30 \text{ dB}.$$

Thus, an amplifier which has an apparent gain of 20 dB has a true gain of 50 dB.

23.41 If the power-output level of an amplifier is given in decibels, how is this converted into watts?

$$\text{Power in Watts} = \text{Antilog} \frac{dB}{10} \times \text{Reference Level}$$

where,
dB is the output level,
reference level is 1 or 6 milliwatts.

23.42 If the power output of an amplifier is stated in watts, how is it converted to decibels? Answer:

$$dB = 10 \, Log_{10} \frac{P_1}{P_2}$$

where,
P_1 is the stated power output in watts,
P_2 is the reference level in watts.

23.43 How is the voltage gain calculated for an amplifier of equal input and output impedances? Answer:

$$dB = 20 \, Log_{10} \frac{E_2}{E_1}$$

where,
E_1 is the voltage at the input,
E_2 is the voltage across the output load termination.

23.44 How is the voltage gain calculated for an amplifier of unequal input and output impedances? Answer:

$$dB = 20 \, Log_{10} \frac{E_2 \sqrt{Z_1}}{E_1 \sqrt{Z_2}}$$

where,
E_1 is the voltage at the input,
E_2 is the voltage across the output load termination,
Z_1 is the input impedance,
Z_2 is the output load impedance.

23.45 What effect does unbalance in a push-pull amplifier stage have on the amount of harmonic distortion, and how is it measured? — Unbalance in a push-pull stage has little effect on the harmonic distortion above 100 Hz. However, below this frequency the distortion rises quite rapidly as the degree of unbalance is increased, as shown graphically in Fig. 23-45A. The measurement data shown were made on a Williamson Ultralinear amplifier using KT-66 tubes in the output with a balancing control in the cathode circuit, as shown in Fig. 23-45B. To balance the push-pull stage, a dc voltmeter is connected between the upper ends of the cathodes, and balancing potentiometer P1 is adjusted for a zero indication on the meter. This balance is made with the

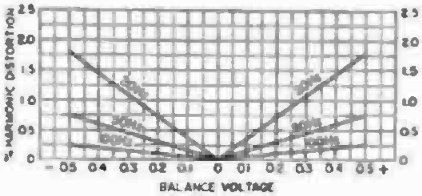


Fig. 23-45A. Harmonic distortion in push-pull output stage due to unbalance of plate current.

input of the amplifier terminated to prevent noise pickup. After a balance has been obtained, a distortion measurement is made at a given power output. To measure the effect of unbalance, balancing pot P1 is set for different values of unbalance as indicated on the voltmeter across the cathodes. Distortion is measured at 20, 50, and 100 Hz for different settings of the balancing pot, using exactly the same power output as was used for the balanced condition. The results are plotted as shown in Fig. 23-45A.

23.46 How may a phase inverter driving a push-pull output stage be balanced for minimum distortion?—It is the general belief that if the two tubes of a push-pull amplifier stage are similar in characteristics, the stage is in balance and that a condition of lowest distortion (second harmonic) exists. However, this is not always true because of discrepancies in components and a difference in the gain between the two sides of the phase inverter driving the push-pull stage. Considerable unbalance can also exist in the push-pull stage because of the difference in transconductance of the tubes and differences in the coupling capacitors, grid resistors, and the two

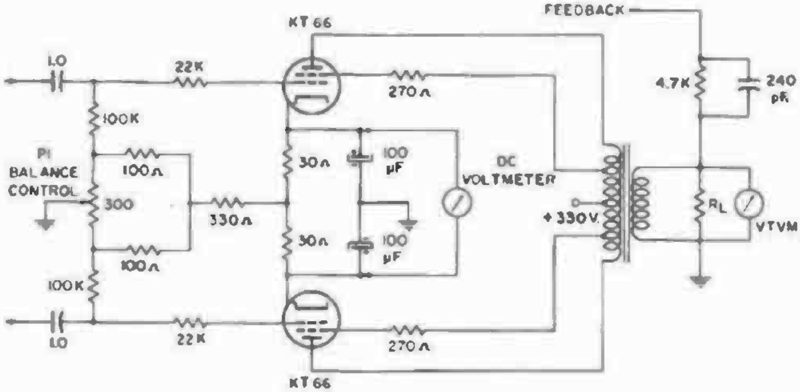


Fig. 23-45B. Output stage and cathode-balancing circuit for Williamson Ultralinear amplifier employing KT-66 tubes.

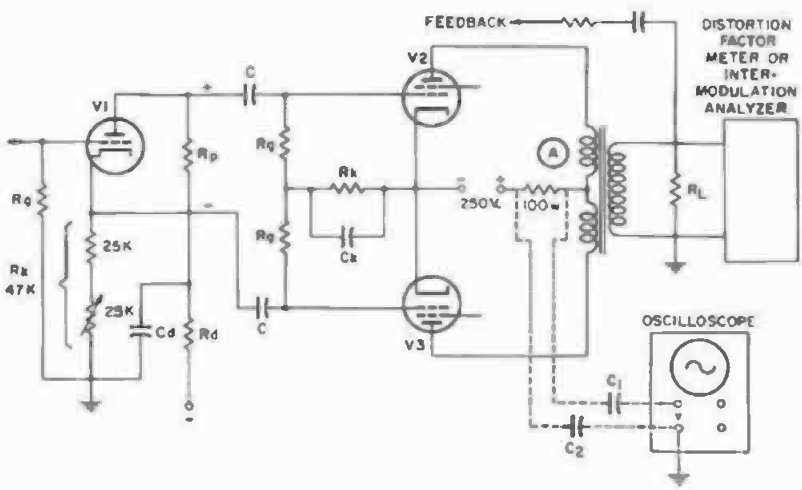
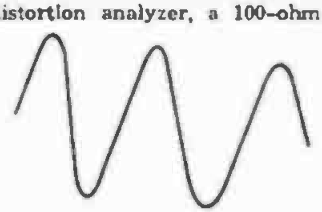


Fig. 23-46A. Phase-inverter and push-pull amplifier stage. Minimum distortion is achieved by the adjustment of the variable resistor in the cathode circuit of the phase-splitter tube.

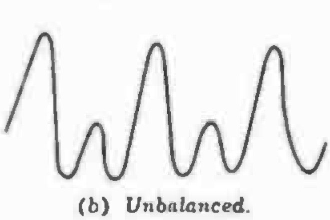
halves of the output transformer. It then appears that some method of balancing the phase inverter should be provided so that the push-pull stage and the phase inverter can be balanced for minimum distortion. Such a balance may be achieved by making a portion of the load resistance of one side of the phase inverter variable, as shown in the circuit of Fig. 23-46A, a split-load phase inverter. To balance the circuit for minimum distortion, a signal is applied to the input of the phase-inverter stage and the output stage driven to near its maximum output level but not into overload. A distortion-factor meter or intermodulation analyzer is connected across the normal load resistor of the push-pull stage and the variable resistor in the cathode of the phase-inverter stage adjusted while the magnitude of the distortion in the output is observed.

Any given adjustment will hold true only for a particular pair or combination of tubes. The circuit will require rebalancing each time the tubes are replaced or their positions changed. The cathode resistor may be readjusted from time to time to keep the distortion at a minimum during the normal life of the tubes. Tubes for both the phase-inverter and the push-pull stages should be selected for the closest possible match by means of a transconductance (mutual-conductance) tube tester.

In the absence of an intermodulation or distortion analyzer, a 100-ohm resistor may be connected in the common B+ lead of the push-pull output stage shown in Fig. 23-46A. A cathode-ray oscilloscope is connected in series with two 1-μF paper or oil capacitors, C1 and C2, across the 100-ohm resistor to permit the observation of the waveform in the push-pull stage as the balance of the phase inverter is adjusted.



(a) Balanced.



(b) Unbalanced.

Fig. 23-46B. Character of waveforms observed across a 100-ohm resistor connected in the B+ lead of a push-pull output transformer while adjusting the balance of the phase inverter for minimum distortion.

The foregoing method is not quite as satisfactory as measuring the distortion at the output of the push-pull stage; however, a condition of balance can be determined fairly well by the character of the waveform observed across the 100-ohm resistor. When the distortion is at a minimum, the waveform will appear as in (a) of Fig. 23-46B, and when at a higher value than minimum, it will appear as in (b) of Fig. 23-46B.

Although a condition of minimum distortion can be determined fairly closely by the use of an oscilloscope, it is not always true that the distortion is at a minimum when the peaks of the waveform are equal. However, since the difference in distortion will be quite small, it may be assumed that if the peaks appear to be of equal height, the phase-inverter stage is in balance and the distortion is at a minimum.

23.47 Show the effects of a cathode-bypass capacitor in a push-pull stage. —In Fig. 23-47 are shown the results of intermodulation measurements made on a conventional negative-feedback amplifier for conditions of no capacitance and for values of 50 and 100 μF across the common cathode resistor. It will be noted that without the bypass capacitor, the amplifier produced a power output of 14 watts with 18-percent intermodulation distortion. Connecting a 50-μF capacitor across the cathode resistor resulted in increasing the power output to 16.10 watts for approximately the same amount of distortion. Increasing the size of the capacitor to 100 μF increases the power output to only 16.4 watts for the same value of distortion.

No attempt was made to select the tubes or to balance the phase-inverter stage. The measurement was made as described in Question 23.113 for making conventional intermodulation measurements.

23.48 What peculiarity will be noted when testing the linearity of a class-AB amplifier? — When testing a class-AB amplifier for linearity, it will be noted

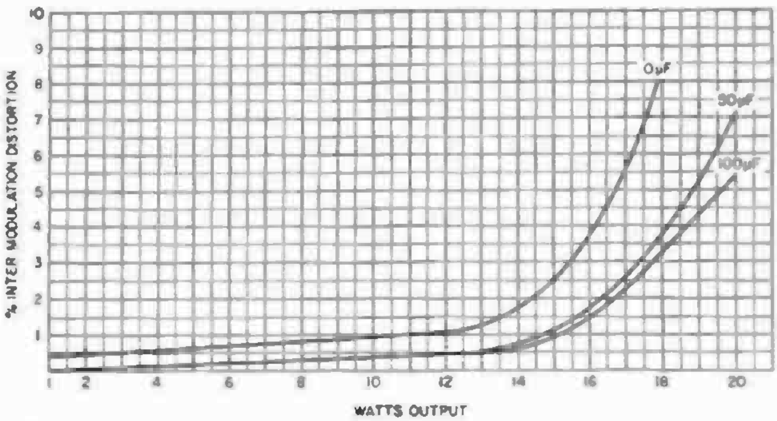


Fig. 23-47. Intermodulation distortion measurements of a conventional negative-feedback amplifier with and without a cathode bypass capacitor in the push-pull output stage.

that a linear response will be obtained up to about one-half the power output. Then the output will increase at the rate of approximately 0.8 dB for a change of 1 dB at the input. (See Question 23.36.) This action will continue to the point where the amplifier is driven into overload, at which point a sharp departure from this almost linear re- point and then function as class-AB.

This is typical of many class-AB amplifiers, as they function as a class-A amplifier up to about the half-power point and then function as class-AB.

23.49 ***What is the procedure for making transient tests?***—Transient measurements may be made in a number of different ways. In one method a square-wave signal is applied to the input of the amplifier and pulsed. A second method also uses a square wave, but changes the amplitude and time duration of the pulses. The third method as given in IHF Standard A-201-1966 is discussed in Question 23.209.

The test circuit of Fig. 23-49A employs a special pulse generator that generates a signal waveform as shown in Fig. 23-49B. The input and output signals are monitored by a dual-trace oscilluscope. Any departure from the waveshape of the input signal appear-

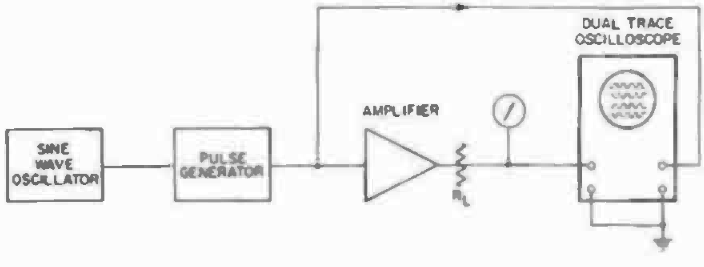


Fig. 23-49A. Block diagram for transient tests, using a pulse or tone-burst generator.

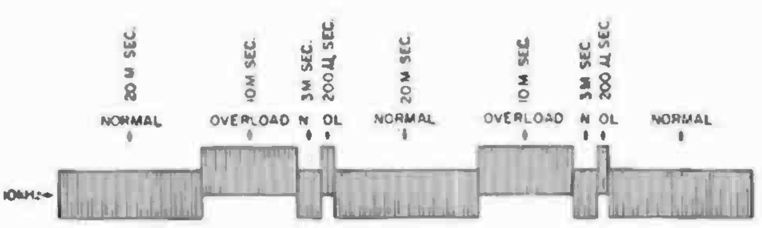


Fig. 23-49B. Waveform characteristic used for testing the transient response of an audio amplifier.

ing in the output signal is considered to be distortion. No hard and fast rules can be given for transient measurements since the circuit components have a great effect on the output waveform. Considerably more transient distortion may be expected if the amplifier employs transformers.

A 10,000-Hz signal of constant amplitude is applied to the amplifier at a given input level well below the overload point. The level of the input signal is suddenly increased to a value that will simulate a reasonable overloaded condition. This sudden increase at the output is observed on the oscilloscope connected at the load termination. Any distortion of the original waveform will indicate ringing or hangover of the amplifier. It is highly important that the increase in level at the input be fast and with as little rise time as possible.

At the end of the overloaded portion of the signal, the system is permitted to return to its normal input level for a short time. The level is again increased, only this time for a short period. This short overload signal close on the long overload period checks the ability of the amplifier to recover after a long overloaded period. The signal is again returned to its normal level and the cycle repeated. The sequencing and time intervals are illustrated in Fig. 23-49B.

Originally, the term "transient" was employed to describe what took place when a piece of equipment was turned off or on, or when some unusual disturbance took place in a piece of equipment or on a power line. It is present-day practice to use the term "transient" to indicate any type of nonsinusoidal disturbance.

23.50 ***Describe an injection circuit for measuring the frequency characteristics of a microphone preamplifier.*** — In measuring the frequency response of an amplifier using an input transformer with an unterminated secondary (open circuit) specified to operate from a given source impedance, the transformer primary is, in reality, terminated by the source impedance.

(The following discussion applies to ribbon-velocity, moving-coil, or other microphones of similar design.)

In a microphone preamplifier operating as an open-circuit device, the microphone supplies the termination for the primary. Therefore, if the amplifier response is measured by injecting a signal in series with the terminating impedance (microphone), the actual frequency response of the amplifier when terminated by the microphone can be measured. To obtain such a measurement, the equipment is connected as shown in Fig. 23-50. A resistor R1 of 1-ohm resistance is connected in series with the output of the microphone and the primary of the impedance-matching transformer contained in the base of the microphone housing. A second resistor, R2, of 600 ohms or more is connected in series with the high side of the measuring circuit (or gain-set send section). These two resistors form an L pad which injects the test signal in series with the primary of the impedance-matching transformer of the microphone. During these tests, the diaphragm of the microphone must be covered to prevent acoustic pickup. For microphones of 250-ohm impedances, resistor R1 should not exceed 5 ohms. For impedances of 50 ohms or less, R1 may vary between 1 and 3 ohms.

The loss created by the combination of R1 and R2 should be on the order of 40 to 60 dB. The signal level injected into the microphone circuit should not exceed the normal voltage developed by the moving element, or the microphone may be damaged.

The gain versus frequency measurement is made by sending frequencies from the oscillator of constant amplitude as read on the VU meter. The gain variation is read on meter M at the output of the amplifier. If a second fre-

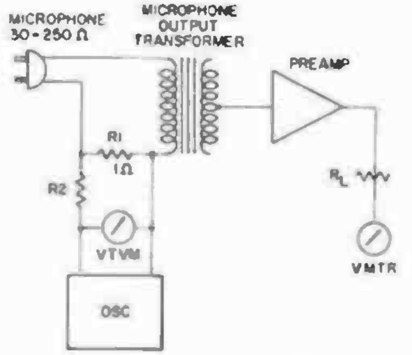


Fig. 23-50. Injection circuit for measuring the frequency characteristics of a microphone preamplifier using a microphone for the primary termination.

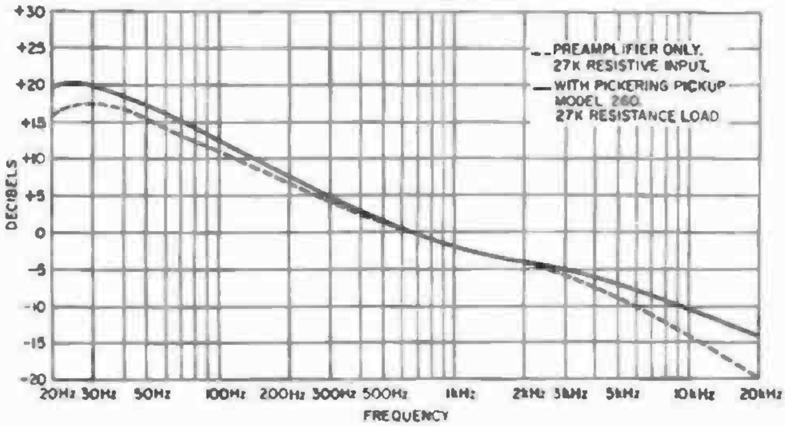


Fig. 23-51A. Frequency response of a magnetic-pickup preamplifier with pickup or resistive termination.

quency measurement is made by feeding the input of the amplifier from a resistive termination as normally used for such measurements, it will be found that the frequency response will be different from that obtained when the microphone is connected and the injection signal method of measurement is used.

The true frequency response is that measured when the microphone terminates the input to the amplifier. Equalization may be induced in the amplifier to correct for any discrepancies caused by the microphone if they are of any consequence.

23.51 ***Describe an injection circuit for measuring the frequency characteristics of a phonograph pickup preamplifier.***—The difference can be considerable, particularly at the low- and high-frequency ends.

Fig. 23-51A is the frequency response of a magnetic pickup working into a 27,000-ohm resistive load and measured using an injection circuit employing a resistor connected in series with the pickup head.

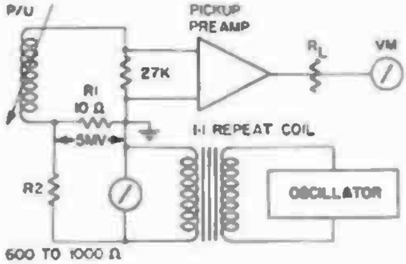


Fig. 23-51B. Injection circuit for measuring the frequency characteristics of a pickup preamplifier using the pickup as a termination.

Fig. 23-51B is an injection circuit suitable for measuring with magnetic pickups. A 10-ohm noninductive resistor R1 is connected in series with the pickup and input of the amplifier. Resistor R2 is connected in series with the output of the oscillator and acts as an attenuator to provide a means of injecting the signal in series with the pickup. Frequencies of constant amplitude are applied to the injection circuit. The voltage applied across the 10-ohm resistor should not exceed the pickup output voltage normally developed, which averages 2 to 10 millivolts.

From the frequency response in Fig. 23-51A, it may be seen that the frequency response varies considerably between the two methods of measurement.

23.52 ***Describe the use of an injection circuit for measuring the frequency characteristics of a magnetic reproducer preamplifier terminated by a magnetic head.***—An injection circuit for this type if measurement is given in Fig. 23-52A, and it may be employed to measure the frequency characteristics of a preamplifier when it is terminated by a reproducing head. The particular circuit shown is for a 30-ohm impedance head; however, it may be used also with high-impedance heads. A resistive network consisting of two noninductive resistors R_1 and R_2 is connected in the head circuit as shown, with a vacuum-tube voltmeter connected at the output of the oscillator for monitoring output voltage over the frequencies of interest. The voltage across resistor R1 (10 ohms) is held constant at a value of 2 to 5 millivolts and the frequency response observed on the voltmeter connected at the output of the preamplifier. A 1:1 repeat coil is connected in the output of the oscillator to isolate it from the measurement circuit. Series resistor R2 (1000 ohms) permits the oscillator to be operated at a fairly high level, thus increasing the signal-to-noise ratio.

It will be assumed for this discussion that the preamplifier has been equalized for the desired transmission characteristics (30 to 600 ohms) and the frequency response plotted for reference. A ground is connected at the low-potential side of the input transformer to prevent leakage at the high frequencies. If this ground is not employed, serious unbalance will be encountered at frequencies above 500 Hz.

The measurement is started by first adjusting the voltage from the oscillator across resistor R1 to, say, 4 millivolts at a frequency of 1000 Hz, and the output level of the preamplifier adjusted for a value 20 dB below the amplitude of the peak frequency of the low-frequency equalization to prevent overloading of the amplifier. A turnover test is then made at 10,000 Hz by reversing the output from the repeat coil and noting any change in the amplitude of the signal at the output of the preamplifier. If this test indicates more than a 0.20-dB unbalance, the unbalance must be cleared before continuing the tests. The output of the preamplifier is tested in a similar manner.

When conducting injection tests on a bench, the magnetic head is placed in a grounded Mumetal container, then rotated for the minimum pickup from

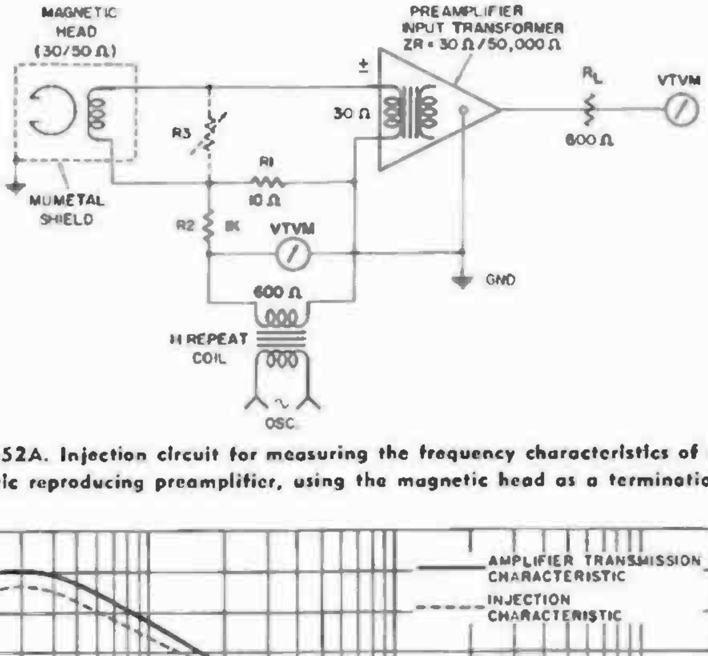


Fig. 23-52A. Injection circuit for measuring the frequency characteristics of a magnetic reproducing preamplifier, using the magnetic head as a termination.

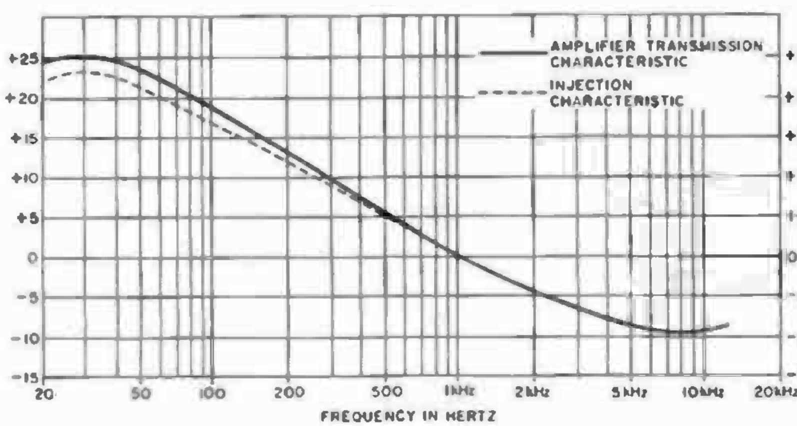


Fig. 23-52B. Comparison of transmission and injection measurement for a magnetic reproducer head.

stray magnetic fields. The pickup as indicated at the output of the preamplifier should be on the order of minus 50 to 60 dBm. Pickup effects may also be reduced by rotating the amplifier position relative to other equipment on the bench. Shielded leads must be used between connections from the test equipment and the network, with the shield grounded at one end only, and to a common ground point.

After the above tests have been completed, the actual frequency measurement is made by starting at 1000 Hz and continuing at a constant voltage downward in frequency. As the lower frequencies are approached the vacuum-tube voltmeter sensitivity will have to be decreased to compensate for the rising frequency characteristic of the preamplifier at the lower frequencies. (This may be 15 to 25 dB with reference to 1000 Hz.) After the low-frequency response is measured, the oscillator is returned to 1000 Hz and the measurement continued up to 12,000 Hz. In this area the voltmeter at the output of the preamplifier will require that its sensitivity be increased to allow for the decreasing high-frequency response.

When the measurement is completed, it is plotted against the transmission measurement (30 to 600 ohms) and compared (Fig. 23-52B). If necessary, the equalization is corrected to offset the discrepancies between the two curves. As a rule, they are held to within plus or minus 0.25 dB of each other.

After the installation of the head in the recorder and the azimuth and other basic adjustments have been made, a standard tape or magnetic film is played back and the high-frequency end is adjusted by connecting resistor R3 across the head. For a 30-ohm impedance head the value of R3 may vary from 100 to 800 ohms and will permit the adjustment of the frequency response between 500 and 12,000 Hz, over a range from minus 2 dB to plus 4 dB around 12,000 Hz.

If during the injection measurement resonant effects are indicated above 5000 Hz, they may be caused by reflections from the secondary due to the input capacitance of the tube and the inherent distributed capacitance of the transformer secondary. This is particularly true if the secondary impedance is on the order of 100,000 ohms. To prevent these effects, secondary impedances of 25,000 to 50,000 ohms are generally employed. It might be well to mention at this point that the input transformer must be of the hum-bucking type, and it must be enclosed in a nested shield. (See Questions 8.51 and 8.98 and 23.58.)

Generally, preamplifiers designed for reproducing magnetic sound track have a maximum output level of about plus 20 dBm. Therefore, it is extremely important that it be operated at a 1000-Hz reference level to prevent overloading at the peak frequency of the low-frequency equalization.

23.53 ***Describe the procedure for measuring transistor input and output impedances.*** — A typical transistor Q1 whose input and output impedances are to be measured is shown in Fig. 23-53, with a second transistor Q2 being used as a current regulator. The voltage and component values are for a typical transistor and can be altered to fit the particular transistor under test.

It is a well-known fact that maximum power is transferred from a generator to its external load when the external load impedance equals the generator impedance. Therefore the maximum power in a transistor circuit is transferred when the external load equals the internal output impedance of the transistor.

Referring to Fig. 23-53, the current-regulating circuit of Q2 offers an impedance of over 1 megohm at the audio frequencies. Resistors R3 and R4 form a bias voltage divider to stabilize the operation of transistor Q1. The procedure for measurement is:

1. Set resistor R1 in the emitter circuit of Q2 to its maximum value of 10,000 ohms.
2. Adjust current control R2 for the proper collector current in Q2 (about 1 mA).
3. Adjust R1 for the proper collector current through Q1.
4. Close S1 and apply a 1000-Hz signal from the signal generator (E1) to the input of Q1, with resistor R5 set to zero resistance and load resistor R6 open. Measure voltage E2.
5. Adjust resistor R6 to reduce the value of voltage E2 to one-half its value. Leave resistor R6 at this setting.

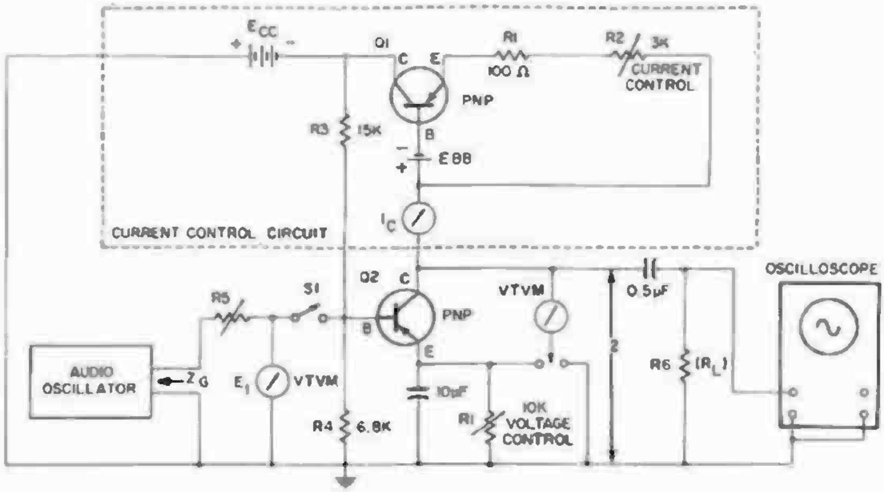


Fig. 23-53. Circuit for measuring the input and output impedance of a transistor. The values given are typical and will require some alteration for a given transistor.

6. Measure input voltage E1 with S1 open.
7. Close switch S1 and readjust resistor R5 until voltage E1 is reduced to one-half its value.
8. With R5 at this new setting and R6 open, read the output voltage E2.
9. Readjust R6 to reduce E2 to one-half its value.
10. Repeat Steps 6 and 7.
11. Record the values of R5, R6, E1, and E2.

While the measurements are in progress, the oscilloscope display must be continuously monitored for overloading and subsequent distortion of the signal. Input and output impedances may be calculated:

$$R_{in} = R5 + Z_g \qquad R_{out} = R6 = R_L$$

where,

R_{in} is the input impedance,
R5 is the value of the input resistor,
Z_g is the impedance of the signal generator,
R_{out} is the output impedance,
R6 is the load impedance.

The power gain may now be calculated:

$$P_g = \left(\frac{E_2}{E_1}\right)^2 \times \left(\frac{R5 + Z_g}{R6}\right)$$

where,

E_1 and E_2 are the input and output voltages respectively.

Power gain may be converted into decibels:

$$dB = 10 \operatorname{Log}_{10} (P_{out}/P_{in})$$

where,

P_{in} and P_{out} are the input and output powers respectively.

Input and output powers may be expressed:

$$P_{in} = \frac{E_{in}^2}{R5} \qquad P_{out} = \frac{E_{out}^2}{R6}$$

23.54 How can the attenuator section of a gain set be used to supply a known voltage?—The send section of a gain set is an attenuator network similar to that shown in Fig. 23-54A. If a known voltage is applied to the input of the network, a known voltage may be obtained at the send terminals by converting the attenuator loss to a voltage ratio. This may be done by referring to the graph in Fig. 23-54B.

To obtain a correct value of voltage at the output terminals of the network, the network must be terminated in a resistive load, as shown, which is equal in value to the attenuator resistance which, for this illustration, is 600 ohms. As an example, if 1 volt is applied to the input of the network and 40 dB of loss is inserted in the attenuators, the voltage at the output terminals will be 0.01 volt, or a ratio of 100:1. If 60 dB of loss is inserted, the voltage at the output will be 0.001 volt, or a ratio of 1000:1.

The VU meter in the send section of a gain set is, in reality, a voltmeter calibrated in decibels. By referring to a decibel versus voltage chart, such as the one in Fig. 23-54B, the VU meter

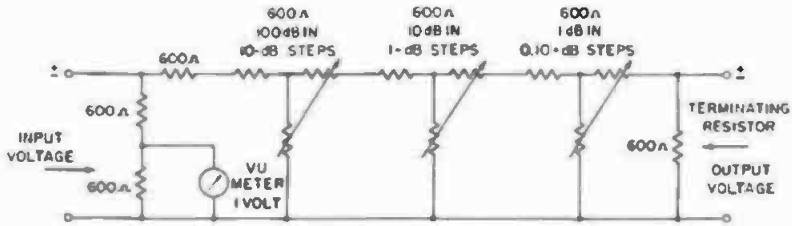


Fig. 23-54A. Attenuator section of a gain set connected to supply a known source of voltage.

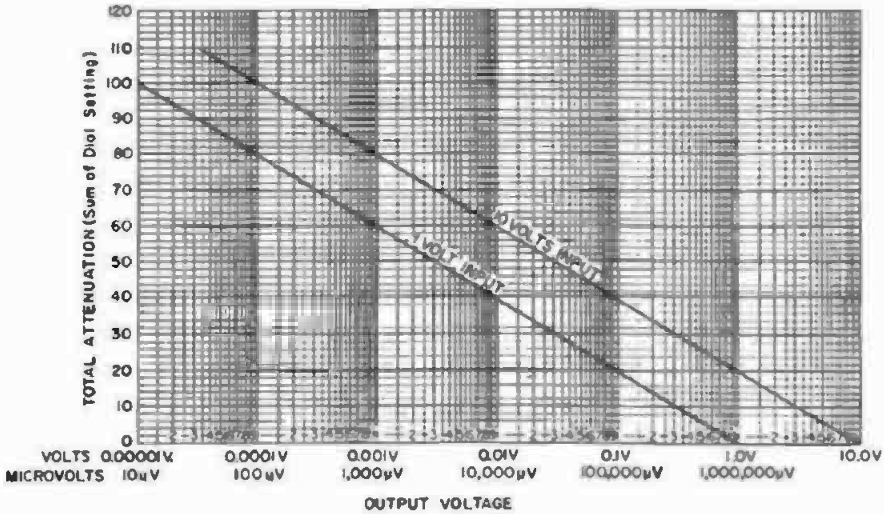


Fig. 23-54B. Voltage at the output terminals of an attenuator panel for a given voltage at the input.

may be set to indicate a known source of voltage at the input of the attenuator group. As an example, a level of plus 2 dBm is equal to 0.975 volt.

23.55 ***How is the insertion loss of a device measured using a gain set?***— The device to be measured is connected as shown in Fig. 23-55A. The send- and receive-meter attenuators are both set to the same level. All loss is removed from the gain-set attenuators. A signal is sent into the device and the reduced output level caused by the insertion loss of the device is read on the receive meter. If the insertion loss is beyond the range of the meters, the sending level is increased and the loss read on the lower figures of the receive meter. In either instance, the insertion loss is the difference in decibels between the two meter readings. As an example, if the send meter reads a plus 10 dBm and the receive meter reads a minus 10 dBm the insertion loss is 20 dB.

If the insertion loss is too great to be read by either of these methods, an

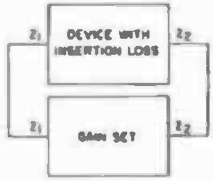


Fig. 23-55A. Measuring low insertion loss using a gain set.

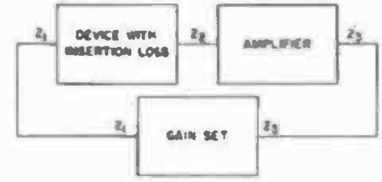


Fig. 23-55B. Measuring high insertion loss using a gain set.

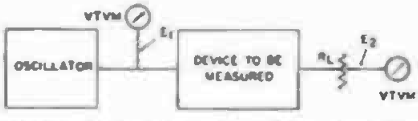


Fig. 23-55C. Measuring of insertion loss using two vacuum-tube voltmeters.

amplifier of known gain is connected in the circuit, as shown in Fig. 23-55B. The gain of the amplifier is measured first; then the device with the insertion loss is connected in the circuit, and the gain is measured again. The difference in gain for the two measurements is the insertion loss of the device under measurement. It is extremely important that all impedance matches be satisfied when making insertion-loss measurements; otherwise, the measurement will be in error.

Insertion loss may also be measured using one or two vacuum-tube voltmeters as shown in Fig. 23-55C. The insertion loss is the ratio of the input to output voltage:

$$\text{Insertion loss (dB)} = 20 \operatorname{Log}_{10} \frac{E_1}{E_2}$$

where,

E_1 is the voltage at the input,
E_2 is the voltage at the output.

When insertion loss is measured, the signal frequency must be such that it passes through the center of the passband of the device. As an example, if the insertion loss of a 400-Hz bandpass filter is to be measured, the oscillator signal is swept through the passband and set for the lowest loss (highest output). In the case of a high-pass filter the oscillator frequency must be well above the cutoff frequency. The same reasoning holds true for any device to be measured.

23.56 ***What type of resistors are most suitable for termination purposes?*** —The most convenient and economical terminating resistor is the wirewound vitreous noninductive type, since they can be obtained in almost any value of resistance and with up to several hundred watts of dissipative capacity. Although these resistors are called noninductive, some small part of inductive reactance is always present; however, at audio frequencies it may generally be ignored. The two cardinal points in the selection of such resistors are the wattage rating and the resistance value.

Terminating resistors used for precise measurements must be noninductive and within plus or minus 1 percent of the desired resistance. Before putting the resistance to use, two measurements are made of its resistance value: one at normal room temperature and a second after it has operated at half its wattage rating for 5 minutes. For example, a 50-watt, 16-ohm resistor was measured to be 15.80 ohms when cold. After operating at 25 watts for 5 minutes, its resistance dropped to 15.60 ohms. When such a resistance is used for a power measurement, the power dissipated in the resistance is computed on the basis of the hot resistance.

It is true that the small change in resistance will amount to only a fractional change in the output power. However, if the amplifier is being tested close to its overload point, the small increase in output power could cause it to overload and increase the distortion, particularly the intermodulation products.

To reduce the effects of heating in a resistor, it is good engineering practice to select a resistor that has a wattage rating at least 50 percent greater than the maximum power to be dissipated.

Regular wirewound vitreous resistors may also be used for terminations, although they are not entirely noninductive. The inductance, however, is still quite small, on the order of 2 to 25 microhenries for values of 10 to 600 ohms, with wattage ratings of 10 to 50 watts. Typical resistors of this type are illustrated in Fig. 5-87.

It is worthwhile to note a few precautions to be taken in the selection of resistors for terminating the input of an amplifier. As a rule, the resistance can be of the composition type, although in some instances such resistors can generate considerable internal noise, and when they are used with an amplifier system of high gain the generated noise appears in the output as amplifier noise. Also, some of these resistors appear to be of the composition type when they are actually wirewound and have considerable inductance. This latter type of resistor should be avoided unless it is known to be noninductive. Resistors used for input terminations are generally mounted in a shield, and the shield grounded to the amplifier chassis. Five-percent accuracy is acceptable for input terminations. (See Question 23.218.)

IHF Standard A-201-1966 for terminating resistors specifies that the resistor is to have not more than 10-percent reactive component at any frequency up to five times the highest frequency of interest. Also, the resistor is to be capable of dissipating the full output power of the device under test while maintaining its resistance within 1 percent of its rated value. (See Question 23.56.)

23.57 ***What does the term "double termination" mean, and what is its effect?*** — A double termination occurs when a device has been terminated in its normal load impedance and is terminated a second time, either accidentally or intentionally, as shown in Fig. 23-57.

A double termination connected at the output of an amplifier will cause the amplifier to overload and increase its distortion products. Also, a double termination will often cause the frequency response to be altered as well as reduce the power output.

For certain types of devices a double termination may be used intentionally to obtain a particular effect or characteristic. However, generally speaking, double terminations are to be avoided.

23.58 ***What is a turnover test?***—It is customary, before making a transmission measurement, to make a turnover test to determine whether any unbalance exists in the system to be tested or in the test setup. This is particularly important if the system contains a number of amplifiers and unbalanced or balanced circuits such as are to be found in a recording channel.

The test is made by sending a signal of 10,000 Hz through the system and noting the gain. The send circuit of the gain set is then reversed and the gain measured. If the gain increases or decreases by more than 0.5 dB, the system shows indication of an unbalance which should be eliminated, if possible, because the gain versus frequency characteristics are not the same for the two phases and, if a turnover in the connections between the units should occur, the frequency response will not be the same. Many times, unbalance in a circuit can be eliminated by connecting a repeat coil in the input of the device under test, although the input circuit does not normally require one. Although the following information mentions the use of a gain set, the basic rules are applicable to any gain-measuring circuit.

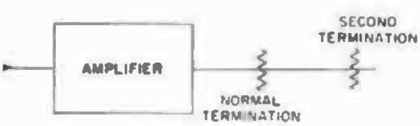


Fig. 23-57. Double termination applied to the output of an amplifier.

If no turnover is indicated when the input connections are reversed, invert the output connection to the gain-set receive section. If the output of the device being measured is designed to operate with one side grounded (such as negative-feedback amplifiers), it cannot be operated in the reverse direction. However, if the circuit is of a level in which a repeat coil can be used, a coil may clear the difficulty. Good transmission practice dictates that if the turnover is 0.5 dB or greater, it should be eliminated before proceeding with the measurement.

If the insertion of a repeat coil in the gain-set send circuit does not correct the difficulty, the turnover may be in the system under measurement. This can be determined by checking for improper grounding of a balanced or unbalanced circuit, or by grounding an unbalanced circuit, by checking the phasing of input and ouput circuits between the several different pieces of equipment in the system and by patching out certain units, such as an equalizer or filter.

Turnover is caused by the distributed capacity to ground of a circuit or piece of apparatus in the system. Applying 10,000 Hz to the system facilitates the measurement of unbalance because the reactance of the distributed capacity is lower than if measured at a frequency of, say, 100. As a rule, turnover cannot be detected at a frequency below 6000 Hz. Many times when an amplifier of high gain is measured, it will indicate considerable turnover. This is generally caused by unbalance between the measuring circuit and the input of the amplifier. The connection of a repeat coil in the input will generally clear up the difficulty.

If the gain-set receive section is grounded and the system being tested is also grounded, the ground should be removed from the gain-set receive section. The ground connection of the attenuator section (send) should never be removed for any measurement, because to do so would cause serious leakage in the attenuator section, resulting in erroneous measurements.

23.59 How is magnetic coupling measured in an amplifier? — Magnetic coupling in an amplifier is caused by the close proximity of the power transformer and filter chokes of the power supply to other components of the circuit. Also, magnetic coupling takes place due to magnetic lines of force induced in the metal chassis by the power transformer and chokes. These lines of force couple with the internal wiring and cause hum frequencies of 60 and 120 Hz to be induced into the signal frequencies.

A simple method of measuring the degree of magnetic coupling caused by magnetic lines of force in the chassis is to remove all tubes from the circuit, including the rectifier tube. Terminate the output winding of the output transformer in its normal load resistance. Connect a vacuum-tube voltmeter in parallel with the load resistor. Apply the ac voltage to the primary of the power transformer. The primary of the power transformer will now be energized along with the ac high voltage of the rectifier circuit. Also, the heater circuit will be energized.

Set the vacuum-tube voltmeter to a scale that will result in a convenient reading. (This will generally be on the order of minus 30 to 40 dBm.) Loosen the mounting screws of the power transformer and rotate it for a minimum reading on the vacuum-tube voltmeter. If more convenient, the power transformer may be left as is, and the output or interstage (if used) transformers may be rotated.

After noting the position of the transformers, install all the tubes and again measure the hum level at the output. Some revision of the transformer position may be necessary. In this latter measurement, both the magnetic coupling and other magnetic fields due to current are measured. As a rule, high-quality input and interstage transformers are well shielded and are designed to be rotated for minimum hum pickup.

The use of a single-point ground system will often eliminate many difficulties due to hum-frequency pickup. Such ground systems are discussed in Section 24.

23.60 What are the procedures for measuring the frequency characteristics of wave filters?—Referring to Fig. 23-60A, the input of the filter is fed from a resistive network of an impedance that matches the input impedance of the filter. The output of the filter is terminated in a resistive load paralleled with a vacuum-tube voltmeter. If the configuration of the filter is unbalanced, it must be grounded as shown. If the filter is of a balanced configuration, it must be grounded using a balanced resistive network as shown in Fig. 23-60B.

Frequencies of constant amplitude are applied from the oscillator to the input of the filter and the resulting frequency response is measured at the output of the filter across the load termination. The meter used across the output termination must have a rather large range of sensitivity to allow the measurement of frequencies beyond the cutoff frequency of the filter.

If two vacuum-tube voltmeters are used as shown, they must have identical frequency characteristics; otherwise,

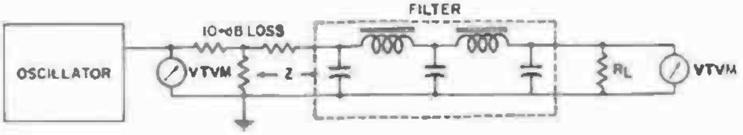


Fig. 23-60A. Circuit for measuring the frequency response of an unbalanced wave filter fed from a resistive network.

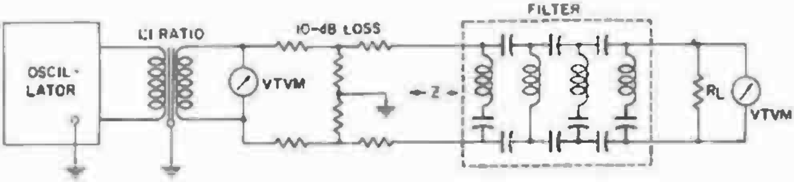


Fig. 23-60B. Circuit for measuring the frequency response of a balanced wave filter.

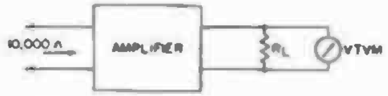


Fig. 23-60C. Amplifier used with voltmeter to increase the sensitivity.

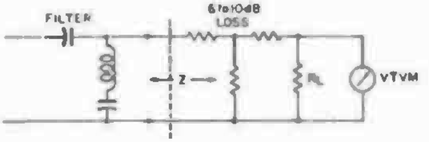


Fig. 23-60D. A 6- to 10-dB pad connected in the output of a filter to isolate the effects of the measuring circuit.

a true frequency response will not be possible. The subject of frequency discrimination in meters is discussed in Question 23.25. If there is any doubt as to the frequency response of the two meters, one meter may be used by switching it from input to output.

The input meter must be connected as shown and not across the input of the filter, because the reactive elements in the filter network will affect the meter readings, resulting in an erroneous frequency characteristic. Because of the characteristic of the iron cores used in filter coils, it is important that the filter be measured at a signal level commensurate with the manufacturer's specifications for the particular filter.

If the sensitivity of the output meter is not sufficient to read the levels beyond the cutoff frequency, an amplifier of known gain may be connected between the output of the filter and the input of the meter, as shown in Fig. 23-60C. The amplifier should have a bridging input impedance of 10,000 ohms or greater, with flat frequency characteristics. This will permit the filter to be terminated in a resistive load rather than an inductive one as when an amplifier of the same impedance as the filter is used. The circuits shown may be used to measure any type of filter configuration.

If the oscillator employs a grounded output, a repeat coil should be connected between it and the resistive network when measuring balanced filters. A ground may or may not be necessary with a balanced filter configuration. This can be determined by the turnover test described in Question 23.58. If the difference in turnover is greater than 0.20 dB, it is an indication of leakage, and the circuit should be grounded at the exact electrical center of the shunt resistor of the input network.

In some instances, it might be desirable to connect a 6- to 10-dB pad between the output of the filter and the input of an amplifier or vacuum-tube voltmeter as shown in Fig. 23-60D to isolate any effects of the amplifier or vacuum-tube voltmeter. For this type of measurement the terminating resistor is connected at the output of the pad. The attenuator resistance must match the output impedance of the filter. The measurement is made in the usual manner.

23.61 What is the procedure for measuring the frequency characteristics of a high-or low-pass filter by using a gain set?—The filter and gain set are connected as shown in Fig. 23-61. The output of the filter is terminated in a resistive termination and a bridging amplifier is connected in parallel with the termination. The bridging impedance of the amplifier should be at least 10,000 ohms and of flat frequency response. The gain of the amplifier is adjusted for a value that will permit the frequency characteristic beyond the cutoff frequency to be measured down to the noise level, which will be apparant when the measurement appears to flatten off.

If the amplifier input impedance matches the output impedance of the filter, an attenuator of at least 6 dB, and preferably 10 dB, should be connected between the output of the filter and the input of the amplifier to isolate inductive effects of the input transformer.

If the gain set employs a balanced configuration, a repeat coil will be necessary in the input, as shown in (c) of Fig. 23-14D. A ground is then connected to the low side of the filter.

If the filter has a high-pass configuration as shown in Fig. 23-61, the frequency run is started at the upper end of the passband. If it is a low-pass filter, the run is started at the low end of the passband. If it is necessary to increase the gain of the amplifier to measure the frequency response in the cutoff region, the increase of gain must be subtracted from, or added to, the measured loss.

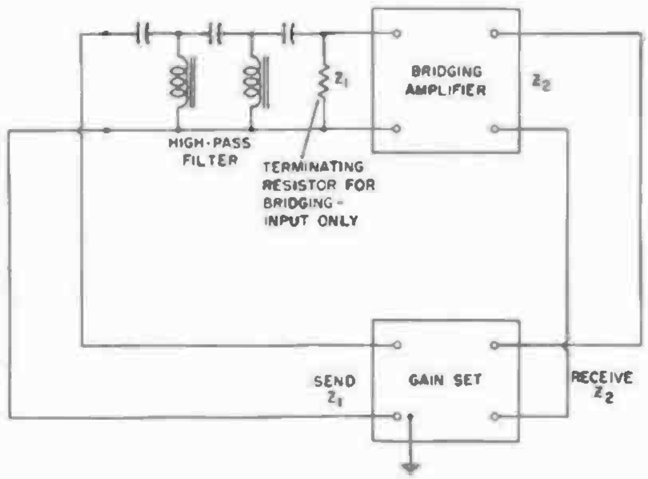


Fig. 23-61. Circuit for measuring the frequency characteristics of a high-pass filter, using a gain set.

Bandpass filters are measured in a similar manner, except that the measurement is started by finding the center frequency and then measuring first in one direction and then measuring in the other direction.

Band-elimination filters are started by finding the center frequency and then measured similarly to the bandpass filter.

23.62 ***How are the frequency characteristics of an equalizer measured by using a gain set?***—As shown in Fig. 23-62A. If the gain set has a balanced configuration, a repeat coil is necessary in the send circuit to isolate the balanced circuit of the gain set from the unbalanced equalizer circuit and, also, to permit the equalizer to be grounded.

As a rule, an amplifier is connected at the output of the equalizer network for the purpose of providing enough gain to permit a reasonable amount of loss to be inserted in the gain-set attenuators. If the amplifier has a bridging input, the ouput of the equalizer is terminated by a resistor.

The measurement is started by making a turnover test as described in Qnestion 23.58, by checking the maximum rise (or dip) of the equalizer, and then by setting the gain of the amplifier. If the equalizer has a rise, the gain of the amplifier is set at the peak of the rise and the measurement made. This ensures that the amplifier will not

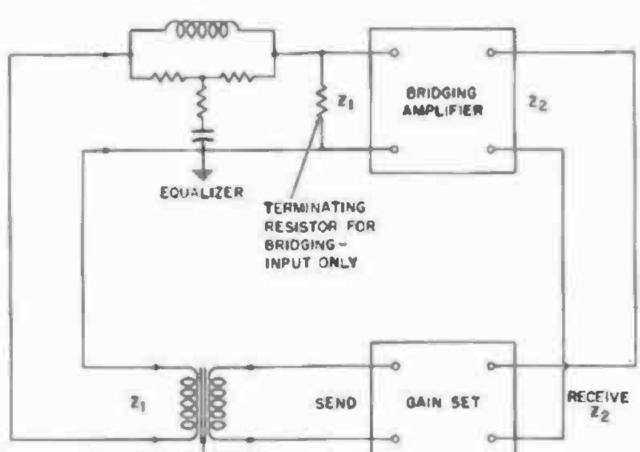


Fig. 23-62A. Circuit for measuring the frequency response of an equalizer, using a gain set.

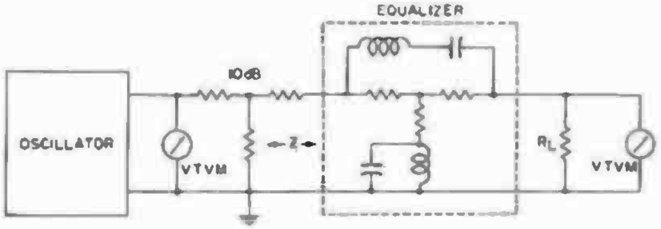


Fig. 23-62B. Circuit for measuring the frequency characteristics of an unbalanced equalizer, using an isolating pad and vacuum-tube voltmeters.

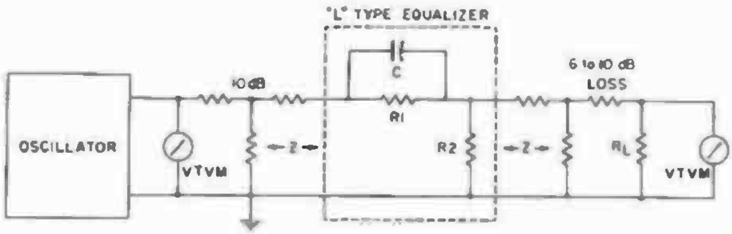


Fig. 23-62C. Circuit for measuring the frequency response of an "L" type equalizer. A 6- to 10-dB pad is connected in the output to supply the proper termination.

be overloaded as the rising characteristic of the equalizer is aproached.

As an example, if the equalizer has a rise of 10 dB at its resonant frequency, the gain of the amplifier is adjusted for a given amount of attenuation loss in the gain set. The gain control is left at this particular setting and is not changed throughout the rest of the measurement.

It will be noted that other frequencies will be lower in amplitude by the amount of the rise (10 dB). Therefore, the amplifier cannot be overloaded by the rising response of the equalizer. The response is then plotted with respect to either 400 or 1000 Hz, depending on the purpose for which the equalizer is designed.

If a gain set is not available, the frequency response may be measured by using one of the two circuits shown in Figs. 23-62B and C. The procedure and precautions are the same as described for the measurement of a filter in Question 23.60.

23.63 ***How are the frequency characteristics of wave filters and equalizers plotted?***—As shown in Figs. 23-63A and B. For wave filters the reference frequency is generally 1000 Hz. For equalizers either 400 or 1000 Hz is employed, depending on the use of the equalizer.

Filters or equalizers may be plotted to show their insertion loss or their transmission characteristics as shown in Fig. 6-29A and B. Generally the transmission characteristic is used because it pictures the action of the device in a transmission circuit. Insertion-loss plots are used in design procedures.

As a rule, the characteristics of equalizers and wave filters are plotted using the major lines on the graph paper to represent 5 dB and each minor line representing 1 dB. In special cases where it is necessary to expand a certain area, the plot of each major division can be used to represent 1 dB and the minor lines can represent 0.2 dB.

Equalizer and filter characteristics are discussed in detail in Sections 6 and 7, respectively.

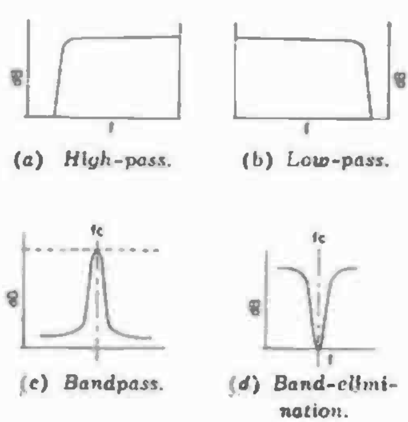
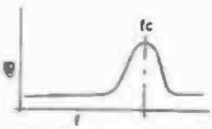
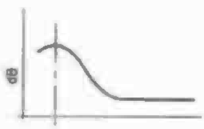


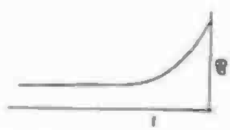
Fig. 23-63A. Frequency-response characteristics of wave filters.



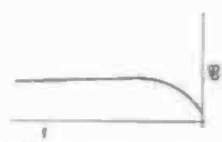
(a) *High-frequency peaking.*



(b) *Low-frequency boost.*



(c) *High-frequency pre-equalizer.*



(d) *High-frequency post-equalizer.*

Fig. 23-63B. Frequency-response characteristics of equalizers.

23.64 ***Describe the procedure for measuring the characteristics of a speaker crossover network.*** — Crossover networks employing air-core coils (as most are) may be measured as given in Fig. 23-64A. Since the coils are air core, they may be measured at a low level because the characteristics of such coils are not affected by the power level. For this example of measurement it will be assumed that a 450-Hz parallel-type crossover network of 16-ohm impedance is to be measured. An impedance-matching network of 600/16 ohms is connected between the oscillator and the input of the network. The output terminals of the network are terminated in 16 ohms, with a vacuum-tube voltmeter connected in parallel. It will be observed that the input voltmeter is connected on the high side of the network to prevent reflections from the active elements of the network from affecting its readings.

The measurement is started with the low-pass section. The oscillator is set to a frequency of 30 Hz as a reference frequency. The output meter readings are noted and tabulated. The high-frequency side is now measured by setting the oscillator to 10,000 Hz as a reference frequency. Frequencies down to 3 octaves below the crossover frequency are applied and their amplitudes noted. The results of the measurements are then plotted as given in Fig. 23-64B. For a properly designed network the crossover frequency will be down 3 dB from the reference frequency.

Series-type crossover networks are measured as given in Fig. 23-64C. The same measurement procedure is followed as is used for parallel networks.

Insertion loss may be measured by noting the voltage at the input and output of a given section, at a frequency somewhat near the center of the passband. The loss is then computed:

$$\text{dB} = 20 \operatorname{Log}_{10} (E_1/E_2)$$

$$\text{Power} = IL = E^2/R_L$$

where,

- E_1 is the voltage at the input of the network,
- E_2 is the voltage across the output termination of the section under measurement,
- R_L is the terminating resistance.

The characteristics of crossover networks can also be measured by using the amplifier normally employed for driving the speaker system. For such measurements the test circuit of Fig. 23-64D is used. Some difference in the

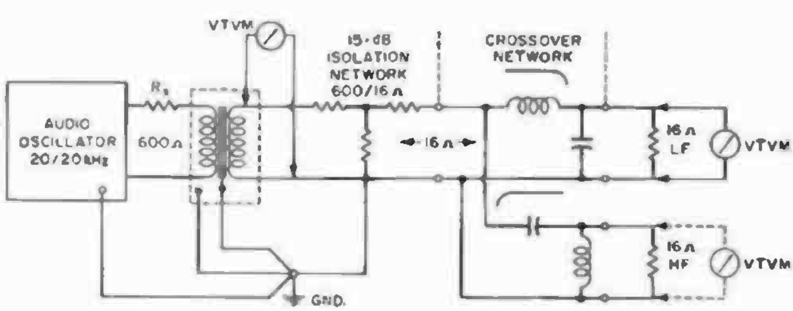


Fig. 23-64A. Test setup for measuring characteristics of a parallel crossover network.

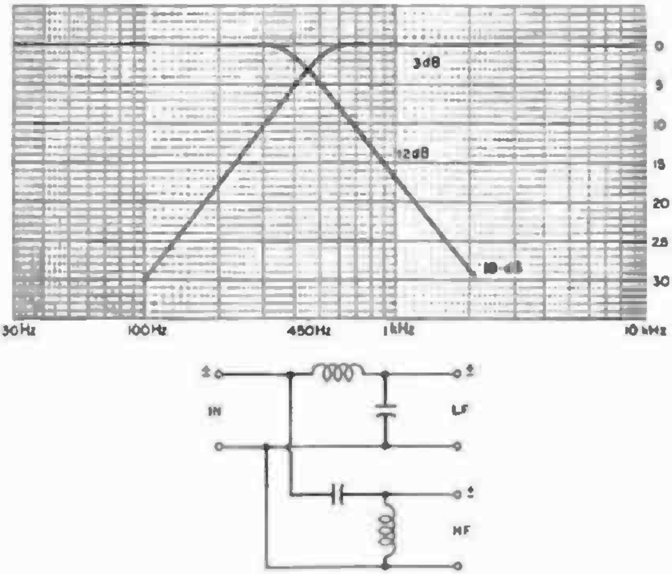


Fig. 23-64B. Frequency response of a parallel crossover network.

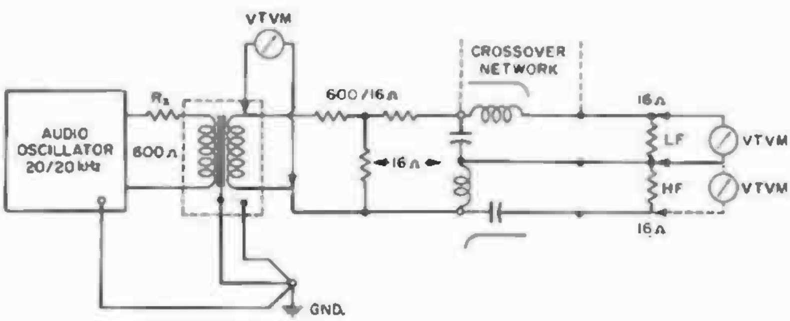


Fig. 23-64C. Test setup for measuring characteristics of a series crossover network.

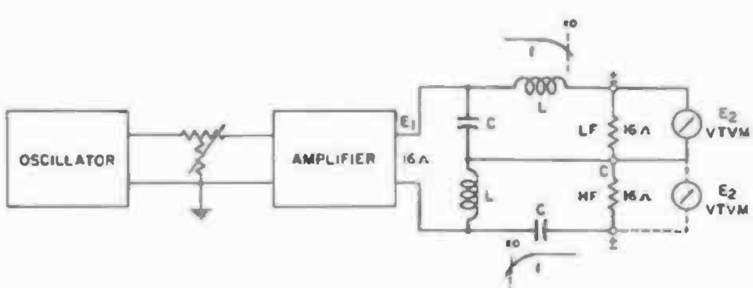


Fig. 23-64D. Test circuit for measuring the characteristics of a crossover network, using its normal driving amplifier.

characteristics of the network may be expected when measured in this manner, since the internal ouput impedance of the amplifier is not a perfect match to the crossover network input, as is the impedance-matching network. The first measurement is an actual measurement of the network characteristics, while the measurement with the driving amplifier is a practical or operational measurement. (See Question 20.90.)

Networks employing more than two sections are measured in a similar manner (Fig. 23-64E), taking care that all unused sections of the network are properly terminated and that the mea-

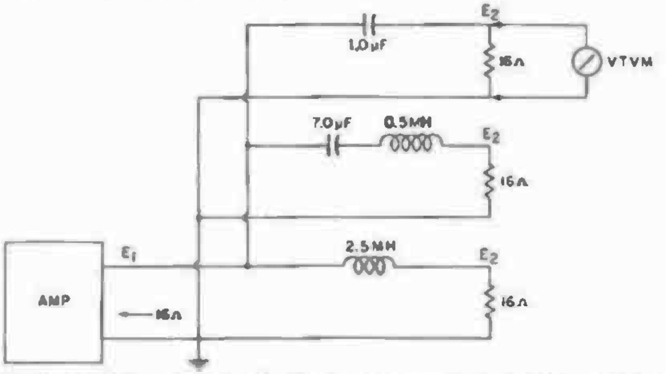


Fig. 23-64E. Method of measuring the frequency response of a three-section, 16-ohm crossover network.

surement is made in a portion of the passband removed from the crossover frequency. The insertion loss variation between sections should be very small. This will be indicated by the difference between the input voltage and the output voltage of the various sections in the passbands. The design of crossover networks is discussed in Section 7.

23.65 ***How are the overall frequency characteristics of a recording channel plotted?*** — The frequency characteristic is measured from the input of the microphone preamplifier to the output of the amplifier driving the recorder. A plot of a typical magnetic recording channel with the high- and low-pass filters removed is given in Fig. 23-65A. Fig. 23-65B shows the same channel using high- and low-pass filters.

In addition to the overall frequency response, it is customary to make frequency measurements of each individual piece of equipment in the channel, to facilitate the servicing of individual units. These plots should include, besides the frequency response, the gain for a given set of conditions, the maximum power output, the harmonic or intermodulation distortion, the signal-to-noise ratio, the equalization, the compression ratio and breakaway

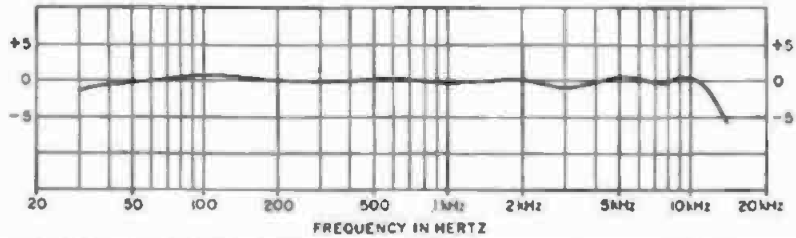


Fig. 23-65A. Method of plotting the overall frequency response of a magnetic recording channel.

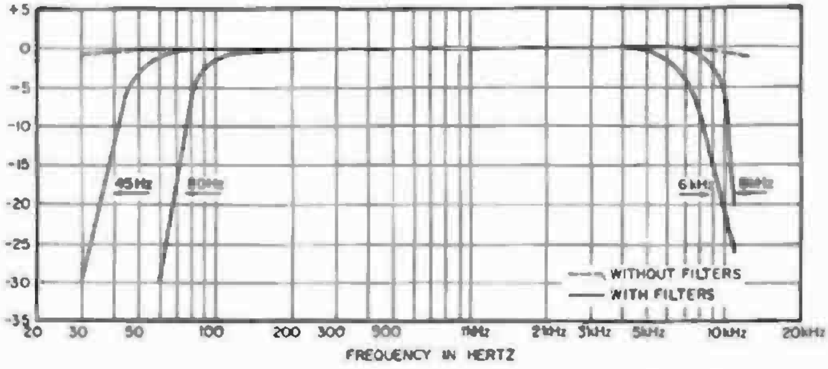


Fig. 23-65B. Frequency characteristics of a typical recording channel using high- and low-pass filters.

points, and any other measurements that may be used for reference. A standard method of plotting should be adopted for the preceding characteristics, so that the various graphs can be laid over each other for comparison.

One of the major lines (counting up from the bottom) of semilogarithmic graph paper such as is used throughout this book is referred to as the zero line. Each major division represents 5 dB and each minor line 1 dB. The frequency range of the paper is 20 to 20,000 Hz. The usual plotting range for a motion-picture recording channel is 40 to 10,000 Hz.

23.66 How are the frequency characteristics of a mixer network measured?—The mixer is connected as shown in Fig. 23-66. If the gain set is the balanced type, a repeat coil is connected in the send circuit and the mixer network is grounded as shown. An amplifier is connected at the output of the network and returned to the gain-set receive terminals. If the amplifier has a bridging input, the output of the network is terminated in its normal operating impedance by means of a resistor.

The mixer input control used for the measurement is set for about 15 to 20 dB of loss. This is about the normal operating position for the average mixer. All other controls are set to their maximum loss positions or off positions. The gain of the amplifier is set for a convenient measuring level at the gain-set receive terminals. Usually 30 dB of gain will suffice.

After the measurements of the network characteristics have been made, the response of any equalizers or high- or low-frequency attenuators is then made.

For motion-picture production and rerecording, the frequency response is generally plus or minus 1 dB from 40 to 10,000 Hz. For radio broadcasting, it is 30 to 16,000 Hz, plus or minus 1 dB.

23.67 How is the leakage between positions of a mixer network measured?—The mixer network is connected as shown in Fig. 23-67. Suppose the leakage of a four-position mixer is to be measured at a frequency of 10,000 Hz. Positions 2 and 3 are terminated in their normal input impedances using a resistive termination. Position 4 is also terminated and a vacuum-tube voltmeter connected in parallel with the termination. The attenuators of positions 2, 3, and 4 are set to their maximum loss or the off position. An audio oscillator is connected to position 1, and the attenuator of that position is set to approximately 20 dB of loss (this is about the loss that would be used in the average mixer pot). The output of

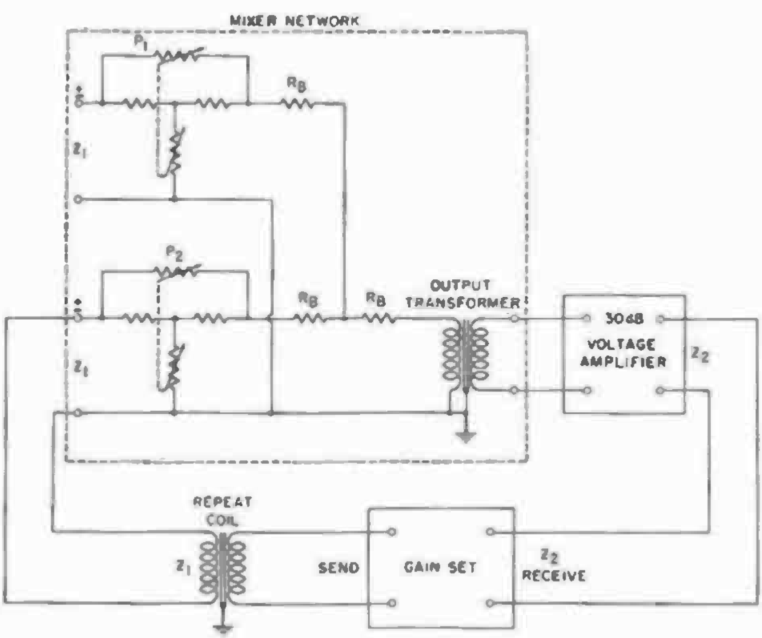


Fig. 23-66. Circuit for measuring the frequency response of a mixer network.

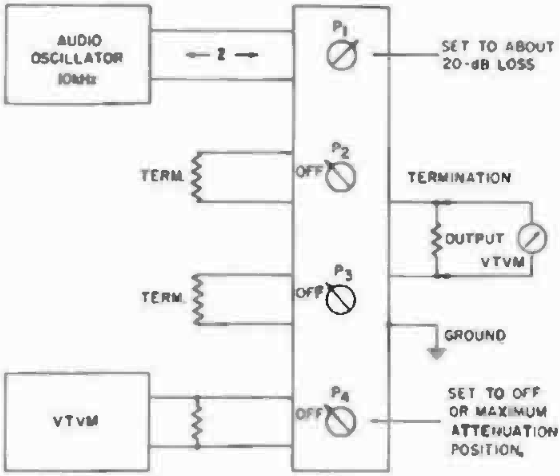


Fig. 23-67. Circuit for measuring the leakage between the various positions of a mixer network.

the oscillator is set for a level comparable to the highest signal level that will be used under normal operating conditions. For motion-picture rerecording, this will range from minus 24 to minus 10 dBm. For normal microphone pickups using a conventional preamplifier, the level will range from minus 25 to minus 12 dBm.

With the test signal applied to input 1, the leakage is read on the vacuum-tube voltmeter across position 4, with that attenuator set to maximum loss. For a well-designed mixer, the leakage will be at least 70 dB or greater. If the leakage is less than 70 dB, it is generally caused by multiple grounds, ground loops, or inadequate shielding between the pairs connecting the input and output circuits of the mixer controls.

The lowest leakage is obtained by the use of a single-point ground system of wiring as described in Question 9.37. It is essential, when a leakage measurement is made, that the mixer network be grounded to the transmission ground of the measuring circuit.

The leakage between the output circuit and a given control is measured in a similar manner, except the fourth control is terminated and the vacuum-tube voltmeter is connected across the output terminating resistor. The oscillator is set for a level comparable to that normally applied to the input circuits and is connected to each input (with the control closed and the other controls in the off position), and the leakage is measured at the output.

The attenuation of the various controls may be measured by setting them to various positions of loss and noting the reduction in level at the output. (See Question 23.66.)

23.68 How is the insertion loss of a mixer network measured?—The mixer network is connected as shown in Fig. 23-68. Mixer networks have two losses: one variable and the other fixed. The variable loss is dependent on the setting

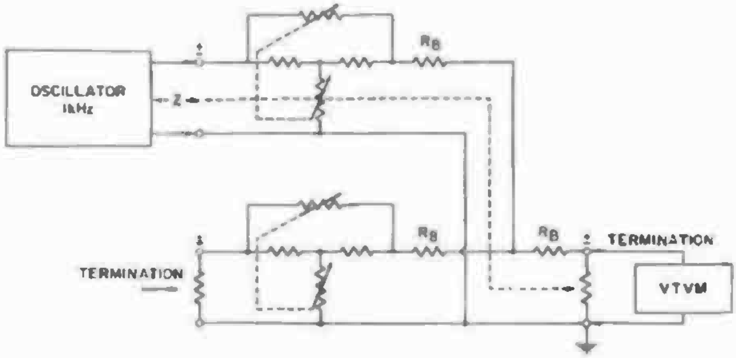


Fig. 23-68. Method of measuring the insertion loss of a mixer network.

of the mixer control; the fixed loss, the value of the building-out resistors, and the type of controls used.

The fixed insertion loss is measured between a given input and the output of the mixer network. The difference between the input signal level and the output level is the insertion loss of the network. The measurement is made with the mixer control of the input under measurement set to its minimum-loss position (wide open) and the controls of all other positions set to their maximum loss or off positions.

The only loss measured under these conditions will be the loss induced by network components R_a (building-out resistors). However, this statement is true only for networks using plain-T or bridged-T attenuators. For networks using ladder potentiometers, the measured loss will be 6 dB greater (the insertion loss of a ladder potentiometer) than that measured with a mixer of the same number of positions using T or bridged-T potentiometers.

As an example, a six-position, 250-ohm network using bridged-T attenuators has an insertion loss of 15.6 dB. The same network using ladder potentiometers will have a loss of 21.6 dB. The path of the test signal is shown by the dotted lines in the diagram. Insertion losses for configurations of different impedances and numbers of positions are given in Fig. 9-44.

The insertion loss of any mixer network may be calculated:

$$\text{dB loss} = 20 \operatorname{Log}_{10} N$$

where,

N is the number of input positions.

If the network uses ladder pots, 6 dB is added to the calculated insertion loss.

23.69 ***How are the frequency characteristics of an audio transformer mea-***

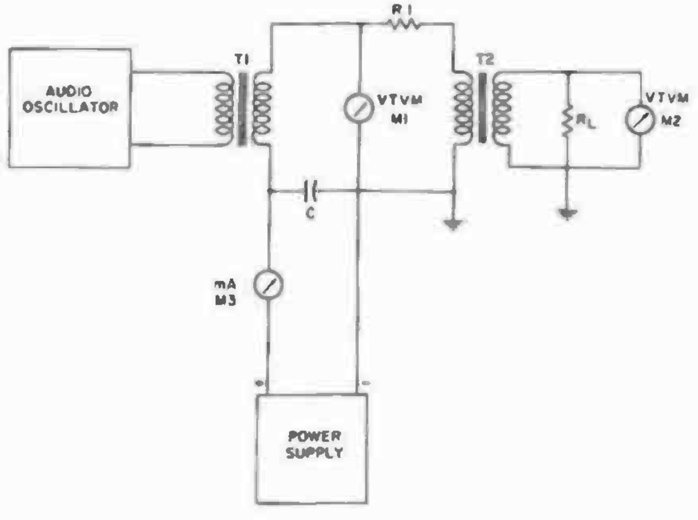


Fig. 23-69A. Circuit for measuring the frequency characteristics of a single-ended audio transformer.

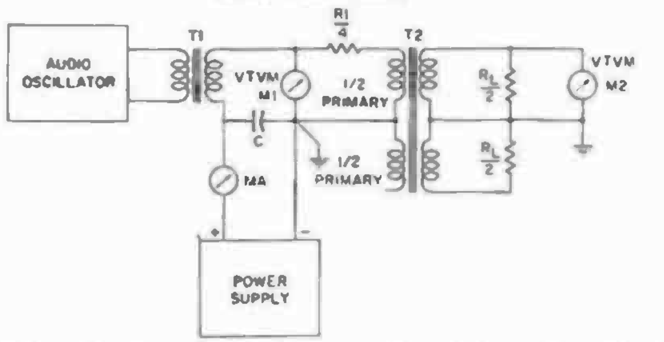


Fig. 23-69B. Circuit for measuring the frequency characteristics of a push-pull interstage transformer.

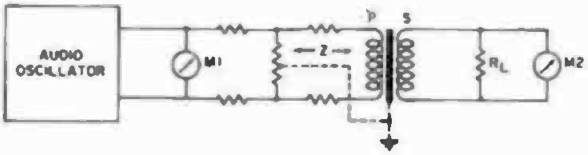


Fig. 23-69C. Method of measuring the frequency characteristics of a repeat or isolation coil.

sured?—As shown in Fig. 23-69A for single-ended transformers, and as shown in Fig. 23-69B for push-pull transformers. R1 is a noninductive resistor equal in value to the plate impedance that the transformer is designed to match. R_L is the secondary load impedance, and C is a large capacitor with a reactance sufficient to pass the lowest frequency to be measured. A dc power supply applies a current equal to the normal plate current that will be passed by the primary of the transformer.

The actnal measurement is made by applying frequencies of a constant amplitude to the primary of transformer T2. This amplitude is read on vacuum-tube voltmeter M1. The frequency response is read on vacuum-tube voltmeter M2. If two meters are used as shown, they must have similar frequency characteristics for the reasons explained in Question 23.25. Transformer T1 has a low-impedance secondary on the order of 30 to 50 ohms.

When measuring a push-pull interstage transformer, only one-half of the secondary winding is measured at a time. This will avoid shunting the total secondary winding with the input capacitance of meter M2, which may affect the response at the higher frequencies.

Output transformers may be measured in a similar manner; however, it is generally better to measure such transformers in an amplifier with characteristics similar to those in which it is to be used.

Fig. 23-69C shows the method used for measuring the frequency response of a repeat or isolation coil. The output of the oscillator at the left is matched to the input impedance of the coil by means of a balanced or unbalanced attenuator network. The secondary side is terminated in a resistive load, R_L.

Frequencies of constant amplitude are fed into the primary side, and the response is measured by a vacuum-tube voltmeter connected across load resistor R_L.

The insertion loss is measured by noting the voltage across the primary and the secondary for a given frequency (generally 1000 Hz). The loss in decibels is then 20 Log_{10} of the ratio of the primary to the secondary voltage.

23.70 How may the frequency response of a transformer be calculated at the lower frequencies if the primary reactance is known?—The equation given below is based on the amount of low-frequency voltage that will appear across the primary turns of a transformer when it is connected in series with a resistance equivalent to the rated primary impedance. The expression for each frequency is:

$$\text{Insertion loss in dB} = 20\,\text{Log}_{10}\frac{1}{\sqrt{1 + (R_1/\omega L_p)^2}}$$

where,

L_p is the primary inductance in henries at the selected frequency,
ω equals $2\pi f$,
f is the frequency in hertz,
R_1 is the rated primary impedance with the following correction.

For transformers snch as an input or interstage working into an open circuit (no secondary termination except for the input capacitance of the tube):

$$R_1 = R_3 = R_r + R_p$$

where,

R_p is the dc resistance of the primary winding,
R_r is the rated primary impedance.

For transformers loaded on the secondary side, such as a driver or output transformer:

$$R_1 = \frac{R_3}{1 + R_3/R_7}$$

$$R_3 = (R_r + R_p)$$

$$R_7 = (R_s + R_L)\left(\frac{N_p}{N_s}\right)^2$$

where,

R_s is the dc resistance of the secondary,

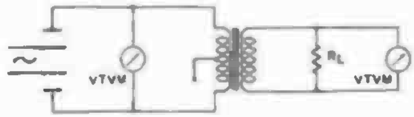


Fig. 23-71A. Method of measuring the insertion loss of an output transformer.

R_s is the secondary load resistance,
N_p is the turns in the primary,
N_s is the tnrns in the secondary winding.

23.71 ***How may the insertion loss of an output transformer be measured?***—As shown in Fig. 23-71A. A signal is applied to the output stage with the transformer secondary terminated in its normal load impedance. If the amplifier employs negative feedback, the feedback loop is disconnected. The voltage is measured across the total primary winding, and the power is computed:

$$P = \frac{E^2}{Z}$$

where,
Z is the plate impedance.

(For a push-pull stage, the voltage is measured from plate to plate.)

The power in the secondary is computed in a similar manner. The ratio of the two powers may then be converted to decibels:

$$dB = 10 \text{ Log}_{10} \frac{P_1}{P_2}$$

where,
P_1 is the power at the primary,
P_2 is the power at the secondary.

Fig. 23-71B shows the insertion-loss curves for a conventional output transformer, and Fig. 23-71C shows those for an ultralinear transformer used in a Williamson-type amplifier.

The foregoing insertion-loss curves should not be confused with a power-versus-frequency response curve. Such curves are run with the negative-feedback loop connected.

23.72 ***What is the procedure for balancing a hybrid coil?***—Referring to Fig. 23-72, the secondary (Z_s) and the two primary windings (Z_1 and Z_2) are each terminated with a resistor equal to the recommended impedance. A vacuum-tube voltmeter is connected across the terminating resistor of primary Z_2, and a signal of 1000 Hz is applied across primary Z_1. Resistor Z_s is made variable during the test and for the average coil will be about 250 ohms. Later, this resistor is replaced with a fixed resistor.

The signal level from the oscillator at Z_1 may be set at a level of 0 dBm or at the normal operating level. The vacuum-tube voltmeter is set for a convenient reading, and variable resistor Z_s is adjusted for a minimum reading on the meter across Z_2. The meter sensitivity is adjusted as the resistor is adjusted. The opposite side (although balanced) may be checked for balance by reversing the procedure. A good coil will generally show about 45 dB of isolation at frequencies between 30 and 10,000 Hz. The value of resistor Z_s is

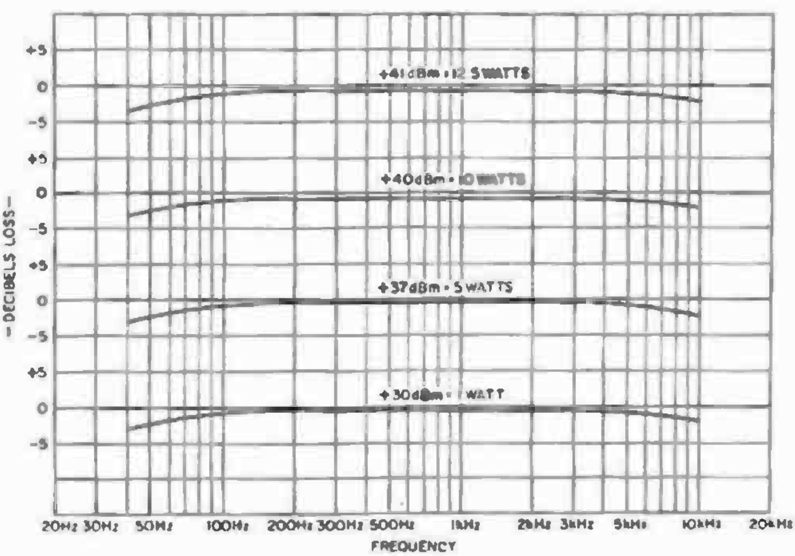


Fig. 23-71B. Insertion loss in dB for conventional push-pull output transformer.

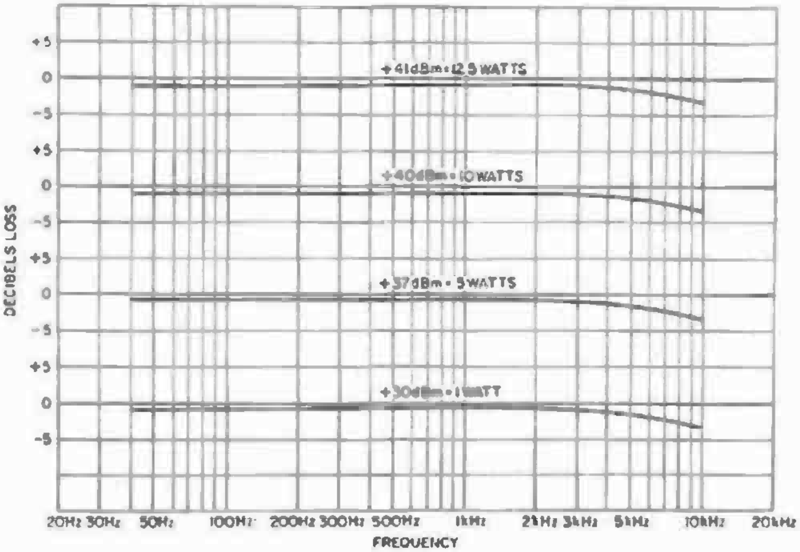


Fig. 23-71C. Insertion loss in dB for ultralinear output transformer.

generally quite critical and is only effective when the terminating impedance across the primaries is constant. If a meter is bridged across resistor Z_3, the signals in windings Z_1 and Z_2 may be monitored simultaneously.

The degree of isolation between windings Z_1 and Z_2 is stated in reference to 1 milliwatt or 0 dBm; however, it may be stated with reference to the signal level at the opposite primary winding. A typical coil of this type is the LS-141 hybrid coil manufactured by the United Transformer Co. The losses encountered in a hybrid coil and other data are discussed in Questions 8.66 to 8.71.

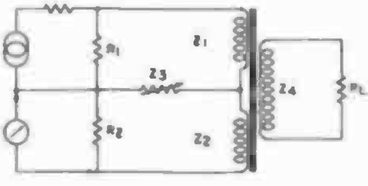


Fig. 23-72. Circuit for balancing a hybrid coil.

23.73 ***How are the frequency characteristics of a magnetic recorder measured?***—As shown in Fig. 23-73. The signal level from the oscillator is adjusted for a recording level at least 10 dB (preferably 20 dB) below the normal recording level, as indicated by the level-indicating device of the recorder. Making the frequency-response measurement in this manner prevents overloading of the magnetic tape.

If possible, a 20,000-Hz low-pass filter, similar to that described in Question 7.107 should be connected in the output of the playback amplifier to remove the effects of the high-frequency bias current on the meter employed to read the frequency response. The filter is used only when frequency and noise measurements are made. It should be removed when a distortion measurement is made. The frequency response of the filter is such that frequencies up to 18,000 Hz are passed without discrimination.

If the machine is equipped with a separate playback head, the frequency

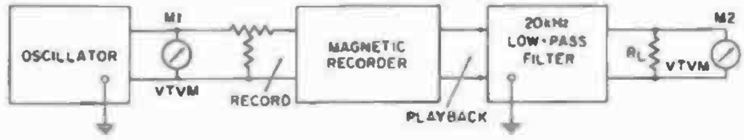


Fig. 23-73. Circuit for measuring the frequency response of a magnetic recorder. The 20-kHz low-pass filter is used to remove the effect of the high-frequency bias current on the meter M2.

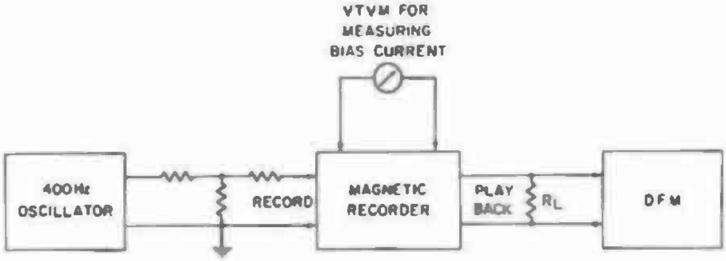


Fig. 23-74A. Circuit connection for making distortion and bias-current measurements on magnetic equipment.

response may be observed as the oscillator frequencies are applied. If not, the tape is recorded and then played back. Before a frequency measurement is attempted, the heads should be aligned as described in Questions 17.104, 17.109, 17.110, and 17.111. A frequency-response measurement made in this manner is called a record/playback frequency response and includes the frequency response of the record and playback amplifiers.

In some recorders the record and playback equalization can be individually adjusted for the best response. If such controls are available, they should be adjusted to the manufacturer's specifications; if not, serious overloading of the amplifiers may result.

It is advisable to demagnetize both the recording and reproducing heads before making any adjustments, because a magnetized head will increase the distortion, reduce the signal-to-noise ratio, and may partially erase the high frequencies. (See Question 7.107.)

23.74 Describe the procedure for making distortion measurements on magnetic recording and reproducing equipment.—The equipment is connected as shown in Fig. 23-74A. The principal reason for making distortion measurements on magnetic recording equipment is to establish the correct operating bias current, the maximum operating levels, and the signal-to-noise ratio for a given operating level.

Both the motion picture industry and manufacturers of professional magnetic recording equipment have established 1 percent as maximum total harmonic distortion (THD) for recording and reproduction. The 1-percent THD represents 100-percent sine-wave modulation of the system. In earlier equipment a 3-percent THD was used and in some instances the distortion was stated in terms of intermodulation distortion. However, to date no standards have been established for intermodulation testing of magnetic recording equipment. The following tests to be described are applicable to both magnetic film and tape.

Harmonic-distortion measurements are made by applying a signal from a low-distortion oscillator to the recording channel and then measuring the amount of harmonic distortion at the output of the playback amplifier. Distortion measurements should be made at several levels above and below the normally specified recording level for several different values of bias current. These tracks are then played back and the distortion and signal-to-noise ratio

Bias Current in mA	*Percent Distortion*		
	+8	*+10*	*+12*
20	1.0	1.6	2.55
22	0.9	1.5	2.4
24	0.7	1.35	2.2
26	0.6	1.2	2.0
28	0.55	1.1	1.75
30	0.50	1.0	1.5
S/N	−59 dB	−61.5 dB	−62 dB

Fig. 23-74B. Tabulation of percent distortion. Bias current versus recording level (in dBm).

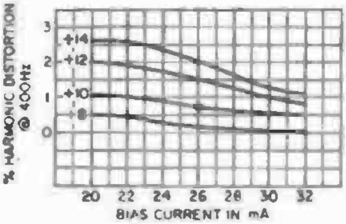


Fig. 23-74C. Harmonic distortion versus bias current for magnetic film recorder (30 to 12,000 Hz).

measured. The combination of bias current and recording level resulting in the greatest signal-to-noise ratio with the lowest distortion and best frequency response is the correct recording level. (See Questions 17.48, 17.49, and 17.52.)

The procedure for making the measurements is as follows. Assume a magnetic film recorder is specified to operate at a recording level of plus 12 dBm at the recording head with a bias current of 25 mA. A vacuum-tube voltmeter is connected at the output of the recording amplifier (or where specified by the manufacturer), and sound tracks are recorded (at 400 Hz) at levels of plus 8, 10, 12, and 14 dBm with bias currents of 20, 22, 24, 26, 28, 30, and 32 mA. A few feet of unmodulated sound track is left at the end of each recording to permit the signal-to-noise ratio to be measured. The tracks are played back, and the harmonic distortion and signal-to-noise ratios are measured. These are then tabulated as in Fig. 23-74B, and, when plotted, will appear

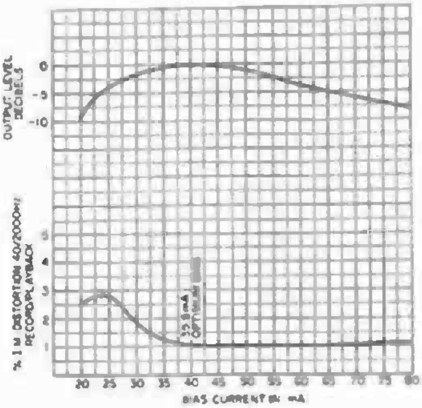


Fig. 23-74E. Record-playback response of ¼" magnetic tape recorder at 40 and 2000 Hz and tape speed of 7½ ips. Plotted as bias current versus intermodulation distortion and output level.

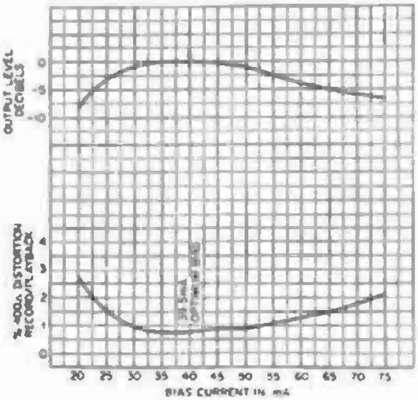


Fig. 23-74D. Record-playback response of ¼" magnetic tape recorder at 400 Hz and tape speed of 7½ ips. Plotted as bias current versus harmonic distortion and output level.

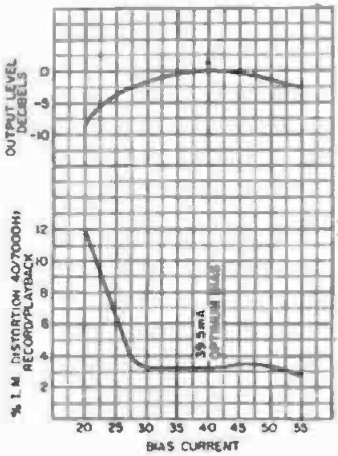


Fig. 23-74F. Record-playback response of ¼" magnetic tape recorder at 40 and 7000 Hz and tape speed of 7½ ips. Plotted as bias current versus intermodulation distortion and output level.

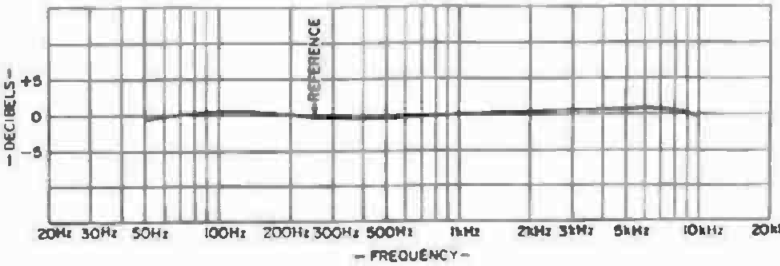


Fig. 23-74G. Frequency response of ¼" magnetic tape recorder.

as a family of bias curves such as those shown in Fig. 23-74C.

After selecting a set of conditions that appears to have the best signal-to-noise ratio with the lowest distortion, a frequency-response measurement is made using the selected bias current and recording level. If the value of bias current is too high, it may erase the high frequencies as they are being recorded. Therefore, it is important that the frequency response be measured before settling for a given set of conditions.

As a rule, a compromise is generally made. If the signal-to-noise ratio is 60 dB for a given set of conditions and the frequency response is better for a signal-to-noise ratio of 58 dB, the 2-dB difference in noise level would be negligible if the frequency response were materially improved. It is assumed that before the foregoing group of measurements is made, the head alignment will have been checked and the heads thoroughly degaussed. (See Question 17.91.)

Although the value of 1-percent THD maximum is specified, operating the equipment at a slight amount above 1 percent will have little effect. Also, if the signal-to-noise ratio will permit, operating the equipment below a 1-percent THD is permissible. In the family of bias curves, Fig. 23-74C, the point of operation has been selected for a distortion value of 1.20 percent, with a signal-to-noise ratio of 61.5 dB.

It will be noted that for a recording level of plus 10 dBm, using a bias current of 28 to 30 mA, the distortion is approximately 1 percent. This would be a good point at which to operate if the frequency response is satisfactory. However, with a bias current of 28 to 30 mA, some difficulty may be experienced with erasure at the higher frequencies. If difficulty is experienced with the high-frequency bias current leaking into the output and affecting the measurements, a capacitor of 0.25 μF may be connected across the output of the playback circuit, if the output is of 600-ohms impedance.

Fig. 23-74D depicts the record/playback harmonic distortion measured on a professional ¼-inch magnetic-tape recorder to show the total rms harmonic distortion for different values of bias current. The bias current was measured by connecting a 1-ohm resistor in series with the recording head and a vacuum-tube voltmeter in parallel with the resistor. The results were plotted as harmonic distortion versus bias current. The upper curve is the variation in output level due to changes in bias current.

Fig. 23-74E shows a similar group of measurements obtained by the use of an intermodulation analyzer. The frequencies used for the intermodulation measurements were 40 and 2000 Hz mixed in a ratio of 4:1 (low frequency 12 dB higher in amplitude than the high frequency).

Fig. 23-74F is the same recorder using 40 and 7000 Hz for a similar group of measurements. A record/playback frequency response is shown in Fig. 23-74G.

One of the attractive features of magnetic recording is that it is inherently free of even-harmonic distortion. The magnetic characteristic is not necessarily linear but is symmetrical for opposite directions of magnetization. In practice, however, wave analysis of the playback of a sine-wave recorded from a distortion-free source will frequently show appreciable second and higher even-order harmonic distortion. When this situation is encountered, it is usually indicative of some malfunctioning of the recording equipment.

If the amplifier system when measured by itself shows negligible distortion, it may be assumed that the observed distortion from the tape is being caused by the recording process itself.

Distortion in the recording process is caused by a direct-current component or magnetization which prevents the recording heads from modulating the tape around a point of symmetry. The asymmetric influences may be clearly visualized at the recording head; but if the erase head leaves the tape in a magnetized condition the same result can occur. High harmonic distortion can, as a rule, be traced to the erase, recording, or reproducing heads magnetized by coming in contact with magnetized tools, or to excessive transient currents generated in the switching circuits. Magnetization of the heads can also occur from leaky coupling capacitors, permitting direct current to flow through the head in question.

If no direct current is present, then the observed distortion may be caused by high distortion of the bias oscillator waveform or a badly overloaded recording amplifier. External stray dc fields will also have the same effect as a direct current through the head circuit.

To test for the existence of an asymmetrical field, a small permanent magnet is held near the recording head. If a given position of the magnet causes a reduction of the even-harmonic distortion, the polarity of the magnet is noted. The polarity is reversed, and the distortion measured again. If the distortion value is increased or decreased, second-harmonic distortion is indicated. The heads should be demagnetized by using a degausser as described in Question 17.90.

It will be noted that when the source of distortion is removed, the signal-to-noise ratio will be increased.

23.75 ***Describe the procedures for recording disc-record light patterns and their measurement.*** — The making of light patterns for plotting the frequency characteristics of cutting heads and disc-recording channels was first devised by Buchman and Meyer. Light patterns are recorded by applying frequencies of constant amplitude to the recording channel and then recording them on a disc record. Each frequency is recorded for approximately 10 seconds with an unmodulated groove of 5 seconds between each frequency. The recording is started with the highest frequency on the outside of the record to reduce the effect of lower groove velocities at the smaller diameters. A typical light pattern made with a magnetic cutting head using a modified constant-amplitude, constant-velocity recording characteristic (see Question 14.6) is shown in Fig. 23-75A.

The usual manner of reading a light pattern is to view it in sunlight or by means of a small light located some distance away from the disc so that the light rays strike the disc nearly parallel to its surface. The pattern is then observed from a distance of about 4 feet, using one eye, or it may be photographed and read. Light patterns are a simple and effective means of making an overall calibration of a recording

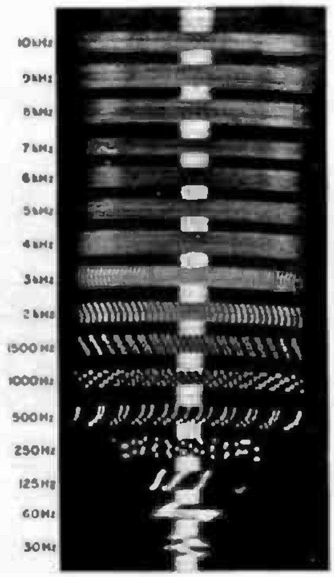


Fig. 23-75A. Typical monophonic light pattern made with a magnetic cutting head using a modified constant-amplitude, constant-velocity recording characteristic (RIAA). Speed 33⅓ rpm.

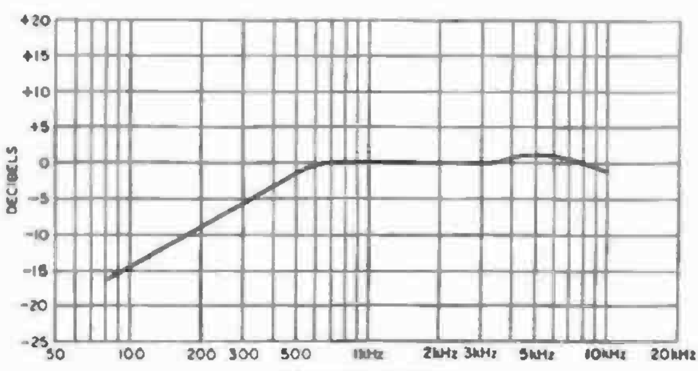


Fig. 23-75B. Frequency response of a magnetic cutting head plotted from measurements of frequency-band amplitudes.

channel or checking the frequency characteristics of a cutting head. The theory of the light pattern may be explained as follows. At the center of the pattern, an unmodulated groove makes an angle of 90 degrees with the incident ray and reflects a beam of light to the eye. Other parts of the groove appear dark. When the groove is modulated, reflections are visible despite the departure of the groove axis from the 90-degree direction. A point will exist on each waveform where the angle due to the modulation cancels the 90-degree angle because of the change iu mean direction.

Again, for a short distance within each waveform, the groove is at the

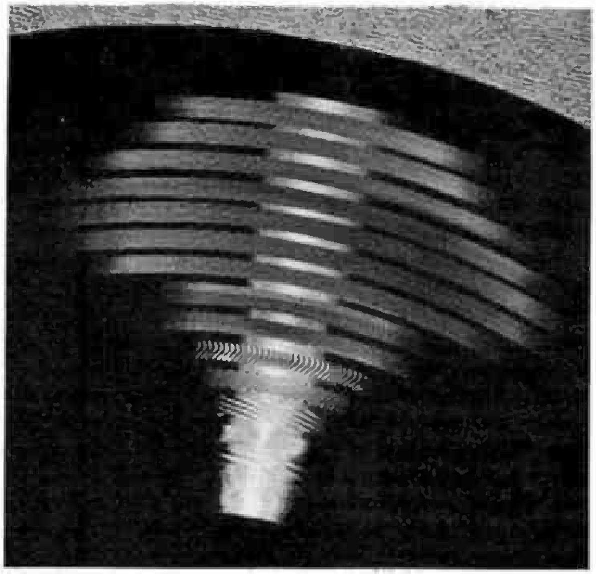


Fig. 23-75C. Monophonic light pattern of a pre-equalized recording channel showing the results of either improper equalization or improper cutting-head frequency response. Pattern made at 33⅓ rpm.

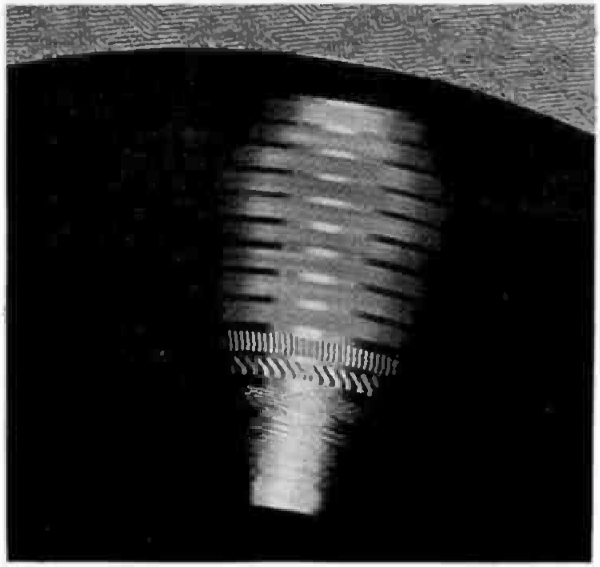


Fig. 23-75D. Monophonic light pattern of a cutting head made at 78 rpm showing a midrange frequency peak and a loss at the high frequencies. The response from this head would be accentuated in the midrange frequencies with a loss at the high frequencies.

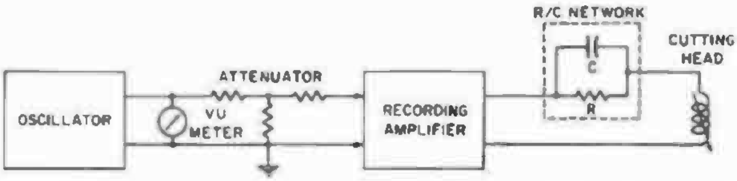


Fig. 23-75E. Circuit for measuring the frequency characteristics of a magnetic cutting head. The RC network is the normal recording network specified for the head.

90-degree position or parallel to the tangent at the center of the pattern. At a given distance from the center, the groove angle becomes so large that the cancellation of angles no longer occurs. This is the edge of the pattern.

As the groove diameters become smaller, the mean curvature increases, but the waveforms are becoming shorter and the modulation slope for a given frequency and amplitude increases in the same proportion. Thus, the width of the pattern is not affected by the changing groove diameter. A light pattern is principally a method of comparing the width of various frequency bands with each other, one band being a reference frequency. Light patterns may also be measured by playing back the pattern using a magnetic pickup. The results are plotted as frequency versus decibels as for any frequency characteristic. The width of the pattern is proportional to the voltage output from a constant-velocity pickup. If the amplitude of the frequency bands is measured mechanically, either by photographing or by viewing, variations from the reference frequency may be plotted in decibels:

$$dB = 20 \log_{10} \frac{A_1}{A_2}$$

where,

A_1 is the reference frequency,
A_2 is any frequency of interest.

A light pattern recorded using a modified constant-amplitude, constant-velocity recording characteristic will show whether the velocity is constant, by its having straight sides between the turnover frequency and the highest frequency at the outside of the disc. Below the turnover frequency, the frequency characteristic slopes off at a rate of 6 dB per octave, if amplitude is constant. (See Question 14.6.)

The light pattern in Fig. 23-75A was made with a magnetic cutting head using a modified constant-amplitude, constant-velocity recording characteristic and is typical of a well-adjusted cutting head. The recorded frequencies are indicated at the left of the pattern. The reference frequency is 1000 Hz, and the turnover frequency is 500 Hz. Variations in the amplitudes of the frequency bands are due to slight variations in the frequency response of the head. The variations are in the order of a few tenths of a decibel.

Fig. 23-75B shows a light-pattern frequency characteristic plotted from information obtained by measuring the amplitudes of the frequency bands. The reference frequency is 600 Hz.

If the frequency response of a cutting head appears to be satisfactory, a second pattern may be made of the complete recording channel to determine whether the recording characteristic (including the equalization) is correct. This is made in a manner similar to that used for the cutting-head light pattern, except that the equalization normally employed during recording is left in the circuit.

A light pattern made using pre-equalization in the recording circuits is shown in Fig. 23-75C. It may be seen the frequency response is not correct because the high frequencies fall off quite rapidly. To be correct, the high frequencies should continue to increase in amplitude.

Fig. 23-75D depicts a light pattern of a cutting-head response at a speed of 78 rpm. In this pattern, the middle frequency range is increased while the higher frequencies drop off, resulting in a peaked frequency characteristic.

Some difference must be expected in the frequency response of a cutting head made at speeds of 33⅓ and 45 rpm. Therefore, a compromise is made if it is to be used at both speeds.

Recording characteristics may vary with different styli. Therefore, the styli should be spot checked by recording a reference frequency band, then one at the high frequencies and one at the low frequencies.

The circuit for making a light pattern of the frequency response of a cutting head is shown in Fig. 23-75E. Only the recording amplifier that is normally used and the RC network permanently connected in the cutting head are used. The initial recording level is set to the reference frequency (100-percent modulation) and all other frequencies to be recorded are sent into the recording amplifier at the same amplitude. The amplifier must have a uniform frequency response to slightly below and slightly above the frequency range to be recorded by the head in order to prevent frequency variations in the light pattern due to frequency discrepancies in the amplifier.

When a light pattern of a complete recording channel is to be made using pre-equalization in the recording circuits, the initial level (100-percent modulation) must be set at the amplitude of the highest equalized frequency. If this precaution is not observed, the cutting head will be overloaded and possibly damaged.

If the recording channel is pre-equalized to complement the RIAA and NAB Standard (17.17 dB at 15,000 Hz) this frequency is used as the reference frequency. As this frequency has the maximum amplitude, all other frequencies will be lower in amplitude. However, if the recording channel and cutting-head characteristics are correct, the pattern will appear on the disc as would the RIAA frequency response of Fig. 13-95.

Since the advent of hot-stylus recording (see Question 15.59) the problem of getting the correct frequency response on the disc has diminished.

If diameter equalization is used, it is measured separately and added algebraically to the characteristic of the cutting head.

An overall pattern of the two foregoing characteristics may be recorded, provided the reference frequency is taken at the point of maximum equalization for the two forms of equalization to prevent damage to the cutting head.

Although the described light patterns were made using a monophonic cutting head, the procedure is the same for stereophonic light patterns. Corrections to the frequency response are made by adjustment of the negative feedback in the stereo cutting-head driving amplifier. (See Questions 14.2, 17.228, 18.330, and 18.331.)

23.76 What is the procedure for measuring turntable rumble and high-frequency noise?—The NAB Standard of March 1964 specifies that for a monophonic disc reproducing system, the low-frequency noise (rumble) generated by the turntable, pickup, and preamplifier, when playing back an essentially rumble-free silent groove, shall be at least 40 dB below a reference level of 1.4 centimeters per-second peak velocity at 100 Hz. The electrical response of the preamplifier is to conform to the standard NAB reproducing curve (Fig. 13-95), and as Question 13.103.

The test circuit in Fig. 23-76A is used. The test amplifier and indicating meter are to have uniform response withiu plus or minus 1 dB between 10 and 250 Hz, with 500 Hz being 3 dB below the 100-Hz response, and an attenuation rate of at least 12 dB per octave at frequencies above 500 Hz. The amplifier and meter are to decrease at the rate of 6 dB per octave below 10 Hz. The meter is to have the ballistics of a standard VU meter. If the meter fluctuates, maximum values are to conform to this requirement. The pickup-arm resonance must fall outside the passband or be sufficiently damped as to not affect the measurement.

The measurement reflects the electrical effect, not the aural annoyance value, of low-frequency noise. Experience indicates that a strong low-frequency noise (rumble) at a frequency below audibility will cause severe in-

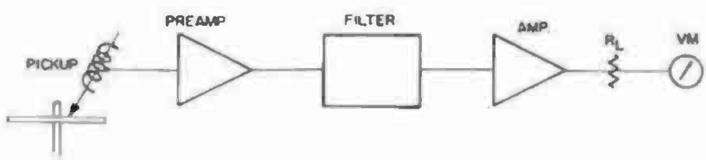


Fig. 23-76A. Test circuit for measuring low-frequency noise (rumble) of a turntable.

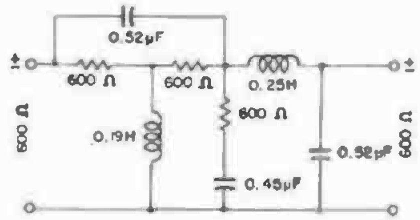


Fig. 23-76B. Filter for measuring the relative rumble loudness level (RRLL) (after Baurer).

termodulation products to be generated and is more serious in a wide-range reproducing system than is an audible low-frequency noise. The reference level of 1.4 centimeters per second at 100 Hz corresponds to an amplitude of 7 centimeters per second at 500 Hz; thus, this falls in the constant-amplitude section of the recording characteristic.

As a rule, the rumble or low-frequency noise developed in a professional turntable will be at least 50 dB or more below the signal. Rumble can generally be traced to bearings, gears, motor-drive systems, pucks, and motor vibration. Rumble will be greater at the larger diameters than at a smaller radius. The waveform error of the amplifier measurement must be negligible down to 10 Hz, as rumble frequencies are of steep wavefront configuration and fall below 20 Hz.

High-frequency noise is measured on a flat velocity basis over a range of 500 Hz to 15,000 Hz, and it is to be at least 55 dB below the level obtained under the same conditions of reproduction, using a reference tone of 1000 Hz, recorded at a peak velocity of 7 centimeters per second. The response of the measuring system at 500 Hz is to be 3 dB below that of 1000 Hz, and fall off at a rate of 12 dB per octave, or more, below 500 Hz. The response at 15,000 Hz is to be 3 dB below the response at 1000 Hz, and then to fall off at a rate of 12 dB or more, above 15,000 Hz.

For stereophonic noise measurements, the electrical requirements for the measuring equipment are the same, except that the low-frequency noise (rumble) is to be not less than 35 dB below a reference level of 1 cm per second peak velocity at 100 Hz in either plane of modulation. This corresponds to an amplitude of 5 centimeters per second peak velocity at 500 Hz, operating in the constant-amplitude portion of the recording characteristic. High-frequency noise is to be at least 50 dB below 100 Hz, at a peak velocity of 5 centimeters per second.

A method of measuring turntable rumble, suggested by Baurer of the CBS Laboratories, includes the use of a filter network which compares the level of a 1000-Hz, 5 cm/s rms lateral tone on a test record with a level of the turntable rumble weighted with a 6-dB per octave network having a 500-Hz turnover point. This conforms to the NAB practice of attenuating frequencies above 500 Hz at the rate of 12 dB per octave to eliminate the surface noise. The configuration of the relative rumble loudness level (RRLL) is given in Fig. 23-76B, and its frequency characteristic is shown in Fig. 23-76C. The procedure for such measurements follows. The system is first adjusted for the conventional NAB-RIAA frequency response, and the output of the system is set for a convenient level indication, using a 1000-Hz test frequency with a lateral displacement of 5 cm/s rms.

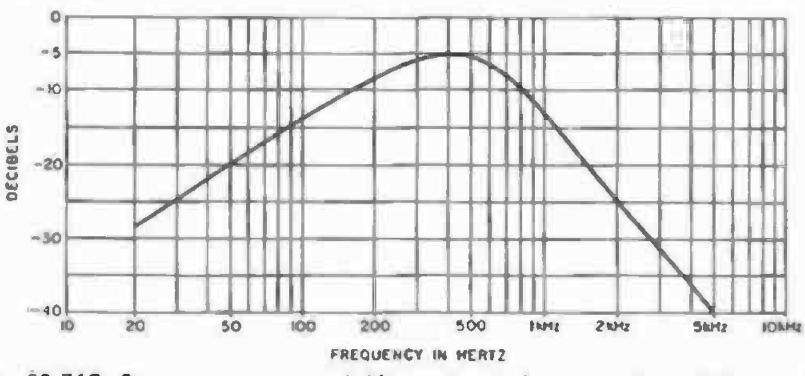


Fig. 23-76C. Frequency response of filter network for measuring relative rumble loudness level (after Baurer).

After noting the level, the network is inserted in the circuit, terminated, and the output voltage again is read using an unmodulated groove. The RRLL may now be computed as 20 Log_{10} times the ratio of the two voltages.

For stereophonic reproduction, two readings are required: one with the pickup connected for lateral reproduction and the other with the pickup in the vertical mode. The total rumble is then the square root of the lateral and vertical voltages. The RRLL is the quotient of the overall rumble voltage and the 1000-Hz tone voltage expressed in decibels. It is important that the network be properly terminated at both the input and output sides.

23.77 ***What is an N-A beam test record?***—A special test record developed by the Cook Laboratories for testing the percentage of intermodulation distortion of a record reproducing system. If the system has 2-percent intermodulation distortion or greater, the listener will hear a code letter, N (— ·). If the intermodulation is less than 2 percent, a letter A is heard (· —).

The test record consists of two frequencies which sweep the audio band while maintaining a constant 1000-Hz separation. One signal sweeps a coded letter N signal from 19,000 Hz down to 3000 Hz, while the second signal (uncoded) sweeps from 20,000 Hz downward to 4000 Hz. During the sweep period of the above frequencies, a 1000-Hz constant-frequency coded A, with an amplitude of 2 percent of the average carrier envelope, is used as a reference for the 1000-Hz difference between the two sweep frequencies.

If the intermodulation distortion is 2 percent, the 1000-Hz frequency A predominates. Over 2 percent, the A signal overrides the 1000-Hz frequency and becomes the letter N.

23.78 ***How is a cutting head stylus tested for noise so as to obtain the greatest signal-to-noise ratio?*** — As shown in Fig. 23-78. A vtvm is connected across the cutting head. The recorder is started and an unmodulated groove is cut while the noise generated by the stylus as it cuts the recording medium is observed. Many times a very slight realignment of the recording stylus will reduce the cutting noise by several decibels. The noise level between different styli may be measured in a similar manner. It is not uncommon to find new styli with unacceptable noise levels. It is common practice in recording activities to make this test at the outside diameter of the record before each recording by switching the necessary equipment into the circuit before the start of each take.

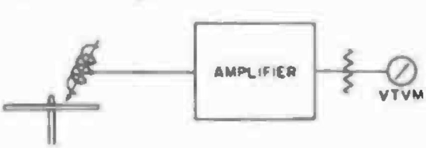


Fig. 23-78. Circuit used for aligning a recording stylus for the greatest signal-to-noise ratio.

23.79 ***How is the translation loss due to decreasing groove velocity measured?***—By recording a frequency of 10,000 Hz of constant amplitude, starting at the outside diameter of the disc and continuing down to the smallest diameter normally recorded. The disc is then played back and the signal level is measured as the pickup travels across the disc. The grooves at the outer edge are used as a reference level. To obtain a true measurement of translation loss in the preceding manner, the signal to the recording system must be carefully maintained at a constant amplitude. (See Question 13.49.)

23.80 ***How may the acoustical frequency response of a microphone be measured if an anechoic chamber is not available?***—Unless an anechoic chamber is available, the true acoustic frequency response of a microphone cannot be measured. (See Question 2.83.) However, if a standard microphone or one of known frequency characteristics is available and it is of the same design as the one to be tested, a comparison measurement may be made, which, for all practical purposes, will be acceptable. The equipment for measurement is set up as shown in Fig. 23-80A.

A speaker of good frequency response and low distortion is supported on a stand, with the microphone to be used as a standard set at a distance of about 12 to 18 inches from the speaker and on the axis of the speaker voice coil. The speaker is driven by an amplifier of low distortion connected to an audio oscillator and attenuator. A vacuum-tube voltmeter is connected across the output of the amplifier and is used to maintain a constant input voltage to the speaker. The output of the microphone is connected to a preamplifier, attenuator, and voltage amplifier to which a vacuum-tube voltmeter or VU meter is connected.

The measurement is started by setting the oscillator to 1000 Hz and adjusting the level of the speaker output to a plus 10 dBm. The voltage amplifier at the output of the microphone is then adjusted for a reading of 0 dBm on the meter. This reading is then used as a reference level. The desired frequencies are applied to the speaker while a constant voltage is maintained across the output of the driving amplifier. The variation in frequency response is read at the output of the voltage amplifier and is the frequency response of the microphone plus the speaker and the room acoustics. If the microphone is placed within 12 to 18 inches of the speaker diaphragm, the room acoustics will have little effect on the frequency response, because the majority of the sound picked up by the microphone will be direct rather than reflected sound.

After a measurement of the standard microphone has been made, it is removed and the microphone to be compared is connected in its place. A second measurement is then made in a similar manner.

It may be necessary to increase or decrease the gain of the voltage amplifier to obtain the same output level as was used with the standard. If a large group of microphones similar in design to the standard is available, a comparison of their sensitivities may be made. If the output level appears to be down several decibels compared to the standard, it is an indication that the magnets may need recharging.

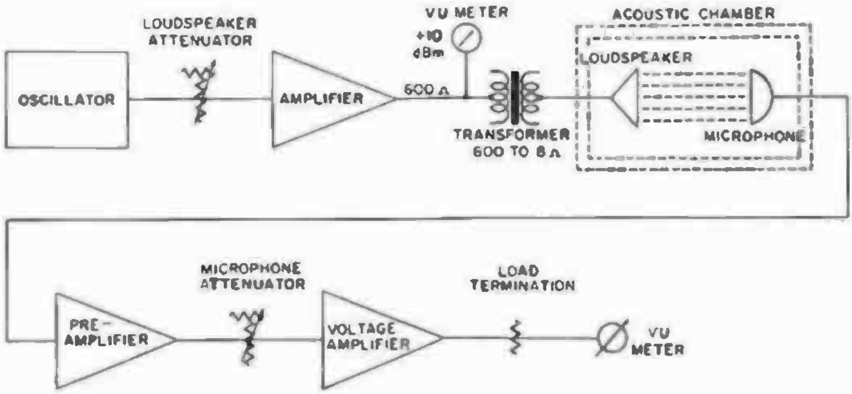


Fig. 23-80A. Method for measuring the frequency response of a microphone using a standard microphone for comparison.

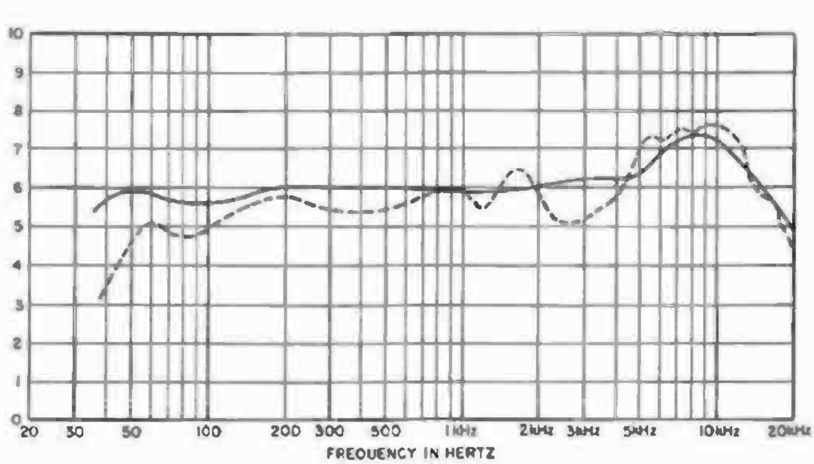


Fig. 23-80B. True curve of a dynamic microphone and a curve made using the comparative method.

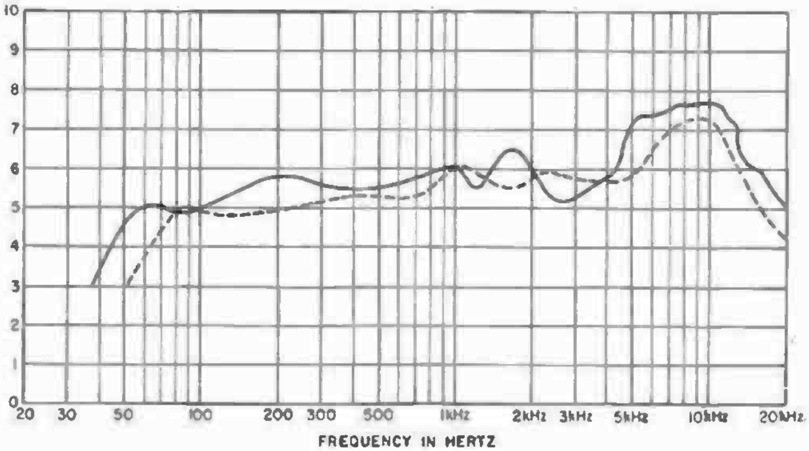


Fig. 23-80C. Frequency response of two microphones measured using the comparison method for determination.

Measurements plotted using the comparison method appear quite different from those supplied by the manufacturer of the microphone. However, this is of no particular consequence because the suspected microphone is being directly compared to one of known quality. Any variations noted between the standard and the suspected microphone are an indication that the suspected microphone does not meet the characteristics of the standard. Under *no* circumstances should the frequency response obtained using the comparison method be construed as a true frequency response of either microphone.

In Fig. 23-80B is shown the true curve (supplied by the manufacturer) of a standard microphone of the dynamic type (moving coil) and a curve of the same microphone plotted with the comparative method of measurement. It will be noted there is a considerable difference in the characteristics between the two methods of measurement, although the two curves are representative of the true characteristics. The peaks and valleys of the comparison curve are due to the microphone characteristics, room acoustics, and speaker frequency response.

Fig. 23-80C plots the standard microphone of Fig. 23-80B and a microphone suspected of needing repair. This microphone appears to have lost its sensitivity at both the low- and high-frequency ends.

The principal points of importance in making comparative measurements are that the microphone under test must be placed in the exact position the standard microphone occupied, and the levels to the speaker must be the same. The acoustics of the room in which the tests are conducted must not be disturbed by objects which would add to or reduce the reflections as originally measured using the standard microphone.

The results of the measurements are plotted as shown. The graphs are then laid over each other, and then 1000-Hz levels matched. The two graphs are next placed over a light box or held up to a light and the differences between the standard and the suspected microphone are compared.

23.81 How can a microphone be tested for extraneous magnetic field pickup?—By placing the microphone in a large loop of wire excited by a source of 115-volt, 60-Hz current as shown in Fig. 23-81, to test the effectiveness of the magnetic shielding used around the impedance-matching transformer in the microphone.

A coil of large diameter and small cross section appears to be the best design for this type of test because of the greater uniformity of the magnetic field close to the center of the coil. This assists in positioning the microphone as the position is then less critical.

The measurement of magnetic field pickup is made by placing a standard microphone, or one of known sensitivity to magnetic fields, in the exact center of the coil and rotating it for a maximum deflection on a vacuum-tube voltmeter connected across the output of the mi-

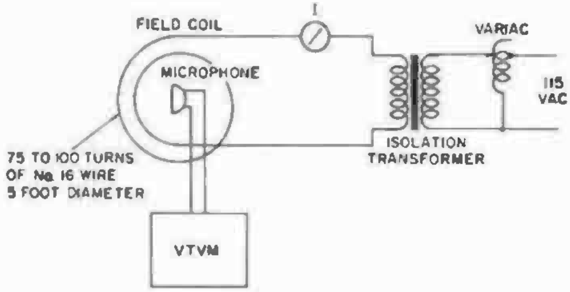


Fig. 23-81. Test for measuring the ac field pickup of a microphone.

crophone. The output of the microphone is measured open circuit, by using a vacuum-tube voltmeter with an input impedance of 1 megohm or greater. Other microphones are then placed in the coil and their sensitivity to magnetic pickup compared to the standard microphone.

This device is quite helpful when a large number of microphones are to be measured for magnetic pickup. The intensity of the magnetic field picked up by the microphone may be calculated as:

$$H = \frac{0.2\pi NI}{r}$$

where,

H is the magnetizing force in oersteds,
N is the number of turns in the coil,
I is the current in the coil,
r is the radius of the coil in centimeters.

The magnetizing force of the test coil should be kept between 0.10 and 0.50 oersted to assure that the field being measured is that of the coil and not the ambient field strength of the location where the measurement is being conducted.

In some instances, it might be advisable to excite the field coil with 120 or even 180 Hz because this frequency is many times encountered around power supplies and is the cause of considerable interference.

23.82 What are the different methods used for measuring the distortion of an amplifier?—The distortion products of an amplifier may be measured in several different ways. They are:

(a) Single-frequency harmonic distortion—A single frequency is applied to the input of the amplifier and the amplitude of the harmonics generated within the amplifier is measured at the output by means of a distortion-measuring set as described in Questions 22.62 and 22.63. The audio-frequency oscillator used for such measurements must have low harmonic distortion.

(b) Harmonic analysis—This measurement is made by the use of a harmonic wave analyzer, such as that described in Question 22.65, connected across the output of the amplifier. The amplitude of each individual harmonic generated within the amplifier is measured relative to the fundamental frequency. The total harmonic distortion is then computed in percent of the fundamental frequency.

(c) Intermodulation distortion—Two frequencies, one high and one low and mixed in a given ratio, are applied to the input of the amplifier. Sum and difference frequencies are measured at the output of the amplifier with an intermodulation analyzer, as described in Question 22.129.

(d) Difference-frequency intermodulation—Two frequencies, one low and one high and mixed in a ratio of 1:1, are applied to the input of the amplifier and varied in frequency while a constant difference frequency is maintained between them. The sum and difference frequencies in the amplifier output are measured by means of an intermodulation analyzer, described in Question 22.133.

(e) Square-wave response—A square wave having a fast rise time is applied to the input of the amplifier. The distortion of the square wave caused by passage through the amplifier is observed by means of an oscilloscope. With a square wave applied to the input, both distortion and transient effects may be observed. Square-wave generators are discussed in Question 22.54.

(f) Transient distortion — Transient distortion is measured by applying a series of sine-wave pulses to the input of the amplifier as described in Question 23.49. The effect of these pulses is observed at the output of the amplifier by means of an oscilloscope connected across the load termination. The method given in the IFH-A-201-1966 Standard is discussed in Question 23.208.

(g) Linearity—A signal of constant amplitude is applied to the input of the amplifier. The input level is increased in steps of exactly 1 dB. If the amplifier is linear, the output level will increase in steps of exactly 1 dB. At the level where the amplifier output does not increase 1 dB for an increase of 1 dB at the input, it is departing from its linear characteristic. Linearity measurements may be made by the method described in Question 23.36 or by means of a step generator described in Question 22.135.

(h) Phase distortion—This measurement can be made several different ways. The most common way is by the use of a dual-trace oscilloscope, a single-trace oscilloscope in combination with an electronic switch, or a phase meter as described in Question 22.108. Phase distortion can also be measured using Lissajous patterns, as described in Question 23.111.

23.83 ***What is the procedure for making a harmonic distortion measurement on an amplifier?***—The amplifier is connected as shown in Fig. 23-83. The signal may be applied from an oscillator and attenuator or from a gain set. The proper input circuit is selected from Fig. 23-14 or 23-15. The output of the amplifier is terminated in the recommended load impedance.

A distortion-factor meter of any one of the several different types described in Section 22 and a VU meter or vacuum-tube voltmeter are connected across the output termination of the amplifier. The signal at the input of the amplifier is increased to develop a given output level, or power, at the output of the amplifier. The distortion is then measured and tabulated as power output versus harmonic distortion. The foregoing procedure is used for several different output powers below the maximum rated power output and slightly above the rated output.

The results of these measurements are plotted as shown in Fig. 23-7G. The same procedure is used for a recording channel or other amplifying device. However, in the instance of a magnetic recorder, if the distortion is being measured while the machine is recording, a 20-kHz, low-pass filter must be connected in the output, as shown in Fig. 23-73, to eliminate the effect of the high-frequency bias current. (See Question 23.74.)

23.84 ***Show the waveform of an overloaded single- and double-ended amplifier.***—Fig. 23-84A shows the appearance of a 400-Hz waveform in an overloaded, single-ended amplifier. It will be noted the upper side of the waveform is flattened off while the lower side, although driven into overload, is only partially flattened off. The distortion of the amplifier was approximately 20 percent and contained principally even-order harmonics.

Fig. 23-84B is the appearance of a 400-Hz signal at the output of a low-wattage amplifier producing a power output of 12.5 watts. The distortion was approximately 12 percent and consisted of odd-order harmonics. The waveform in Fig. 23-84C is the appearance of the signal in the cathode circuit for the amplifier in Fig. 23-84B.

The trace in Fig. 23-84D is the appearance of the distortion of a push-pull amplifier measured to show the change in the linearity of the amplifier as it is driven into overload. Although this amplifier is overloaded, it is being driven into overload in a symmetrical manner; that is, it is being overloaded by being driven into saturation and cutoff by equal amounts. The distortion

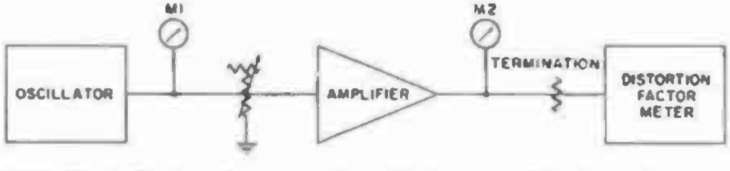


Fig. 23-83. Block diagram for measuring the harmonic distortion of an amplifier.

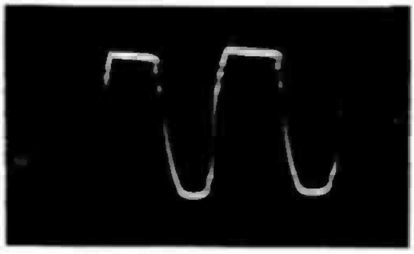


Fig. 23-84A. Single-ended amplifier driven into overload. Bias voltage high. Even-order harmonics are indicated.

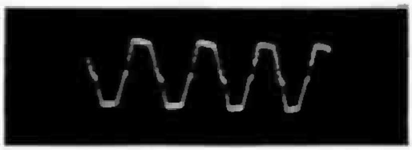


Fig. 23-84B. Output signal of an overloaded push-pull amplifier. Odd-order harmonics are indicated.

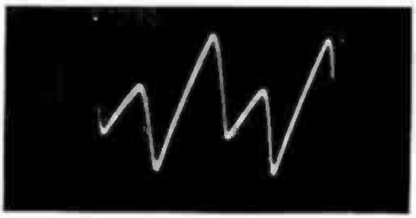


Fig. 23-84C. Waveform of a 400-Hz signal at the cathodes of an overloaded push-pull amplifier.

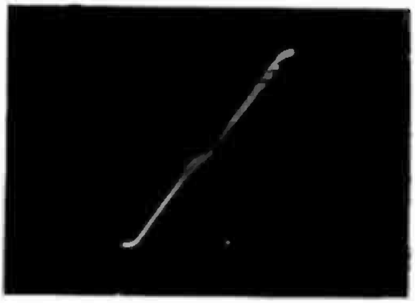


Fig. 23-84D. Symmetrical overloading of a push-pull amplifier. Curvatures at the ends of the trace indicate odd-harmonic distortion, but the amplifier is being driven equal amounts into saturation and cutoff.

shown is approximately 5 percent. The circuit in Fig. 23-92A was used to obtain the trace shown.

23.85 ***If a sine wave is flattened on one peak only, what order harmonics are indicated?***—Even-order harmonics. (See Fig. 23-84A.)

23.86 ***If a sine wave is flattened on both peaks, what order harmonics are indicated?***—Odd-order harmonics. (See Fig. 23-84B.)

23.87 ***If two amplifiers, one having 3-percent harmonic distortion and the other having 1 percent, are connected in tandem, is the total harmonic distortion 4 percent?***—No. The distortion products cannot be directly added. However, the total distortion may be calculated, if the percentages of the individual harmonics are known. If the amplitude of the individual harmonics is measured using a harmonic wave analyzer as described in Question 22.65, the total harmonic distortion for the two amplifiers can be computed. A more practical way is to connect the two amplifiers in tandem and measure the total harmonic distortion as described in Question 23.83.

23.88 ***What is a weighted distortion factor?***—A harmonic measurement in which the harmonics are weighted in proportion to the harmonic relationship.

23.89 ***What causes cross-modulation products in an amplifier?***—When a tone of constant amplitude is present with one which is varying in both amplitude and frequency, the constant-amplitude tone is modulated by the second tone and is distorted. This is called *cross modulation.*

23.90 ***What is scale distortion?***—Frequency discrimination noted when program material is reproduced at a level greater or lower than that at which it was originally recorded. Scale distortion may be corrected for by the use of a loudness control, as described in Question 5.65.

23.91 ***What are combination tones?***—Combination tones are frequencies produced in a nonlinear device, such as an audio amplifier, having an appreciable amount of harmonic distortion. Frequencies generated in the amplifier because of nonlinearity consist of the original and sum and difference frequencies between the fundamental and the harmonic frequencies of the original frequencies. This type of distortion is commonly known as intermodulation distortion and is described in Question 23.113. (Also see Question 22.129.)

23.92 ***Can an oscilloscope be used for the measurement of harmonic distor-***

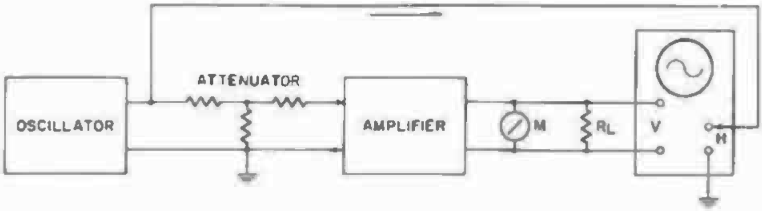


Fig. 23-92A. Connections for measuring harmonic distortion using a cathode-ray oscilloscope.

tion?—Yes, except it is rather difficult to see small values of distortion unless the oscilloscope has a magnifier for spreading the image. As a rule about 3- to 5-percent distortion is about the least that can be seen on a 5-inch screen. However for higher-order harmonics, this limit may extend downward to 3 or even 2 percent under certain conditions. It is highly important that the vertical and horizontal amplifiers have equal phase-shift characteristics. If this is not known, it should be measured and corrected before such measurements are made. (See Question 23.110.)

To check harmonic distortion using an oscilloscope, the oscilloscope is connected across the amplifier output load termination. The gain of the amplifier is increased to a point where the waveform shows the first indication of departure from the true sine-wave pattern. A frequency of 40 to 60 Hz is generally used for this test, although other frequencies may be used, depending on the circumstances.

A method of measuring harmonic distortion with an oscilloscope is shown in Fig. 23-92A. The vertical input of the oscilloscope is connected across the amplifier load termination. The horizontal input of the oscilloscope is connected to the signal source ahead of the attenuator network at the input of the amplifier. The internal sweep oscillator of the oscilloscope is turned off.

For output levels below the overload

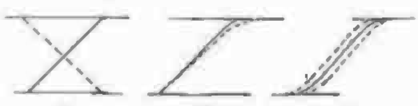


Fig. 23-92B. Distortion patterns seen on an oscilloscope using the connections in Fig. 23-92A. (a) Initial adjustment. (b) Even-harmonic distortion. (c) Odd-harmonic distortion.

point of the amplifier, the pattern will appear as a straight line, as in (a) of Fig. 23-92B, and leaning either to the left or right, depending on the number of stages in the amplifier. When a straight line has been obtained at a low level by adjusting of the oscilloscope controls, the input signal to the amplifier is increased until the trace starts to show a curvature at either top or bottom, or both, as in (b) and (c) of Fig. 23-92B.

If the curvature is at the top of the line (b) of Fig. 23-92B, even-harmonic distortion is indicated, and if it is curved at both ends (c) of Fig. 23-92B, odd-harmonic distortion is indicated.

The only drawback to this measurement is that it is difficult to judge the percent harmonic distortion and, at best, it is only an indication that distortion is present and consists of a given-order harmonic.

23.93 ***What effect does a rectifier-type meter have on harmonic-distortion measurements?***—If, when making a harmonic-distortion measurement, a VU or volume indicator meter employing a rectifier element such as copper oxide or selenium is used for setting the output level, the meter attenuator must be turned to its highest setting after adjusting the output level. If this precaution is not taken, harmonic distortion may be added to the measurement, resulting in erroneous values.

The added distortion is caused by the rectifier element clipping or squaring off the peaks of the sine-wave signal used for the distortion measurement. As the squaring off of the sine-wave peaks causes the generation of harmonics, the total measured harmonic distortion is increased. The effect of the meter may be easily demonstrated by measuring the distortion of a low-powered amplifier such as a preamplifier at an output level of plus 4 dBm, leaving the VU meter attenuator set to the plus 4-dBm position. A second measurement is made at the same output level, except this time the VU meter attenuator is set to its highest point, or around 30 dBm. It will be noted the harmonic distortion has been reduced by several tenths of a percent.

Setting the meter attenuator to a higher level after adjusting the output level inserts a higher value of resistance in the attenuator network and isolates the rectifier element of the meter from the output circuit of the amplifier.

23.94 How is the linearity of an amplifier measured?—By setting up the amplifier as for a normal distortion measurement, as described in Question 23.83. The output level is set for a value approximately 20 dB below the rated maximum power output. Loss is removed from the input attenuators in 1-dB steps. For each step removed, the output should increase exactly 1 dB. This procedure is continued until the removal of 1 dB (or less) results in a change of 0.10 to 0.20 dB at the output.

The level where the amplifier just starts to become nonlinear is the point where the output does not increase 1 dB for a change of 1 dB at the input. As a rule, this test requires an input attenuator that may be varied in 0.10-dB steps to arrive at the exact point of departure from linearity. (See Question 23.36.)

23.95 How are the characteristics of a motion picture projection system measured?—Special test films for both magnetic and photographic sound-track reproducers are available for measuring the frequency response, overload, and distortion. These films are run through the projection system, using the test circuit shown in Fig. 23-95. Overload is measured by using a constant-amplitude sound track, recorded to increase in 1-dB steps from a fairly low level to 100-percent modulation of the sound track. (Test films are available from the SMPTE.)

The output of the projection system is terminated in a resistive load, and the gain control is set for a level approximately 10 dB below the maximum rated power output level. The test film is run through the projection system, and the point of overload is observed on an output meter connected across the resistive termination, as in Fig. 23-95.

In a normal system, the output level should increase in 1-dB steps in a linear manner up to the point of overload. The overload point is that level where the output does not increase 1 dB for an increase of 1 dB from the test film. The frequency response is measured as for any amplifier system, using a test film of the type described in Question 19.68.

23.96 What is the procedure for measuring the frequency characteristics of a telephone or transmission line?—The line is terminated at both ends, using a balanced attenuator network with 6 to 10 dB of loss. The purpose of the pad is to supply a solid termination to the line and to isolate the measuring equipment from the line. Frequencies of constant amplitude are sent from an oscillator at the far end of the line, as in Fig. 23-96. After establishing a reference level at the receiving end, readings are made for each frequency and plotted with reference to 1000 Hz.

If the line is to be equalized, series resistance R of the equalizer configuration is adjusted for a uniform frequency response. The equalizer is always connected at the receiving end of line as shown. Design of telephone-line equal-

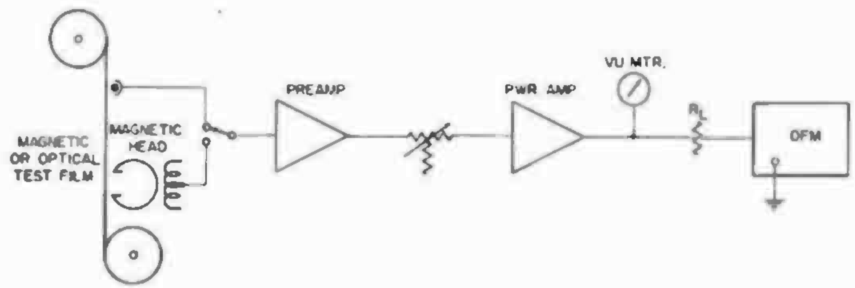


Fig. 23-95. Test setup for measuring the frequency response, overload, and distortion characteristics of a motion picture projection system for both magnetic and optical sound track.

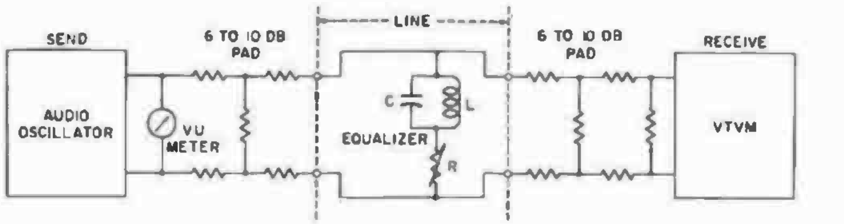


Fig. 23-96. Circuit for measuring the frequency response of a telephone or transmission line.

lizers is discussed in Questions 6.37 to 6.43.

23.97 ***What is cross talk and what is its cause?***—Cross talk is the introduction of an unwanted signal into one circuit by another, causing interference. Such interference is caused by inductive or capacitive coupling between circuits, as in Figs. 23-97A and B. Cross talk may also be caused by leakage between pairs.

The strength of a magnetic field around a wire varies with the square of the distance from the wire. The magnetic field has the same waveform as the current which is producing it. If the current is alternating, the magnetic field will also alternate. If the field cuts an adjacent conductor, an alternating emf will be induced in the second wire. The magnitude of this induced emf will vary inversely as the distance between the two conductors. This is termed inductive coupling (Fig. 23-97A). Capacitive coupling causes unbalance by inducing an emf into another circuit by virtue of the capacities existing between pairs 1 and 2 and ground, as shown in Fig. 23-97B. The capacitance to ground is shown by the two dotted capacitors, C_1 and C_2. These two capacitors may be considered to be in series. Therefore a voltage exists across each capacitor. Assuming that the capacitors have equal capacitance, the voltage drop across each is of the same magnitude but opposite in polarity.

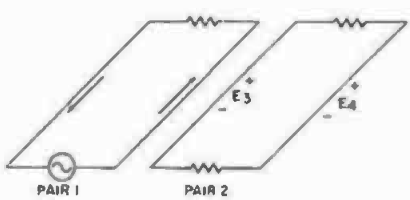


Fig. 23-97A. Causes of inductive cross talk.

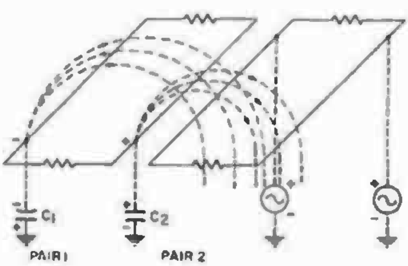


Fig. 23-97B. Causes of capacitive cross talk.

In a second circuit, pair 2 is running adjacent to pair 1. As shown, each wire of its pair is linked by the field existing at that point. The inner wire of pair 2 will have a stronger induced field than the outer wire. This causes the inner wire to be raised to a higher potential than the outer wire, and this unbalance of potentials causes cross talk in pair 2.

The unit for expressing cross talk is the *cross-talk unit* (CU). As this unit generally involves the use of small values of current, the values are multiplied by 10^6 if the circuits are of equal impedance, using the ratio of the currents in the circuits. When the circuit impedances are different, the amount of coupling is expressed by using the square root of the power ratio between the two circuits. Cross talk is directly proportional to the amount of coupling, expressed in CU between circuits. The greater the value of CU, the greater will be the cross talk.

Coupling between circuits can also be related to the strength of the signal in the disturbed circuit and the power in the disturbing circuit, by stating the ratio of the two powers in decibels. As this will result in a negative power gain, it is stated as a positive value and designated *coupling loss* in decibels. Therefore, coupling loss in dB is inversely proportional to the amount of cross talk.

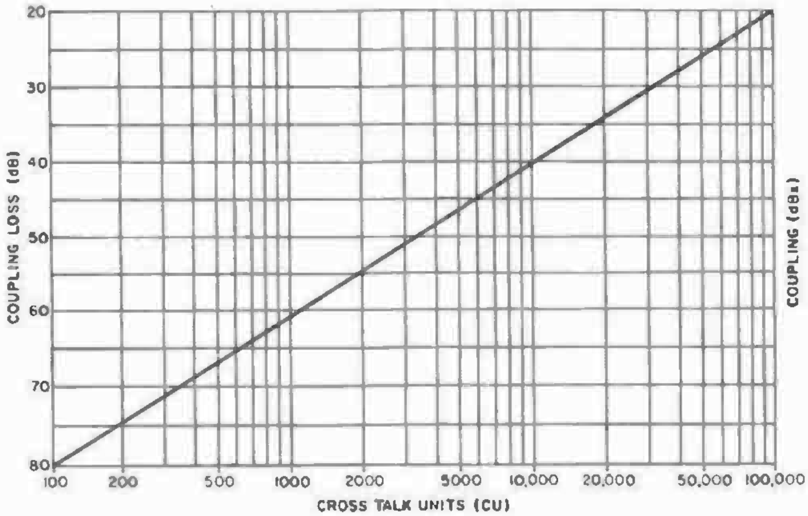


Fig. 23-97C. Relationship among Cu, dB, and dBx.

Cross talk is also stated in dBx, which takes into account the interfering signal effect on a 1000-Hz reference signal. As a rule, cross talk measurements are made using a standard test set employing a given weighting characteristic. (See Questions 2.93 and 10.36.) The relationship between cross talk and coupling in CU, dB, and dBx is shown in the graph of Fig. 23-97.

23.98 ***How is cross talk reduced in a transmission line?***—By transposition of the pairs as shown in Fig. 23-98. It will be noted pair 2 has been transposed or crossed over. This induces a greater emf in sections A and B. Similarly, a smaller emf is induced in wires C and D. This makes the resultant emf induced in the whole circuit nearly equal. Because unbalance has been reduced, the cross talk is at a minimum. In actual practice, pair 1 would also be transposed.

23.99 ***What is the maximum cross-talk level that can be tolerated in a cable pair?***—About 40 dB below the maximum signal level transmitted over the line.

23.100 ***What is the equation for calculating the loss of an audio-frequency line, neglecting the phase angle?***

$$dB = 20 \text{ Log}_{10} \frac{Z_1 + Z_2 + Z_{LP}}{Z_1 + Z_2}$$

where,

Z_1 is the source impedance,
Z_2 is the terminating impedance,
Z_{LP} is the dc resistance of the line in ohms per loop-mile.

Thus, for a loop one-mile long having a dc resistance of 4.02 ohms with a 250-ohm source and terminating impedance, the loss is:

$$dB = 20 \text{ Log}_{10} \frac{250 + 250 + 4.02}{250 + 250}$$
$$= 20 \text{ Log}_{10} \frac{504}{500}$$
$$= 20 \text{ Log}_{10} 1.01$$
$$= 20 \times 0.0043$$
$$= 0.086$$

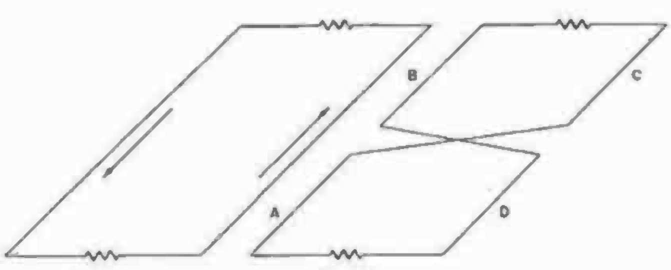


Fig. 23-98. Reduction of cross talk by transposition of pairs.

23.101 *What is near-end cross-talk and how is it measured?* — Near-end cross talk is interference between two transmission lines which is propagated in the disturbed pair in a direction opposite to that of the transmission of the disturbing pair. If a signal is sent into pair 1 at Station A in Fig. 23-101, and the cross talk measured at Station A in pair 2, it is called near-end cross talk.

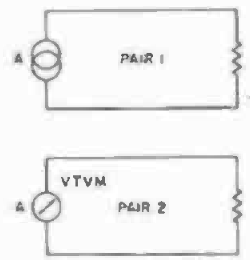


Fig. 23-101. Measuring near-end cross talk in a cable pair.

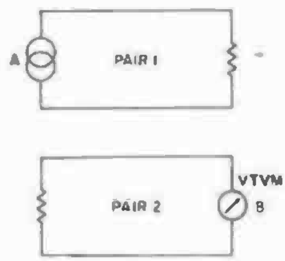


Fig. 23-102. Measuring far-end cross talk in a cable pair.

23.102 *What is far-end cross talk and how is it measured?*—Far-end cross talk is interference which travels along the disturbed circuit in the direction in which the signal normally travels in the circuit. To measure far-end cross talk, the test signal is transmitted on pair 1 at Station A and the cross-talk level measured on pair 2 at Station B as shown in Fig. 23-102.

23.103 *What effects does the type of terminating impedance have on a long transmission line?*—Fig. 23-103 shows the effect of terminating a long line with two different values of terminating impedances and the effect when the line is terminated using a 600-ohm pad of 10-dB loss at each end of the line. The best characteristic is obtained when the line is terminated using the pads, as may be seen by the wide variation of impedance when only a terminating resistance is used.

Terminating the same line with a 6-dB pad resulted in the impedance rising to 700 ohms at 50 Hz and dropping to 450 ohms at 10,000 Hz (not shown).

23.104 *What is the purpose of phasing an amplifier and how is it accomplished?*—It is the practice in large installations such as recording and broadcast networks to completely phase the system from the microphone to the transmission line or recording equipment. This is particularly important in

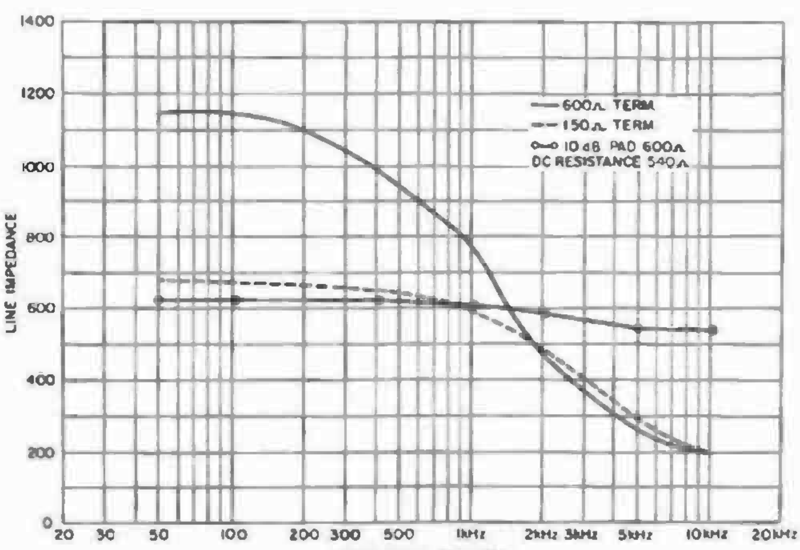


Fig. 23-103. Characteristic impedance of a long line when terminated using a pad and a terminating resistor.

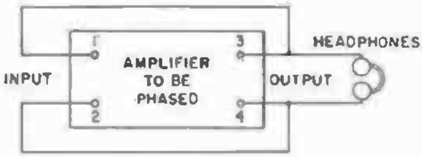


Fig. 23-104. Amplifier connected for phasing input and output circuits.

a system employing optical film-recording systems using noise-reduction amplifiers. Phasing the various units of a transmission system facilitates patching of equipment, with the assurance that equipment may be substituted in an emergency without running a complete test to determine whether the system is functioning properly.

Amplifiers may be tested for phasing by connecting the input and output terminals in parallel with a set of headphones as shown in Fig. 23-104. If a high-pitched squeal or no sound is heard, the amplifier input and output are out of phase. If a low-frequency motorboating type of sound (putt-putt) is heard, the amplifier is in phase. Out-of-phase conditions may be corrected by reversing the leads to either the input or output, but not both. When proper phasing has been established, the upper terminals of both the input and output are marked with a plus-or-minus sign (±).

23.105 How is a complete recording system checked for phasing?—By applying an unsymmetrical waveform to the input of the system and observing the waveform at the output of the system by means of an oscilloscope as shown in Fig. 23-105A.

In an optical recording system used for the recording of speech, it must be so phased that the noise-reduction system of the optical film recorder is at its position of maximum opening when the long side of the pressure wave (acoustic) at the microphone is applied to the light modulator. This is particularly true for a variable-area film recorder, because if this is not observed, the peaks of the modulation may be clipped.

The most practical way of phasing a recording channel is to place a common door buzzer in front of the microphone and observe its waveform on an oscillo-

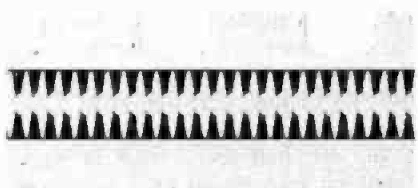


Fig. 23-105B. Nonsymmetrical photographic sound track used for phasing sound heads and recording systems.

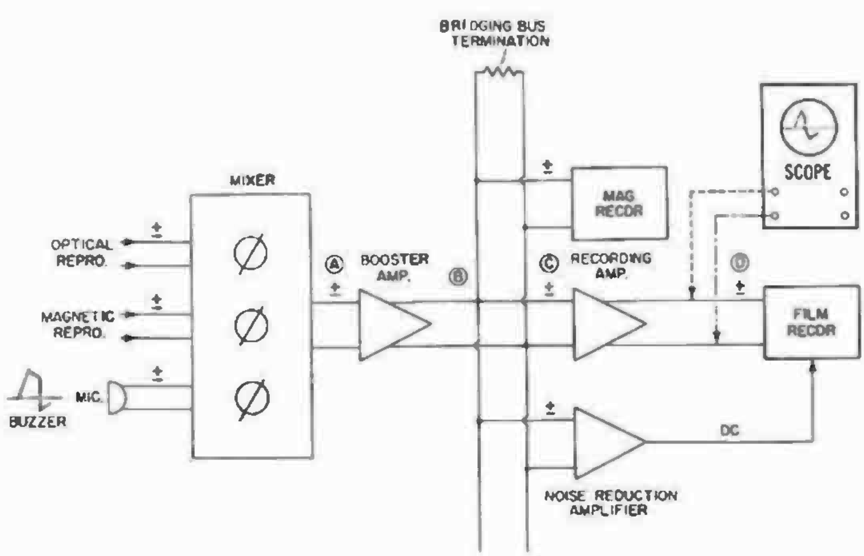


Fig. 23-105A. Procedure for checking the phasing of a recording channel. The distorted image of the buzzer at the microphone, or a distorted sound track must all appear to have the same phasing from points A to D.

scope connected at the output of the system. If the waveform is not sufficiently unsymmetrical, the adjusting screw on the buzzer is turned until the waveform on the oscilloscope appears similar to that in Fig. 23-105A.

With such a waveform applied to the microphone, the pressure wave (when the microphone diaphragm moves inward) will be in the correct phase. The noise-reduction amplifier is phased by reversing the audio input leads to obtain a maximum opening when the light modulator is deflected to 80-percent modulation.

To phase a group of optical sound-track reproducers, such as are used for rerecording, an optical sound track of 1000 Hz and of nonsymmetrical waveform is required.

This nonsymmetrical sound track can be made by connecting an oscilloscope across the output of a single-ended amplifier and overloading the amplifier until a nonsymmetrical waveform is obtained, as shown in Fig. 23-105B. The distorted waveform is recorded on a photographic film recorder at 80-percent modulation. Since only the negative will be used, no prints are made.

The distorted sound track is reproduced on each sound head in the installation, while the attitude of the waveform is observed on an oscilloscope connected across the bridging bus. The attitude of the waveform must he the same for each sound head; that is, the peak of the distorted waveform should be either above or below the oscilloscope reference line for all machines. The attitude of the waveform may be reversed by interchanging the leads from the photocell-amplifier output.

It is equally important that magnetic recording and reproducing equipment be in phase with the optical equipment. This is necessary to ensure that when magnetic sound tracks are transferred to optical, the signal voltage will be of the correct phase to open the noise-reduction equipment to its maximum, relative to the long side of the waveform.

To accomplish this latter phasing operation, the distorted optical sound track is played back and a magnetic sound track recorded. This sound track is then played back on each magnetic reproducer. The waveform must be of the same attitude as that of the optical reproducers. If not, the phase may be reversed by turning over the lead to the magnetic reproducer head, or the leads at the output of the reproducer amplifier, but not both.

The magnetic recorders are phased by playing back the magnetic sound track from a reproducer known to be in phase with the optical sound heads. If the waveform from the magnetic recorder is reversed, it may be corrected by reversing the leads to the recording head or the input of the recording amplifier.

As a final check, the distorted 1000-Hz sound track is again played back from an optical reproducer and the oscilloscope patched to the output of each individual piece of equipment in the system. The waveform must have the same attitude for each piece of equipment. If the channel is also used for dialogue, the microphone input must be phased to correspond. This is accomplished as described in the beginning of this Question.

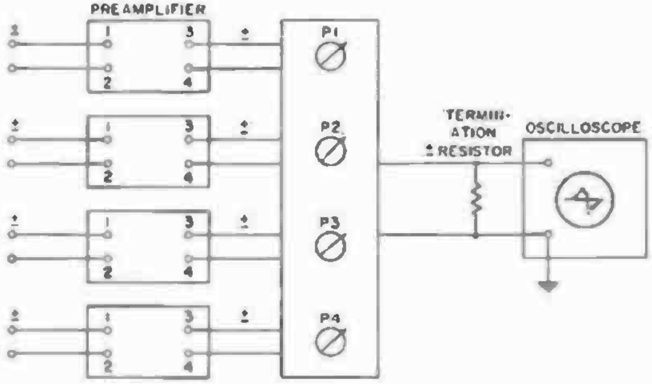


Fig. 23-106. Phasing procedure for monophonic mixer console.

Amplifier phasing is accomplished as described in Question 23.104.

The polarity of the oscilloscope can be determined by the used of a flashlight cell as described in Question 23.192. The polarity of the oscilloscope must be maintained the same throughout the phasing operations.

As a rule, most professional recording equipment is phased in advance by the manufacturer and, if the polarity markings are observed at the input and output terminals, the system will automatically be in phase.

23.106 How are mixer networks phased? — Mixer networks are first wired in phase; that is, the mixer pots and associated equipment are phased by color code, making certain that the grounds are connected to the same sides of the circuits. After the wiring has been completed (unless the network includes a coil), it should be possible to buzz from the ground terminal of each input to the ground side of the output circuit.

If the mixer network includes transformers, it is phased using the techniques outlined in Question 23.105. (See Fig. 23-106.)

Mixer inputs must be in phase with each other; otherwise, microphones which are in phase will be out of phase when used close to each other, resulting in a loss of level and increased distortion. The phasing of microphones is further discussed in Questions 4.84 and 4.85.

23.107 Describe the procedure for phasing a multiple or split-section mixer network.—For multiple or split-section mixer networks phasing of the individual sections and their components is essential for proper operation. Suppose that a three-section mixer network, described in Question 9.47, is to be phased. Refer to Fig. 23-107A, a simplified diagram of the mixing network. It will be assumed that the components have been phased by observing color codes and terminal markings, amplifier gains have been adjusted, and all equalizers and filters have been keyed out of the circuit. An oscillator test signal of 400 Hz is applied to mixer controls 1, 5, and 9, through a resistive combining network consisting of four branches. Mixer controls 1 to 12 are turned to their off positions (maximum attenuation). The attenuators of the four VU meters at the bridging buses are set for plus 14 dBm. Before starting the tests, *precautions must be taken to ensure that the four branches of the combining network and its patch cords are in electrical phase.* A suitable network with a 600-ohm impedance is given in Fig. 23-107B.

The tests are started by opening mixer control 1 to a position where the levels at bridging buses 1 and 4 (composite) indicate a plus 14 dBm. Leaving control 1, set control 5 (section 2) for a plus 14 dBm at bridging buses 2 and 4. The signal level at bus 4 will increase slightly or remain essentially the same. If the level at bus 4 *decreases*, the sections are out of phase with each other. Applying the test signal to controls 2 to 4, and 6 to 8, in turn, will determine if the turnover is in a single control, or if it is a complete section. If a control or section is found to be out of phase, the difficulty must be corrected before continuing the tests. The test signal is now applied to inputs 1 and 9 in section 3, and similar group tests for phasing are made.

The three sections are now tested by applying the test signal to mixer controls 1, 5, and 9. First, the level is set to plus 14 dBm for section 1, then for section 2, and finally for section 3. If the three sections are in phase, the signal will increase at bus 4 as each section input is opened. With the three sections operating, the signal will increase from 2 to 4 dB (above 14 dBm), with slight fluctuations because the different phase delays between the sections, cause a beat between the three signals. This is quite normal. If the sections are out of phase, the level will decrease when the out-of-phase input is opened.

After completion of the phasing tests, the oscillator is set to a frequency in the center of the passband of each filter and the filters are keyed in and out of the circuit. Some small irregular action of the meter across bus 4 may be expected as the three sections are tested together since the delay time is slightly different for each filter section. Equalizers, however, present a somewhat different problem. Although each equalizer may be set to the same frequency and value of equalization, because of manufacturing tolerances a slightly different phase shift is presented to bridging bus 4, and a beating effect will be noted.

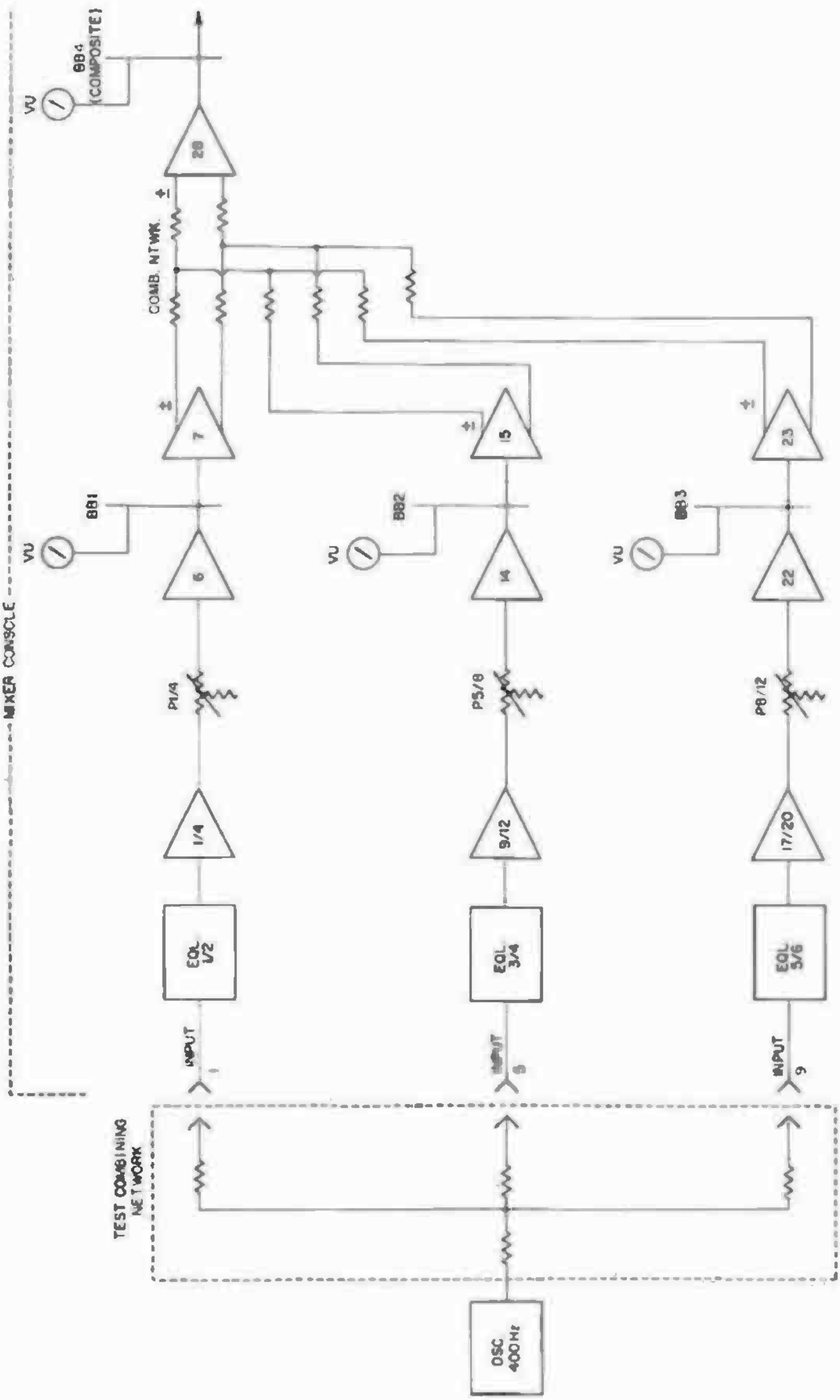


Fig. 23-107A. Simplified diagram of three-section mixer network used for the phasing example.

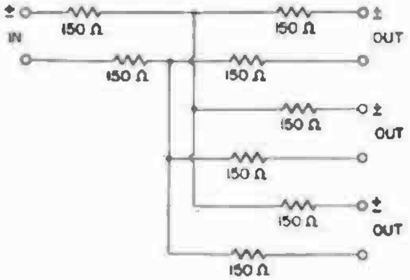


Fig. 23-107B. Combining network for phasing tests.

It should be remembered that when the mixer network is being used the three sections are independent of each other. This is important, since each section must be in phase within itself and in phase with respect to the other sections. If not, cancellation will affect the recording of single tracks from bridging bus 4, as all three input sections of the network feed this bus through the combining networks at the input of amplifier 28. This is the reason why the three head tones (one for each track) when applied to a three-track recorder are recorded in sequence rather than simultaneously. If they were recorded simultaneously, the level at bridging bus 4 would vary from 2 to 4 dB with a beat effect between the three tones. The tone would then be useless for setting future levels, such as could be required for transfer or for the matching of additional sound tracks.

23.108 How are stereophonic mixer networks phased?—The same general procedures as outlined in Questions 23.106 and 23.107 are followed. Each individual channel must be in phase within itself and in phase with the other channels. The best procedure to follow after the wiring has been buzzed out is to apply an unsymmetrical waveform to the input of the channel and to check the phasing at each individual component. After installation, each channel is to be checked against the other channels, the several sections being checked for similar phasing.

23.109 What precautions must be observed when making phase-shift measurements with an oscilloscope?—If the internal amplifiers of the instrument are to be used, they must have identical frequency and phase-shift characteristics. As a rule, the manufacturer states in the instruction manual the phase-shift characteristics for each amplifier.

With older-model oscilloscopes, the phase-shift characteristics were not always the same, and therefore phase-shift measurements were made by connecting directly to the deflection plates inside the tube. However, unless considerable signal voltage is available, this method is not recommended.

23.110 How are the phase-shift characteristics of an oscilloscope measured? — When phase-shift measurements are made on an unknown piece of equipment, both the horizontal and vertical amplifier of the oscilloscope being used for the measurement must have the same phase-shift characteristics. If the phase-shift characteristics are unknown, they may be measured by connecting the inputs of the two amplifiers in parallel, as shown in Fig. 22-110.

First, the signal is applied to the horizontal amplifier and its gain is adjusted for a convenient deflection of the signal (6 centimeters pk-pk) with the beam centered on the graticule vertical and horizontal lines. Leaving the gain control set, the signal is then applied to the vertical amplifier, and its gain control adjusted for a similar deflection. The signal is then applied to both amplifiers simultaneously. If the amplifiers have similar phase shift, the display will be a 45-degree straight line. The amount of phase shift in degrees can be determined as discussed in Question 23.111.

If the oscilloscope is of dual-trace design, the two traces may be superimposed, one on the other, and compared. This same measurement should be made at various settings of the input attenuators, and for both the vertical and horizontal attenuators, particularly for the positions of high attenuation. As

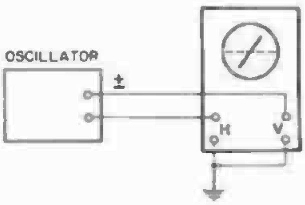


Fig. 23-110. Connections for checking the unknown phase-shift characteristics of an oscilloscope.

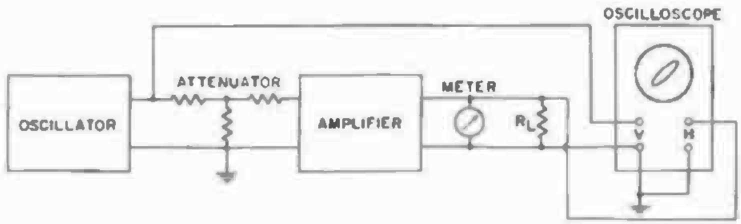


Fig. 23-111A. Block diagram for measuring the phase shift of an amplifier by using the internal amplifiers of a cathode-ray oscilloscope.

a rule, in modern oscilloscopes the phase shift between amplifiers is 1 degree or less.

23.111 ***How is the phase shift of an amplifier measured with an oscilloscope?*** —The amplifier is connected to the vertical and horizontal amplifiers of an oscilloscope, as shown in Fig. 23-111A. With the sweep circuit off, the measurement is started by sending a 400- or 1000-Hz signal from the oscillator into the amplifier and adjusting the gain for a given output level. The vertical amplifier of the oscilloscope is turned to its off position and the horizontal gain is adjusted for a convenient deflection (the larger, the better). The signal to the horizontal amplifier is removed without disturbing the controls. The signal from the oscillator is now applied to the vertical amplifier and the vertical gain is adjusted for exactly the same deflection as was used for the horizontal circuit. Both signals are now applied simultaneously to the horizontal and vertical amplifiers. The result is a 45-degree diagonal-line display.

For a condition of zero phase shift, the line is straight and of normal width. As the phase shift increases, the line will separate into two traces until it becomes an elliptical or circular pattern. Changing the frequency of the oscillator from a low to high frequency will generally tend to increase the in-phase shift.

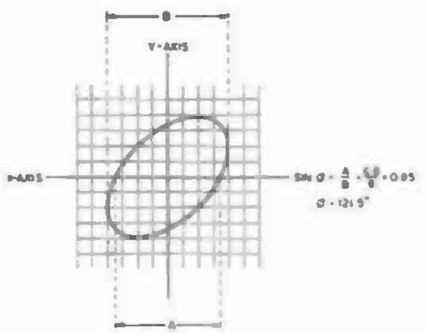


Fig. 23-111B. Phase displacement between two signals.

Fig. 23-111B shows a typical elliptical display arrived at with the method given for calculating phase shift. If the width of the display is measured along the horizontal line indicated A and the vertical line B, the phase shift for the pattern shown is:

$$\text{Sin } \phi = \frac{A}{B} = \frac{6.8}{8.0} = 0.85 = 121.5°$$

It will be noted that when phase-shift measurements are made with an oscilloscope, two displays are obtained for each degree of phase shift. The question then becomes one of determining whether the shift is lagging or leading. This may be determined by inducing a small amount of phase shift in series with the lead from the oscillator output. As an example, a 40-degree ellipse will become rounded and tend to become a circle as the induced delay is increased, while a similar 320-degree pattern will shift toward a straight line as delay is added.

When the initial measurements are made, the delay component in the leads from the oscillator must be out of the circuit; otherwise, the phase-shift display will be in error. The induced delay is used only for determining the final phase angle. Each time the frequency of the oscillator is changed, both the vertical and horizontal deflections must be rechecked and the beam centered. If the oscilloscope amplifiers are within 1-degree phase shift of each other, the measurements are easily made.

23.112 ***How is phase shift measured using a calibrated phase-shift network?*** —The equipment is connected as shown in Fig. 23-112. The signal is applied to the phase-shifting network and then to the input of the amplifier. The oscilloscope is adjusted as described in Ques-

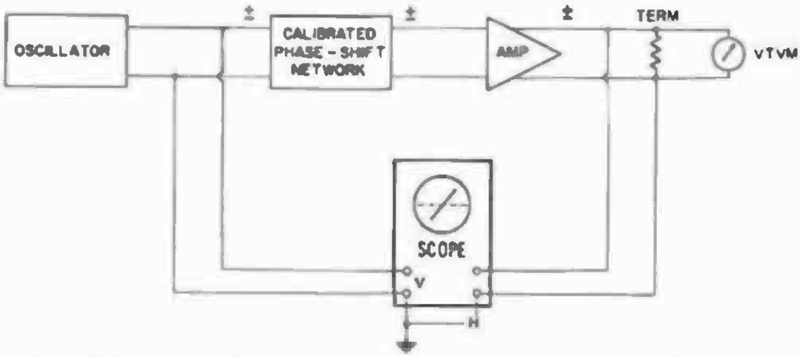


Fig. 23-112. Phase-shift measurement using a calibrated phase-shift network.

tion 23.111. After the pattern on the oscilloscope is obtained, the phase-shifting network is adjusted to bring the pattern to zero phase shift. The phase shift may now be read directly from the calibrated dial of the phase-shifting network.

This method of measurement is considerably faster than that described in Question 23.111, particularly if a large number of measurements is required. A given measurement is good only for a given output level of the amplifier under test. Each time the level is changed, the oscilloscope will require readjustment.

23.113 What is the procedure for measuring the intermodulation distortion of an amplifier?—The amplifier is connected to an intermodulation analyzer, as shown in Fig. 23-113A and described in Questions 22.129 and 22.130. The output section of the intermodulation signal generator is connected to the input of the amplifier to be measured, using a repeat coil if necessary. If a repeat coil is reqnired it should be measured individually to determine whether it contributes any intermodulation to the measurement. (See Question 23.114.) As a rule, the amount is negligible and may be ignored.

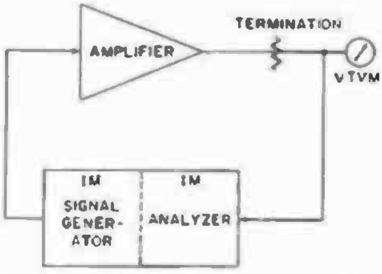


Fig. 23-113A. Amplifier connected for intermodulation distortion measurement.

The output of the amplifier is terminated and connected to the input of the intermodulation analyzer section. The analyzer section is generally designed for bridging the output termination of the device under test to absorb as little power as possible. However, in some types of analyzers, the input circuit is designed to supply a 600-ohm power termination. This feature may be checked by referring to the manufacturer's data sheet for the instrument.

The procedure for measuring the intermodulation distortion of a typical amplifier is as follows: The signal generator section is set to the desired combination of frequencies, which, for this discussion, will be assumed to be 40 and 2000 Hz in a ratio of 4:1; that is, the low frequency is 12 dB higher in amplitude than the high frequency. This combination signal is applied to the input of the amplifier at a level which will result in the desired output level. The gain of the analyzer is then advanced for the proper carrier-current level, and the percent of intermodulation is measured in the manner described for the particular analyzer being used.

If a rectifier-type meter is employed for measuring the output level of the amplifier, its attenuator should be advanced (after the output level is set) to a setting at least 10 dB higher than the measured level. This will prevent the rectifier element from affecting the waveform of the output signal and adding to the distortion. This subject of rectifier distortion is discussed in Question 23.93. If a vacuum-tube voltmeter is used to set the output level, its attenuator may be left at any setting without fear of affecting the measurement.

Level (dB)	*Equiv. Watts*	*Impedance* 4	8	16	150	250	500	600
+20.0	0.1	.519	.734	1.04	3.18	4.10	5.80	6.35
23.01	0.2	.774	1.04	1.47	4.49	5.80	8.20	8.98
26.99	0.5	1.16	1.64	2.32	7.12	9.17	13.0	14.2
28.45	0.7	1.37	1.94	2.75	8.41	10.9	15.4	16.8
30.0	1.0	1.64	2.32	3.28	10.1	13.0	18.3	20.1
33.01	2.	2.32	3.28	4.64	14.1	18.4	25.9	28.4
34.67	3.	2.84	4.02	5.68	17.4	22.5	31.8	34.8
36.02	4.	3.28	4.64	6.56	20.1	25.9	38.7	40.2
36.99	5.	3.67	5.19	7.34	22.5	29.0	41.0	44.9
37.78	6.	4.02	5.68	8.04	24.6	31.8	44.9	49.2
38.45	7.	4.34	6.14	8.68	26.6	34.3	48.5	53.2
39.03	8.	4.64	6.57	9.29	28.4	36.7	51.9	56.9
39.54	9.	4.92	6.96	9.84	30.2	38.9	55.0	60.3
40.0	10.	5.19	7.34	10.4	31.8	41.0	58.0	63.5
40.42	11.	5.44	7.69	10.9	33.3	43.0	60.8	66.6
40.79	12.	5.68	8.04	11.4	34.8	44.9	63.5	69.6
41.14	13.	5.91	8.36	11.8	36.2	46.8	66.1	72.4
41.46	14.	6.14	8.68	12.3	37.6	48.5	68.6	75.2
41.76	15.	6.35	8.98	12.7	38.9	50.2	71.0	77.8
43.01	20.	7.34	10.4	14.7	44.9	58.0	82.0	89.8
43.98	25.	8.20	11.6	16.4	50.2	64.8	91.7	101.
44.67	30.	8.98	12.7	18.0	55.0	71.0	101.	110.
45.44	35.	9.70	13.7	19.4	59.5	76.7	109.	119.
46.02	40.	10.4	14.7	20.8	63.6	82.0	116.	127.
46.53	45.	11.0	15.6	22.0	67.4	87.0	123.	135.
46.99	50.	11.6	16.4	23.2	71.1	91.7	130.	142.
47.78	60.	12.7	18.0	25.4	77.8	101.	142.	156.
48.45	70.	13.7	19.4	27.5	84.1	109.	154.	168.
49.03	80.	14.7	20.8	29.4	89.9	116.	164.	180.
49.54	90.	15.6	22.0	31.1	95.3	123.	174.	191.
50.0	100.	16.4	23.2	32.8	101.	130.	183.	201.

Fig. 23-113B. Equivalent sine-wave voltage for intermodulation measurements.

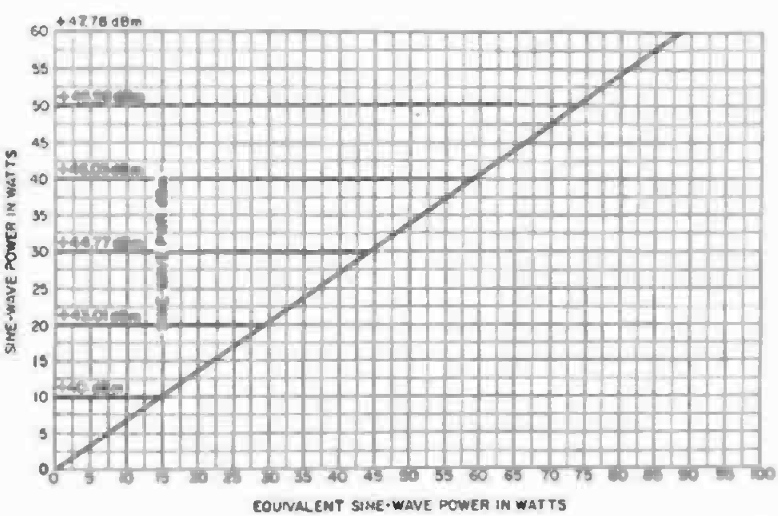


Fig. 23-113C. Graph for converting sine-wave power to equivalent power for two frequencies mixed in a ratio of 4:1. Equivalent sine-wave power equals sine-wave power times 1.47.

Gain-set terminations using an autotransformer in the receiving section shonld not be used to terminate the output of a device when intermodulation measurements are made, as they often add intermodulation distortion due to overloading of the autotransformer core material. As a rule, a high-wattage, vitreous, wirewound resistor will function quite satisfactorily for a termination because its inductance is small; therefore, it may be considered to be noninductive.

The amount of low-frequency intermodulation distortion measured is governed by the frequency of the lowest test signal, and the high-frequency distortion is governed by the frequency of the higher signal. For amplifiers of wide frequency range, 40 Hz and 7000 Hz are generally used. For medium frequency range, 60 Hz and 2000 Hz are used; and for limited frequency range, 100 and 2000 Hz are used. The low-frequency test signal should approximate the low-frequency cutoff frequency of the amplifier.

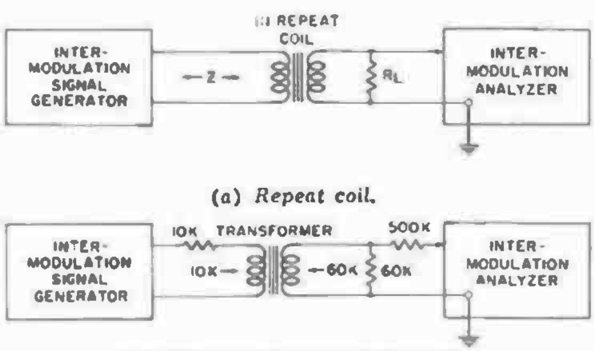
One of the most important points of an intermodulation measurement is to evaluate the developed output power in terms of *equivalent sine-wave power*. If this factor is not taken into consideration, the amplifier may easily be misrated relative to its output power capabilities. When an intermodulation measurement is made, the input signal level is adjusted for a value that will result in the same output voltage for a single-frequency distortion measurement. The power output is then calculated from the rms voltage developed across the load termination. Using a 4:1 ratio for the test signals, the indicated power is multiplied hy a factor of 1.47. This new power level is called the "equivalent sine-wave" power output.

As an example, an amplifier developing 10 watts of power with two input signals in a ratio of 4:1 will have an equivalent sine-wave power output of 14.70 watts. The power output is then plotted, using the equivalent sine-wave power output values, similar to that shown in Fig. 23-7E.

For convenience of measurement, a table of sine-wave power and equivalent sine-wave power for impedances of 16 ohms and 600 ohms is given in Fig. 23-113B. Fig. 23-113C shows the relationship graphically. It should be pointed out there is no simple way of expressing a mathematical relationship between an intermodulation and a harmonic-distortion measurement. A rough approximation indicates the ratio of intermodulation distortion to harmonic distortion falls between 3.2 and 4.0.

One of the advantages of an intermodulation measurement is that such measurements can be made in the face of considerable flutter, such as could be encountered in the measurement of a turntable or magnetic recorder. Intermodulation distortion in high-quality amplifiers generally measures between 0.10 and 1 percent, and up to 5 percent in a fairly good amplifier.

23.114 How may the intermodulation distortion contributed by a repeat coil or transformer be measured?—The coil is connected, as shown in Fig. 23-114, between the send terminals of an intermodulation signal generator and the receive terminals of the analyzer section. If the coil has a high-impedance winding, it must be properly terminated as shown. The send level to the coil will depend on the purpose for which



(a) *Repeat coil.*

(b) *High-impedance transformer.*

Fig. 23-114. Circuits for measuring the intermodulation distortion of transformers.

it was designed and must not be exceeded.

Output transformers should not be measured in this manner since they are required to handle considerable power. Therefore they are measured in the output of an amplifier similar to the one in which they are to be operated. Input transformers are best measured connected in the input of a single-stage amplifier of low distortion.

As a rule, a high-quality coil will add little distortion to the measurement. Tests on typical repeat coils used for transmission measurements reveal that, with the coil in the circuit, the intermodulation distortion is increased approximately 0.01 percent.

23.115 How should the intermodulation characteristics of an amplifier be stated?—In the following form: The amplifier has 1-percent intermodulation distortion at a power output of 10 watts, using two frequencies of 40 and 2000 Hz mixed in a ratio of 4:1 (or other), the amplitude of the low frequency being 12 dB greater than the high frequency. The characteristic shown is equivalent sine-wave power output.

23.116 How is intermodulation distortion related to harmonic distortion?—A very rough approximation assumes the relationship to be:

$$IM/HD = 3.2 \text{ for single-ended amplifiers}$$
$$= 3.8 \text{ for push-pull amplifiers}$$

An intermodulation measurement should be considered to be a separate and distinct measurement of a certain type of distortion and should not be compared to a single-frequency measurement. There is no simple mathematical relationship between the two types of measurement.

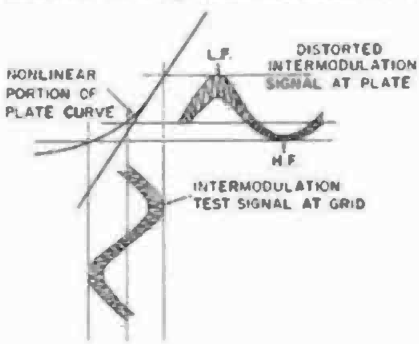


Fig. 23-117A. The generation of intermodulation distortion in a vacuum tube operating on the nonlinear portion of the characteristic curve.

23.117 Describe how intermodulation distortion is generated in a vacuum tube.—Basically, the generation of intermodulation distortion in a vacuum tube is similar to the generation of harmonic distortion as described in Question 12.114. The difference between intermodulation and harmonic-distortion measurement is that the harmonic measurement employs only one frequency and intermodulation testing makes use of two frequencies.

For illustration the intermodulation distortion of a single-ended triode amplifier stage operating class-A with normal operating voltages and load impedance will be investigated (Fig. 23-117A). Two frequencies of 40 and 2000 Hz, as shown in Fig. 23-117B, will be

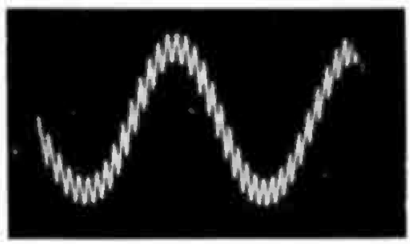


Fig. 23-117B. Two frequencies of 40 and 2000 Hz in a mixed ratio of 4:1.

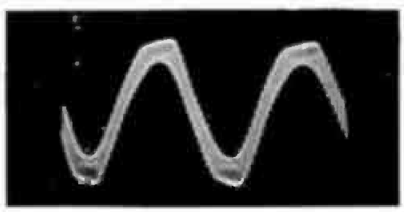


Fig. 23-117C. Distorted intermodulation signal.

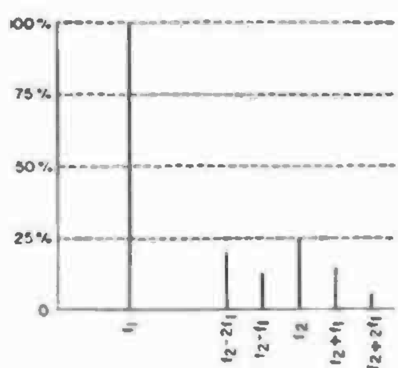


Fig. 23-117D. Intermodulation spectrum with sum and difference frequencies.

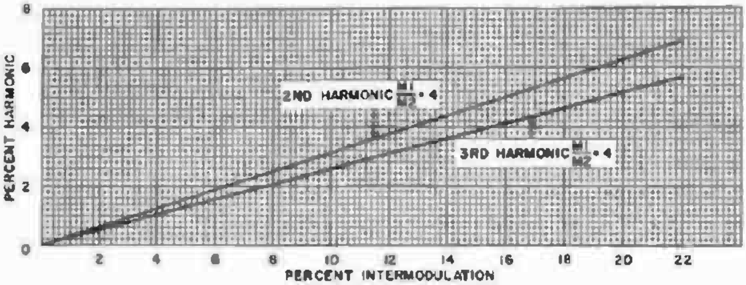


Fig. 23-117E. Relation of intermodulation to harmonic generation.

used in a mixed ratio of 4:1. The ratio of 4:1 is obtained by setting the low frequency 12 dB higher than the high frequency. It will be assumed the combined output signal of these two frequencies is sufficient to drive the plate-circuit signal into the toe region of the plate-current characteristic.

Because of the curvature of the toe end of the transfer characteristic, the symmetrical test-signal waveform will be distorted at the output. Since the low-freqnency component of the test signal is of greater amplitude than the high-frequency component, it is distorted to a greater extent due to the deviation in the slope of the transfer characteristic. Thus, the high-frequency component is modulated by the low frequency at a rate of 40 Hz. Modulation of the high-frequency component by the low-frequency component causes a flattening off of the high-frequency peaks in the output (Fig. 23-117C) and a generation of sum and difference frequencies (Fig. 23-117D).

Percent intermodulation distortion may be defined as the ratio of the average deviation of the amplitude of the high-frequency signal above and below the mean value to a mean value expressed in percent. Any mathematical relationship between readings of percent intermodulation and percent harmonic distortion must be made to only a second or a third harmonic. Therefore:

$$\text{IM distortion} = \frac{\text{percent intermodulation}}{\text{percent 2nd harmonic}}$$

$$= \frac{4m_1}{m_1 + m_2}$$

where,

m_1 is the amplitude of the lower frequency,
m_2 is the amplitude of the higher frequency.

When m_2 (second harmonic) is small in comparison to m_1 (low frequency), and the ratio is 4:1, the theoretical intermodulation percentage is 3.2 times that of the second-harmonic percentage. Where m_1 equals $4m_2$, the theoretical intermodulation is 3.84 times that of the percentage third harmonic for reasonably low values of distortion. Therefore:

$$\text{IM distortion} = \frac{\text{percent intermodulation}}{\text{percent 3rd harmonic}}$$

$$= \frac{6m_1^2}{(m_1 + m_2)^2}$$

where,

m_1 is the amplitude of the lower frequency,
m_2 is the amplitude of the higher frequency.

For an average set of conditions, the numerical relationship between percent intermodulation and percent harmonic distortion lies somewhere between the ratios of 3.2:1 and 4:1. A graphic relationship of these two types of distortion is shown in Fig. 23-117E. The foregoing discussion applies only to the SMPTE method of measuring intermodulation distortion.

23.118 ***If the power falls off appreciably at the low frequencies when an intermodulation distortion measurement is being made, what component should be suspected?*** — As a rule, the ontput transformer is the cause of low power output at the lower frequencies. It is more difficult for a transformer to develop a given power output at the lower frequencies using an intermodulated signal than it is to produce an equivalent amount of power using a single freqnency.

23.119 ***What frequencies are recommended for intermodulation measurements of a lateral disc recorder?*** — 400 and 4000 Hz. A low freqnency of 400 Hz is in the constant amplitude portion of the recording characteristic; therefore, it is subject to maximum displacement.

23.120 When one is listening for distortion, at what level should the program material be reproduced? — Because the characteristics of the human ear are such that it produces internal distortion which tends to mask distortion products in the program material, a given percent of distortion is more easily observed at the lower listening levels. This applies equally well to both intermodulation and harmonic distortion. Therefore, it is somewhat easier to judge the reproduction qualities of a reproducing system if played at a medium-low level.

Intermodulation distortion will be found to be more irritating than harmonic distortion.

23.121 What effect will hum frequencies in an amplifier have on an intermodulation distortion measurement? —If the ac ripple frequency is excessive in the output of the amplifier, intermodulation-distortion products are generated. As an example, if the power supply employs a 60-Hz full-wave rectifier, the 120-Hz ripple frequency will beat with a 1000-Hz signal frequency and produce intermodulation distortion products at 1120 and 880 Hz as well as other frequencies. These frequencies will be in addition to any distortion generated by the actual amplifier stages. It is important in a high-quality amplifier that hum frequencies be reduced to an absolute minimum.

23.122 How can an amplifier be tested for hum modulation?—By cutting off the low-frequency intermodulation test signal and measuring the intermodulation distortion, using the 120-Hz ripple frequency of the amplifier power supply for the low-frequency component of the intermodulation signal generator.

Because the hum-frequency amplitude will not be correct for the usual ratio of 4:1 employed in the conventional intermodulation analyzer, the measure of the intermodulation distortion will not be correct. However, this test will determine whether hum modulation is present.

23.123 When is a one-to-one test signal ratio used for intermodulation measurements?—When a more stringent test is required. However, in tests to determine the power-carrying characteristics of the output transformer in an amplifier, the 4:1 ratio is preferred.

23.124 Show the difference between an intermodulation and a harmonic distortion characteristic for a given amplifier.—Fig. 23-124 shows two curves plotted for the same amplifier. One curve is for a single frequency of 400 Hz, and the other, an intermodulation curve, is for 40 and 2000 Hz.

It will be noted that up to about 5.5 watts the amplifier being tested with two frequencies performs fairly well with 2-percent intermodulation. Using a single frequency of 400 Hz, the amplifier produces approximately 7 watts of power with 2.5 percent total rms

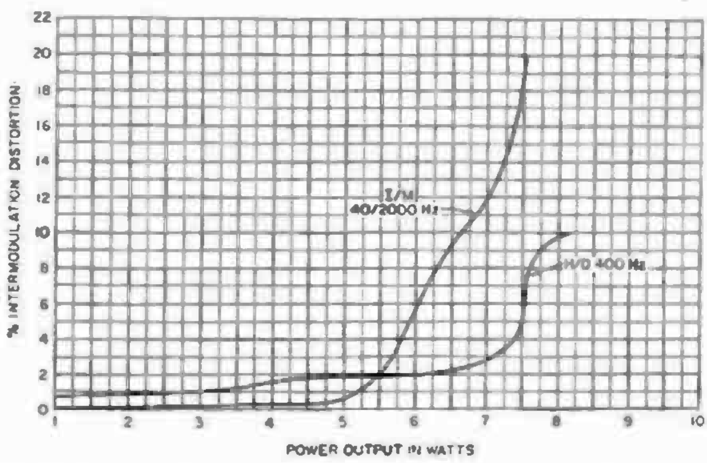


Fig. 23-124. Intermodulation and harmonic distortion curves for an inexpensive 10-watt amplifier.

harmonic distortion. This clearly illustrates the performance difference of this amplifier when measured with an intermodulation analyzer. No doubt the reason for the reduced output power when measuring by the intermodulation method is that the output transformer will not handle the power at a frequency of 40 Hz. This is typical of what may be expected in an inexpensive amplifier.

23.125 ***How does a capacitive load affect the intermodulation-distortion characteristics and stability of an amplifier?***—A considerable amount, if the amplifier is of unstable design. Fig. 23-125A shows the intermodulation distortion characteristics of a 20-watt amplifier of the Williamson ultralinear type when terminated with a 16-ohm resistive load. The waveforms were traced from photographs taken at the oscilloscope terminals of an intermodulation analyzer similar to the one described in Questions 22.129 and 22.130, using frequencies of 40 and 2000 Hz mixed in a ratio of 4:1. The equivalent sine-wave power output and percent intermodulation distortion measured is indicated under each tracing.

Waveforms from the same amplifier but with a capacitive load of 0.025 μF

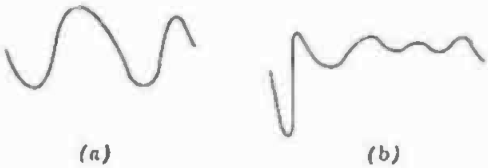


Fig. 23-125A. Intermodulation products of a Williamson ultralinear amplifier with 16-ohm resistive termination. (a) 10 watts output, IM = 0.15%. (b) 20 watts output, IM = 0.55%. (c) 22 watts output, IM = 3.2%.

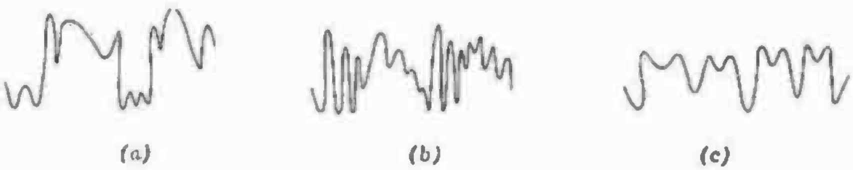


Fig. 23-125B. Intermodulation products of the Williamson ultralinear amplifier with an 0.025-μF capacitor connected in parallel with a 16-ohm resistive termination. (a) 10 watts output, IM = 0.75%. (b) 20 watts output, IM = 1.25%. (c) 22 watts output, IM = 3.75%.

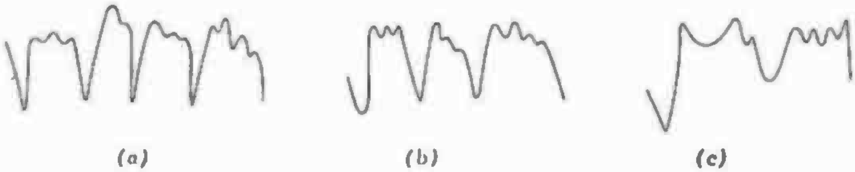


Fig. 23-125C. Intermodulation products of the Williamson ultralinear amplifier with an 0.06-μF capacitor connected in parallel with a 16-ohm load resistor. (a) 10 watts output, IM = 0.75%. (b) 20 watts output, IM = 1.5%. (c) 22 watts output, IM = 4%.

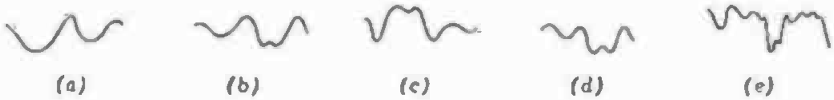


Fig. 23-125D. Intermodulation products of a McIntosh MC-30 amplifier with 16-ohm resistive load termination. The connection of a capacitor across the load termination makes no change in the distortion components. (a) 10 watts output, IM = 0.14%. (b) 20 watts output, IM = 0.20%. (c) 25 watts output, IM = 0.20%. (d) IM = 0.21%. (e) IM = 0.25%.

connected in parallel with the 16-ohm resistive termination are shown in Fig. 23-125B. In (a) of Fig. 23-125B, the distortion products have increased from 0.15 percent to 0.75 percent, or five times the value measured with only a resistive termination. At a 20-watt output, the distortion is 1 percent; but the distortion products have increased aud are of different frequency. Fig. 23-125C shows what happens when the capacitive loading is increased to 0.06 μF. The distortion patterns vary considerably from those previously discussed. The values of capacitance of 0.025 to 0.06 are representative of the capacitive loading an amplifier is subjected to when an electrostatic speaker and a conventional speaker system with 50 feet of interconnecting cable are used.

When a capacitive load is connected across the output winding of an amplifier, a large value of capacitive reactance is reflected to the plates (or plate) of the output stage. In some instances, this capacitive reactance may reach the equivalent of several microfarads. The magnitude of the reflected capacitive reactance is dependent on the turns ratio of the output transformer. (See Question 8.34).

The tracings shown in Fig. 23-125D illustrate similar measurements on a McIntosh MC-40 amplifier (rated 40 watts) using 6L6B beam-power tubes in push-pull in the output stage, as described in Question 12.231. Adding a capacitive load of up to several microfarads across the load termination had no effect on the intermodulation-distortion characteristics. From the foregoing discussion it may be assumed that intermodulation-distortion products vary with circuit design and the capacitive reactive component of the load impedance.

In some instances, the amplifier may oscillate as shown in Figs. 23-125E, F, G, and H, which are photographs of measurements made on a negative feedback amplifier of mediocre design. The tracings shown were made under the same conditions as for the previously described experiments. Using a resistive termination, the amplifier appears to be quite stable. However, when a capacitive load was connected in parallel with the resistive termination, the amplifier oscillated. A well-designed amplifier should not oscillate when a capacitor is connected in parallel with the normal load resistance. The effects of speaker load for transistor and tube-type am-

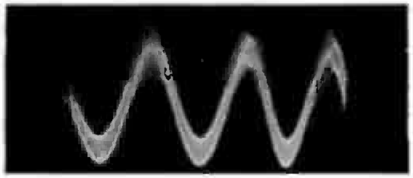


Fig. 23-125E. Waveform at output of an unstable negative-feedback amplifier when a capacitive load is connected in parallel with the resistive load termination. Waveform shown is a 40/2000 Hz intermodulation signal with a 4:1 ratio.

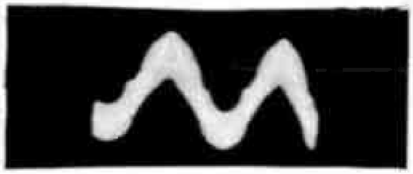


Fig. 23-125F. Waveform at output of unstable negative-feedback amplifier with an 0.01-μF capacitor connected across a 16-ohm resistive termination. Waveform is an intermodulation signal of 40 and 2000 Hz. The wide white band is high-frequency oscillation superimposed on the low-frequency signal.

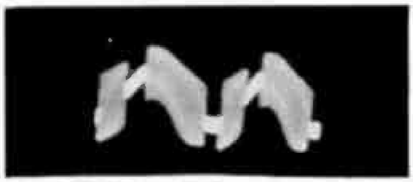


Fig. 23-125G. Waveform at output of an unstable negative-feedback amplifier with an 0.05-μF capacitor connected in parallel with a 16-ohm resistive termination. The light line is 40/2000 Hz intermodulation signal. Heavy portion of signal is high-frequency oscillation.

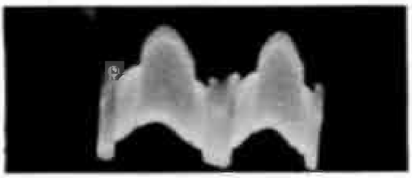


Fig. 23-125H. Waveform at output of an unstable negative-feedback amplifier with an 0.10-μF capacitor connected in parallel with a 16-ohm resistive termination. Intermodulation signal of 40 and 2000 Hz. Large peaks are high-frequency oscillation.

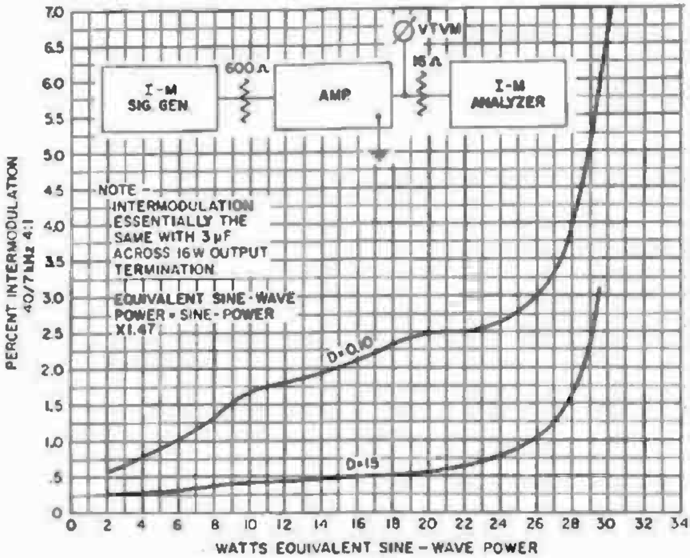


Fig. 23-126. Effect of a variable damping control on the intermodulation distortion of a 20-watt amplifier.

plifiers are discussed in Questions 20.90 and 20.91.

23.126 ***How are the distortion characteristics of an amplifier affected by the use of a variable damping control?*** —Fig. 23-126 shows two intermodulation-distortion measurements made on an amplifier having variable damping and working into its normal load resistance. The curve designated 0.10 is the minimum setting of the damping control, while the curve designated 15 is the maximum setting. As may be seen, the power output and distortion are both affected by the position of the damping control.

Variable damping control is discussed in Question 12.242.

23.127 ***How may a speaker be tested for intermodulation distortion?*** —As a rule, the speaker is set up and excited by an intermodulation generator. The sound from the speaker is picked up on a calibrated microphone and the intermodulation distortion is measured, using an intermodulation analyzer similar to that described in Questions 22.129 and 22.130. In the absence of the foregoing equipment, a simple test may be made by applying 60 Hz to the voice coil for a full deflection of the diaphragm. A frequency of 2000 or 6000 Hz is then added to the 60-Hz signal. If, by listening, a low-frequency warble is heard, intermodulation distortion is present because of the nonlinearity of the cone movement. (See Question 20.96.)

23.128 ***How are cross-modulation or intermodulation measurements made on an optical film recording channel?*** —A cross-modulation or intermodulation analyzer is generally used with a photographic film recording channel for the express purpose of determining the optimum negative and print densities and for the quality control of optical sound tracks. The cross or intermodulation signal generator is connected to the input of the recording channel or directly to the recording amplifier, and a series of negative sound tracks are made at several different negative densities. Prints of a given density are then made from these negatives and the intermodulation or cross modulation is measured. The results of these mea-

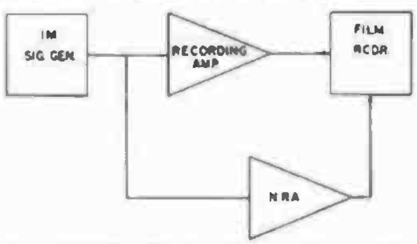


Fig. 23-128. Block diagram for making intermodulation sound tracks, using a photographic film recorder. The intermodulation signal is fed directly to the recording amplifier which is driving the film recorder.

surements are then plotted as described in Section 18.

A block diagram for such measurements is shown in Fig. 23-128. The exact procedure will depend on the type equipment used for the measurements. The measurement of intermodulation distortion from an optical sound track is similar to the measurement of an amplifier with the sound track replacing the signal generator section of the analyzer. The measurement includes any intermodulation induced by the photocell preamplifier and other amplifiers used for the reproduction of the sound track.

23.129 ***How is the intermodulation distortion of a magnetic tape recorder measured?***—It is measured in a manner similar to that described in Question 23.74. An intermodulation measurement is the ideal way of determining the correct bias-current value, because it will permit the measurement in the face of considerable flutter.

Single-frequency harmonic distortion measurements on magnetic tape recorders are at times rather difficult, particularly if an appreciable amount of flutter is present, because the oscillator signal floats in and out as the balance is adjusted on the distortion analyzer. The final results are plotted as shown in Figs. 23-74E and F. The block diagram of the test setup is shown in Fig. 23-129. The purpose of the 0.10-μF capacitor is to remove the residual bias signal from the measuring device.

23.130 ***How is the signal-to-noise ratio of a system or device measured?***—In the case of a disc-recording channel, 400 Hz is applied to the input of the system and several grooves are recorded at 100-percent modulation, followed by several unmodulated grooves cut using the normal groove dimensions. The recorded 400 Hz is played back and the level noted. The unmodulated grooves are then played back and the noise level measured. The difference between the two measurements in decibels is the signal-to-noise ratio of the system.

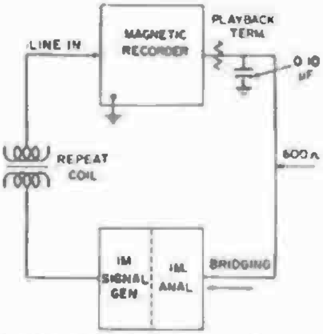


Fig. 23-129. Block diagram for measuring the intermodulation distortion of a magnetic recorder.

For optical film recording systems, a 400-Hz, 100-percent modulated sound track is recorded, followed by an unmodulated sound track. These tracks are processed in the normal manner for the system. The tracks are played back and the noise level of the unmodulated sound track is compared with the level of the 400-Hz signal. The difference between the two, in decibels, is the signal-to-noise ratio. Magnetic film or tape-recording systems are measured in a similar manner, with reference to a given value of distortion.

Amplifiers are generally measured with reference to 1 milliwatt (0 dBm). However, some manufacturers rate their products with reference to the maximum power output. (See Questions 12.207, 13.96, 13.97, and 17.139.)

23.131 ***How are noise measurement values added?***—To combine two electrical noise measurements is sometimes rather difficult. As a rule, the noise values are converted to noise powers, added, and then reconverted to dBa. (See Question 10.36). To eliminate such procedure, the chart in Fig. 1-88 may be used.

The decibel difference between the two known noise levels is located on the horizontal axis of the chart. The point on the curve on the vertical axis corresponding to this difference is the number of decibels to be added to the larger noise quantity to obtain the combined value of the noise.

As an example, if the amplitude of one source of uoise is 18 dBa and the second 23 dBa, the difference is 5 dBa. Referring to the chart for a difference of 5 dBa, 1.2 dBa is added to the larger value of noise which results in a combined noise level of 24.2 dBa. This same method of combining noise is also used for devices connected in tandem.

23.132 ***What is the equation for calculating the theoretical noise level of an amplifier when the bandwidth is known?***—The theoretical noise level is equal to:

$$dB = -198 + 10\ \mathrm{Log}_{10} R + 10\ \mathrm{Log}_{10}\ f_2 - f_1$$

where,

R is the output impedance in ohms,
$f_2 - f_1$ is the bandwidth in Hz at 5 dB down.

The reference level is 1 volt at the output. Thus, an amplifier with a bandwidth extending from 30 to 15,000 Hz with an output impedance of 600 ohms would have a theoretical noise level of:

$$\begin{aligned} &-198 + 10\ \mathrm{Log}_{10}\ 600 \\ &+ 10\ \mathrm{Log}_{10}\ 15{,}000 - 30 \\ &= -198 + 27.78 + 41.74 \\ &= -128.48\ \mathrm{dB/1\ volt} \end{aligned}$$

23.133 Show a simple method of increasing the signal-to-noise ratio of an amplifier.—The signal-to-noise ratio may be increased by connecting an attenuator of the correct configuration in the output circuit as shown in Fig. 23-133. The increase of the signal-to-noise ratio is possible because the internal noise of an amplifier is *constant*, although the signal varies in amplitude. As an example, suppose an amplifier is operating at an output level of 0 dBm. A 20-dB pad is connected in the output as shown. The insertion loss of the pad lowers the internal noise of the amplifier 20 dB. However, this also decreases the signal level by 20 dB. Now, to return the signal to its original level of 0 dBm will require that the amplifier be capable of operating at a level of plus 20 dBm without overloading. As can be seen, the signal-to-noise ratio was increased 20 dB because the signal level was increased 20 dB above the noise, which is at a constant level. Thus, if the amplifier originally had a signal-to-noise ratio of 40 dB, it has now been increased to 60 dB. The use of a pad in the output of a voltage amplifier is generally quite satisfactory as the output levels are low, and most small amplifiers will deliver more than a plus 20-dBm output signal. However, pads should not be used in the output of a power amplifier unless the amplifier is capable of producing a considerable amount of power.

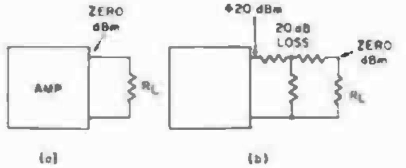


Fig. 23-133. Attenuator connected in the output of an amplifier to increase the signal-to-noise ratio. (a) No pad in output. (b) Pad in output.

It should be remembered that every time 3 dB of loss is inserted in the output of an amplifier, the power must be doubled to obtain the same output level that was available before the pad was connected. Thus, if a pad with 6 dB of loss is connected in the output of an amplifier, four times the power is required to compensate for the insertion loss of the pad and to bring the signal level to its original value.

23.134 What is the appearance of a signal with external or internal noise superimposed?—Figs. 23-134A to D show waveforms containing hum and noise;

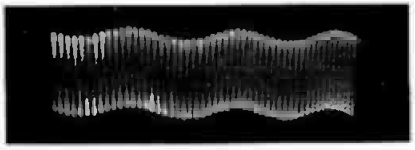


Fig. 23-134A. A 1000-Hz signal with 60-Hz hum modulation superimposed.

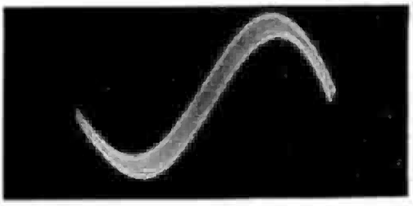


Fig. 23-134B. A single cycle of a 400-Hz waveform with noise superimposed. The width of the waveform is caused by the noise present.

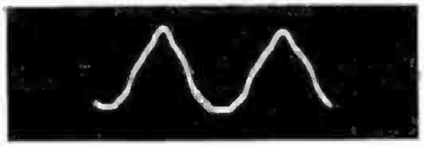


Fig. 23-134C. A 60-Hz hum pattern caused by cathode leakage in a low-level amplifier stage.

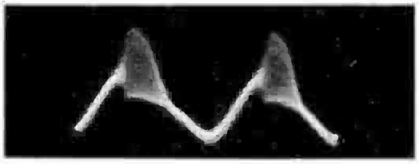


Fig. 23-134D. A 400-Hz waveform with high-frequency oscillation superimposed.

Fig. 23-134D shows a 400-Hz signal with a high-frequency oscillation superimposed.

23.135 How is a white-noise generator used to measure the frequency characteristics of a speaker?—It may be connected as shown in Fig. 23-135A. However, for these types of measurements a pink-noise filter is generally connected at the output of the white-noise generator to result in a straight-line response, when the response is measured using a constant-percentage octave-band analyzer.

A second method uses the white-noise generator, power amplifier, and speaker (Fig. 23-135B). The response is amplified and displayed on an oscilloscope. The microphone must be of known characteristics, otherwise the measurements may show peaks and valleys not actually in the speaker characteristics.

23.136 How may the noise and attenuation of a mixer control be measured?—The control is connected as shown in Fig. 23-136. As a rule, a well-designed mixer control will have a noise level of minus 120 to 140 dBm. As this is beyond the normal range of a vacuum-tube voltmeter, a decade amplifier will be required to extend the meter range. The noise may be monitored by using headphones and by connecting a very low level signal of 40 to 60 Hz to the input of the control.

When the noise is measured using the voltmeter, the input of the control is terminated with a shielded resistor and the noise generated is observed on the meter as the wiper arm passes over the control contacts.

Contact noise is best removed by burnishing the contacts by rapidly rotating the wiper arm. A light mineral oil such as Nujol will prevent the contacts from becoming oxidized.

When listening to contact noise, a signal is necessary to produce a minute current; otherwise, the noise cannot

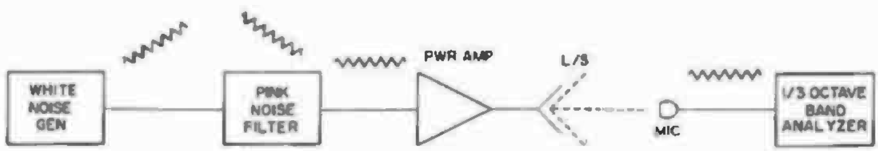


Fig. 23-135A. White-noise generator with pink-noise filter.

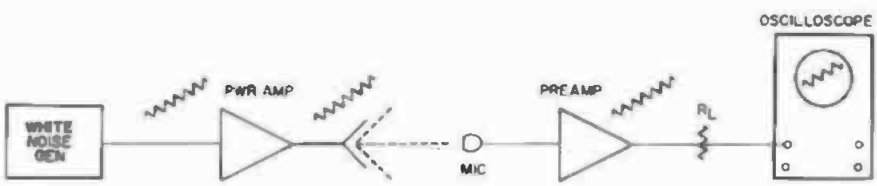


Fig. 23-135B. White-noise generator without pink-noise filter.

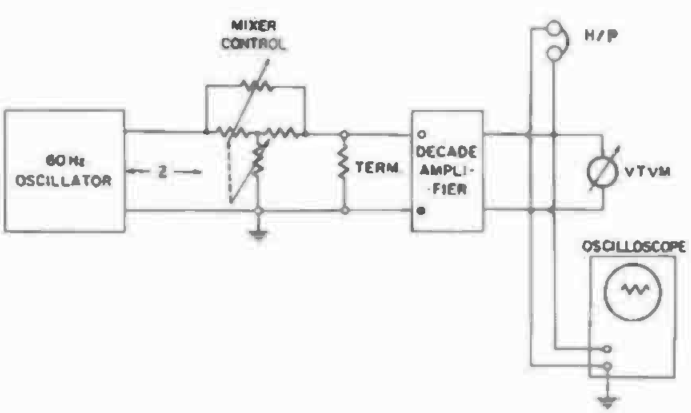


Fig. 23-136. Circuit for measuring or monitoring the noise of a mixer control.

readily be heard. A low frequency of 40 to 60 Hz is not very easily heard in the headphones; thus, the contact noise can be made to predominate. The attenuation per step may be measured by applying a signal to the input of the control and noting the change of level on the vacuum-tube voltmeter. (See Question 5.85.)

23.137 Describe the procedure for measuring input impedance.—Input impedance may be measured by several different methods, the simplest being the use of an impedance bridge. The bridge must be capable of being operated over a frequency range of 20 to 20,000 Hz, and the effects of extraneous magnetic fields reduced to a value where they have no effect on the measurements.

A second method that may be employed in lieu of an impedance bridge is given in Fig. 23-137A. A calibrated noninductive resistor or decade box is connected in series with the high side of the input circuit of an oscillator. The oscillator is set to a reference frequency of 1000 Hz. The measurement is started by first establishing a reference voltage of say 10 volts across load resistor R_L, with resistor R1 out of the circuit. Next, resistor R1 is adjusted to decrease the amplifier output reference voltage to 5 volts, or a decrease of 6 dB. The input impedance is then equal to the value of resistor R1. This is true because the input voltage is now equally divided across resistor R1 and the internal impedance of the amplifier input circuit. Therefore, the output voltage drops 50 percent, or 6 dB. Under these conditions the input resistance will be approximately 5 percent high, and the input impedance approximately 5 percent low. The same procedures are used for all frequencies of interest.

Fig. 23-137A. Circuit for measuring input impedance below 10,000 ohms.

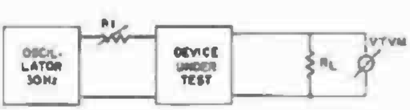


Fig. 23-137B. Circuit for measuring input impedance above 10,000 ohms.

For amplifiers having an input impedance of 10,000 ohms or greater, the circuit in Fig. 23-137B is used. Here a fixed resistor R1 of approximately the estimated input impedance is connected in series with the input, and a second resistor R2 in shunt with the input. The purpose of resistor R1 in this instance is to isolate the oscillator output from the amplifier, thus removing the effect of its low impedance. The reference voltage is established without R2 in the circuit. Then R2 is connected and adjusted to decrease the output reference voltage 50 percent (5 volts). To reduce the effects of hand capacitance and pickup from extraneous fields, the interconnecting wiring must be well shielded and a single ground point used if high frequencies and high-impedance inputs are under measurement.

The method of measuring input impedance, specified in Institute of High Fidelity (IHF) Standard A-201-1966, is similar to that shown in Fig. 23-137A, except that the procedure is slightly different. The amplifier is set for a reference level as previously explained, with R1 out of the circuit. Resistor R1 is then adjusted for a value that will permit the input voltage to be increased 21 dB (10.5:1 ratio) to return the output reference voltage to its original value. The input impedance is then the value of R1 divided by 10.

23.138 How can the internal output impedance of an amplifier be measured?—By the use of a calibrated resistor and voltmeter connected across the output of the amplifier as shown in Fig. 23-138A. A 400-Hz signal is applied to the input of the amplifier and the open-circuit voltage (unterminated) measured at the output. After a reference voltage is obtained, resistor R_L is connected across the output and its value reduced to obtain a voltage reading one-half that of the unterminated reference voltage. The internal output impedance is then equal to the value of resistor R_L in ohms.

A complete series of measurements is then made by applying different frequencies and measuring the impedance for each frequency. The results are plotted as shown in Fig. 23-138B, which is an actual measurement made on a push-pull amplifier using 14 dB of negative feedback and designed for a 16-ohm load termination. (See Ques-

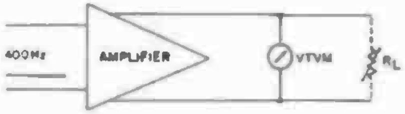


Fig. 23-138A. Circuit that is used for measuring the internal output impedance of an amplifier.

tion 12.142). It may at times be difficult to obtain a satisfactory impedance measurement using the method just described because of the dc resistance of the output winding of the transformer. When this occurs, the internal output impedance may be computed using the following formula:

$$Z_{out} = \frac{R_L \times E_{out}}{E_{RL}} - R_L$$

where,

R_L is the load impedance in ohms,
E_{out} is the unterminated output voltage,
E_{RL} is the voltage obtained with the specified load resistance.

23.139 How Is the internal output impedance of an amplifier using variable damping measured?—The damping control is set for several different values of damping and the internal output impedance measured as described in Question 23.138.

Fig. 23-139 is a plot of the internal output impedance of an amplifier employing variable damping in the output.

The distortion characteristics of an amplifier are also affected by the position of the damping control. Its effect is shown in Fig. 23-126.

23.140 How is the internal impedance of a speaker or system measured?—As a rule, the impedance characteristics of a speaker or system are measured in the enclosure used with the system. A simple method of measuring the impedance of any speaker or system is shown in Fig. 23-140A. A variable resistor R is connected in series with the output of the amplifier used to drive the speaker. If the speaker employs a crossover network, it is connected in the circuit as normally used, as in Fig. 23-140B. The common terminal of the vacuum-tube voltmeter is connected to the lower side of the measuring circuit and the high side of the meter is left free for connection to points A and B as the measurement progresses.

With a signal of 400 Hz applied to the input of the amplifier, the vacuum-tube voltmeter is connected to point A and the voltage is measured. The meter is then connected to point B and the

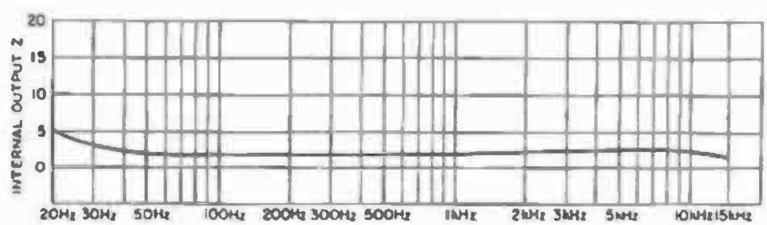


Fig. 23-138B. Internal output impedance of a negative-feedback, push-pull amplifier.

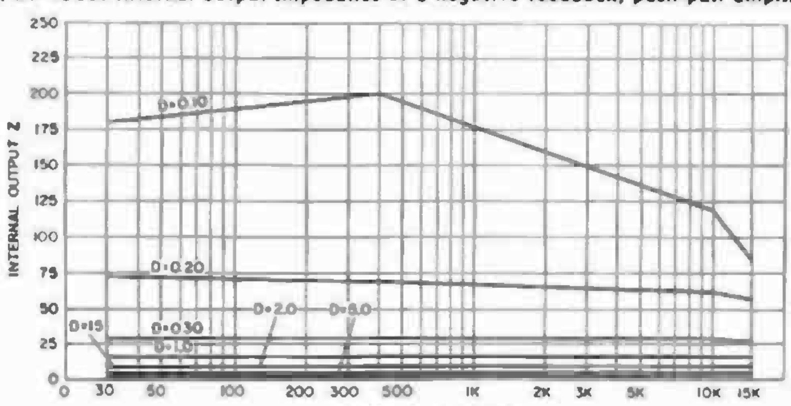


Fig. 23-139. Internal output impedance versus damping-control settings from 0.10 to 15.

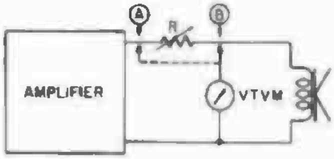


Fig. 23-140A. Circuit for measuring the internal impedance of a speaker system with a single speaker not using a crossover network.

resistor R adjusted for a voltage drop equal to one-half of the voltage measured at point A. (It may be necessary to switch the meter back and forth between points A and B several times until the voltage at point B is equal to one-half that at point A.) When the voltage at point B is equal to one-half the voltage at point A, the impedance of the system is equal to the value of R in ohms. The amplifier output impedance will in no way affect the measurement, as it is used only as a voltage source.

Measurements are made at frequencies of interest and plotted on semilog paper, as shown in Fig. 23-140C, or they may be plotted using 3 × 5 log-log graph paper. Referring to Fig. 20-89, it will be noted that a speaker presents only its rated impedance value over a small portion of the frequency spectrum; therefore, a uniform impedance cannot be expected. For best results, amplifiers used for driving a speaker or system with a low internal output impedance should employ a relatively large amount of negative feedback.

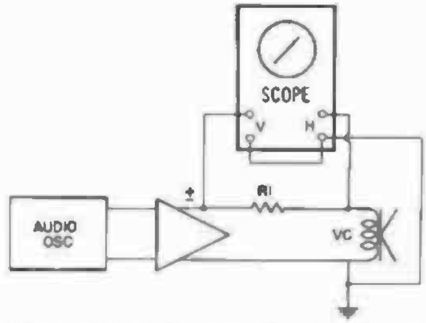


Fig. 23-140D. Visual method of displaying changes in loudspeaker impedance with frequency (after Crowhurst).

A visual method of observing impedance changes with frequency in a speaker may be accomplished by the use of the circuit in Fig. 23-140D. A frequency of 400 Hz is applied to the speaker through resistor R1, whose value is equal to the rated impedance of the speaker. The vertical and horizontal controls of the oscilloscope are adjusted for equal deflection of the beam in both the vertical and horizontal planes. If the impedance is equal to R1, the display will be a 45-degree diagonal line. If it is not, the oscillator frequency is adjusted until a 45-degree line is obtained (the line may tilt in either direction). Note the frequency and vary it above and below the reference frequency. As the impedance of the speaker decreases, the display will shift toward the vertical axis. With an increase of impedance, it shifts toward the horizontal axis. As long as the im-

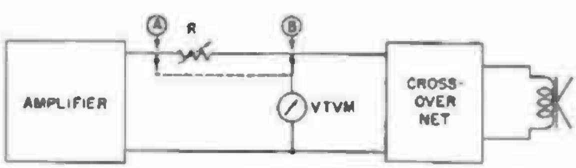


Fig. 23-140B. Circuit for measuring the internal impedance of a speaker system with a crossover network.

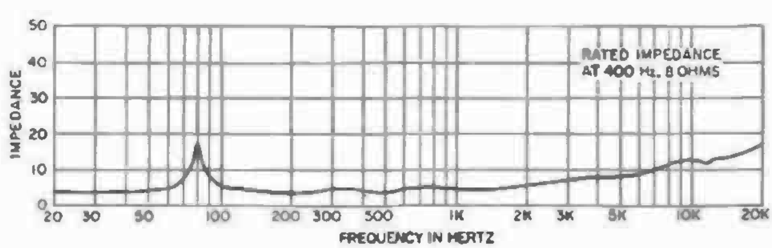


Fig. 23-140C. Impedance variation of a 10-inch single-cone speaker housed in an open-back cabinet.

pedance remains equal to R1, the trace remains at a 45-degree angle. When the impedance becomes reactive, the display becomes an ellipse. (See Question 20.103.)

23.141 ***Describe a sweep record and how it is used.***—The record is played back through the system to be tested,

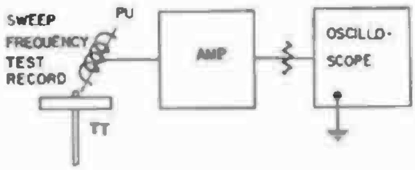


Fig. 23-141A. Method of using sweep-frequency record.

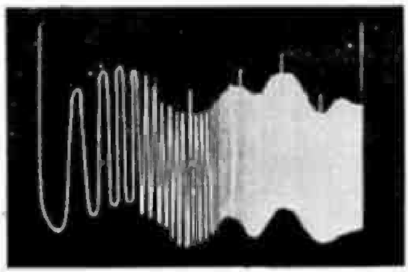


Fig. 23-141B. Frequency response pattern with a turntable that has excessive vibration.

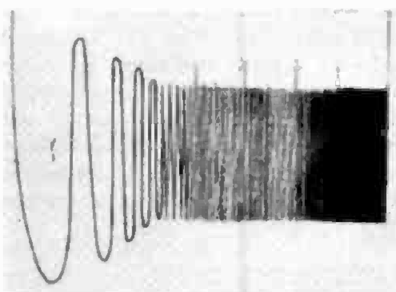


Fig. 23-141C. Frequency response when low frequencies have been accentuated.

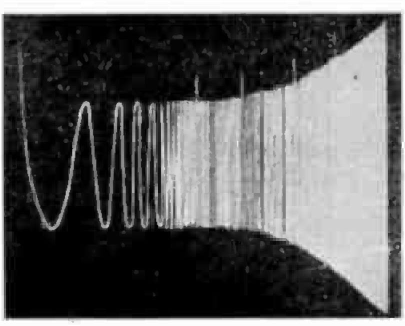


Fig. 23-141D. Frequency response when high frequencies have been accentuated.

with an oscilloscope connected at the output, as shown in Fig. 23-141A. Sweep-frequency records, as explained and illustrated in Question 16.49 cover a frequency range from 50 to 15,000 Hz and are swept over this band at a rate of 20 times per second. The frequency characteristics of such records are the standard RIAA reproducing characteristics (Fig. 13-95).

The record is played back using the normal system equalization and, if equalized for use with the RIAA recording characteristic, the display on the oscilloscope should be uniform within plus or minus dB over a frequency range of 50 to 15,000 Hz. If not, the equalization is adjusted until a uniform characteristic is obtained.

Variations in the frequency response may be computed in decibels:

$$dB = 20 \text{ Log}_{10} \frac{E_1}{E_2}$$

where,

E_1 is the reference frequency height displayed by the oscilloscope.
E_2 is any frequency of interest.

The turnover frequency of the system must be the same as that of the sweep-frequency record. This information is supplied by the manufacturer of the record.

Various frequency-response patterns for different types of tests are shown in Figs. 23-141B to D. Fig. 23-141B shows how the response appears on the oscilloscope when a turntable has excessive vibration, although the frequency response of the system is flat. In Fig. 23-141C the low frequencies have been accentuated, and in Fig. 23-141D the high frequencies have been accentuated.

Sweep-frequency records are, in some respects, similar to the sweep-frequency generator described in Question 22.53, because similar response patterns are obtained.

23.142 ***What is the procedure for measuring the frequency response of a pickup?***—The first step is to ensure that the arm alignment and stylus angle are correct. The output of the pickup is terminated in its specified load resistance and played back through an equalized preamplifier, using one of the several standard test records. The amplifier equalization must meet the standard reproducing response as specified by the RIAA Standard shown in Fig.

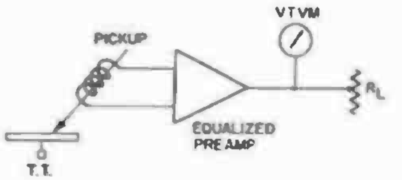


Fig. 23-142A. Measuring the characteristics of a pickup using an equalized preamplifier.

13-95. Frequency-response measurements are then made by establishing a reference frequency, and all other frequencies are plotted with reference to this frequency. The instructions for a given test record must be adhered to, since the method in which they are recorded varies somewhat. The test circuit appears in Fig. 23-142A.

If the actual characteristics of the pickup are to be measured, the preamplifier is omitted and the vacuum-tube voltmeter is connected across the load resistor (Fig. 23-142B). As a rule, only the first circuit is necessary. The test records most commonly used are the CBS STR-100 and the CBS STR-100-111.

23.143 How is stylus force (vertical pressure) determined?—The average high-quality phonograph pickup requires a stylus force (vertical pressure) ranging from 0.75 to 1.5 grams. The pickup is tested with an intermodulation test record and analyzer. The correct pressure is that pressure which results in the lowest intermodulation distortion in keeping with good tracking.

The amount of intermodulation distortion is not a true reading of the pickup, but rather a combination reading of record, pickup, and preamplifier. If an intermodulation test record is not available, the stylus force may be adjusted using a record similar to the CBS STR-100 or CBS STR-100-111, or by the use of an N-A record, described in Question 23.77. The instructions for the test record must be adhered to for best results. An oscilloscope should be connected across the load resistance for observing the waveform.

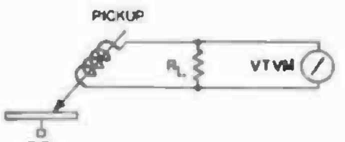


Fig. 23-142B. Measuring the characteristics of a pickup without preamplifier.

23.144 How may the actual voltage output of a pickup be measured?—If the pickup is designed only for monophonic reproduction, a test frequency of 1000 Hz, recorded at a peak velocity of 7 centimeters per second is used. The output voltage is measured across the load terminations as shown in Fig. 23-142B. If it is a stereophonic design, both sides are terminated and the voltage is measured across each side (Fig. 23-144). In this instance the test frequency is also 1000 Hz, but recorded at a peak velocity of 5 centimeters per second.

23.145 How may a reproducer stylus be tested for wear?—By the use of a special intermodulation or the CBS STR-100 test record. A typical measurement shown in Fig. 23-145 was made using a new and a worn sapphire stylus, and record diameter was plotted versus intermodulation distortion. It will be observed the distortion products rise quite rapidly for diameters less than 8.5 inches. The test frequencies used were 100 and 7000 Hz, mixed in a ratio of 4:1, the lower frequency being 12 dB greater in amplitude than the higher frequency.

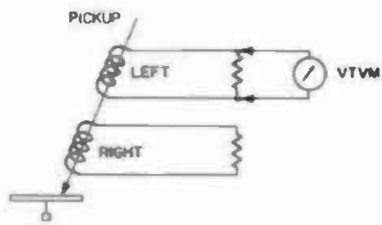


Fig. 23-144. Circuit for measuring the voltage developed by a stereophonic pickup, using a standard test record.

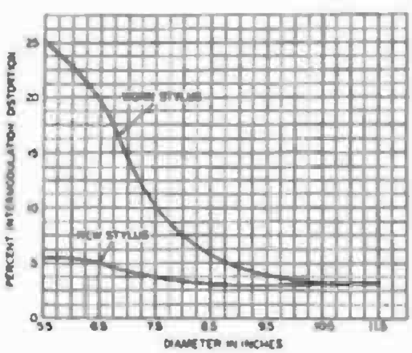


Fig. 23-145. Intermodulation distortion caused by stylus wear.

23.146 How is pickup-arm resonance measured?—Test records such as the CBS STR-100 and STR 100-111 both have bands for testing pickup-arm resonance. Since pickup resonance is a function of both the pickup cartridge and the arm, this test is a combination test of both components. Pickup-arm resonance is of low-frequency character, being in the range of less than 10 Hz. The circuit for making such tests is given in Fig. 23-146. Test record band consists of a glide-tone sweeping downward from 200 Hz to 10 Hz. Arm resonance is indicated by an excessive indication of the meter, generally accompanied by a rattling noise as the stylus tends to momentarily leave the groove walls. It is possible by changing to a different pickup for the resonance frequency to change or disappear completely. Pickup arms should be selected with the resonant frequency below 10 Hz if possible.

If for testing purposes frequencies below 10 Hz are required, they may be generated by playing a test record of 33⅓ rpm at reduced speeds. A 33⅓-rpm recording of 40 Hz played back at a speed of 18 rpm becomes a frequency of 2.16 Hz as given below.

$$F_1 = \frac{f}{T_1/T_2} = \frac{40}{33.3/18} = \frac{40}{1.85} = 2.16\ \text{Hz}$$

where,

F_1 is the new frequency,
f is the known frequency,
T_1 is the normal speed of the test record in rpm,
T_2 is the new speed in rpm.

The 300-Hz low-pass filter is not absolutely necessary; however, if one is available it will be of help in eliminating the effects of surface noise from the measurements.

23.147 How is the sensitivity of a pickup to extraneous magnetic fields measured?—By the method similar to that described for measuring the effectiveness of magnetic shielding in Question 23.81. The pickup to be measured is terminated in its normal load resistance and placed in the center of the field helix. The effect of the field on the pickup unit is measured at the output terminals of the pickup with a vacuum-tube voltmeter.

This type of measurement is very helpful for comparing the effect of magnetic fields on pickups of different design and manufacture. The foregoing measurement is a comparative one only. To determine the absolute level of pickup for a given unit under measurement, the strength of the magnetic field, as well as several other factors, must be known.

23.148 How is stylus force of a pickup measured using a scale?—By suspending the front end of the pickup (mounted in its arm) by means of a small spring scale calibrated in grams. A scale calibrated from 0 to 10 grams will be satisfactory, because most modern pickups use a vertical force of 0.5 to 1.5 grams.

Several small scales are available which are designed to be placed on the record surface, with the pickup stylus resting on a spring. The weight of the pickup depresses the spring and the pickup weight is indicated in grams on a calibrated weight scale at the rear of the spring.

23.149 What is the procedure for making flutter measurements? — Two methods are used for the measurement of flutter, the method depending on the type of equipment to be measured.

If the device is a magnetic-tape recorder, a 3000-Hz signal of constant amplitude with fairly low harmonic distortion is recorded, then played back, and the total rms percent flutter measured using a flutter meter or bridge as described in Question 22.41. The connections are shown in Fig. 23-149A. If the machine includes a separate play-

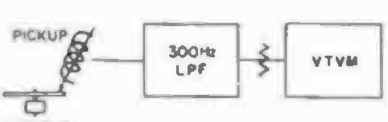


Fig. 23-146. Circuit for measuring pickup arm resonance using a low-frequency test record.

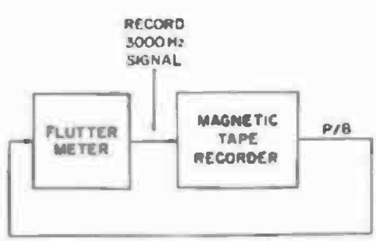


Fig. 23-149A. Circuit for measuring the percentage flutter of a magnetic tape or film recorder.

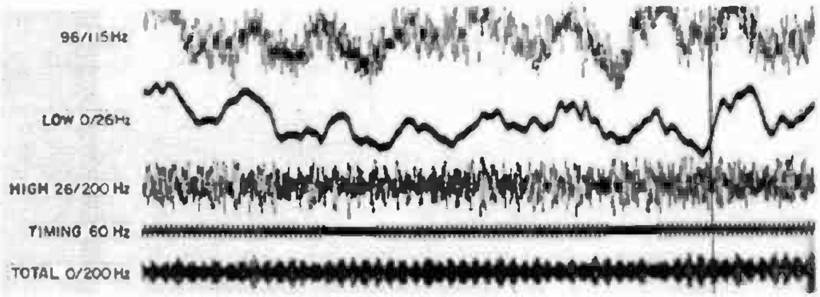


Fig. 23-149B. A typical fluttergram made using an automatic recorder in conjunction with a flutter meter.

back head, the percentage of flutter may be measured as the 300-Hz flutter signal is recorded.

For measuring the flutter of a photographic film recorder, a 3000-Hz sound track is recorded and the negative processed in the prescribed manner for the system of recording. The negative is played back and the percent flutter measured on a sound head for which the flutter has previously been measured. This latter measurement includes both the flutter of the playback sound head and the recorder. The percent of flutter of the sound head is subtracted from the measured flutter. As a rule, a professional film recorder will have about 0.05 percent of flutter.

The flutter of a disc-recording machine is measured by recording a 3000-Hz signal, and then playing it back on the same machine or one one of known flutter characteristics. Special disc records for measuring flutter are available from several manufacturers of test records. Motion-picture projection equipment is measured using a special 3000-Hz flutter test film. (See Question 19.63.)

For any type of flutter test, the use of flutter-measuring equipment is similar to that described for the magnetic recorder. Either the signal is recorded and played back or the flutter is measured using a special flutter tape, film, or disc. The output of the equipment under measurement should be properly terminated and run at a level which will not cause overload in any form.

Many times it is desirable to know in which frequency band flutter is being generated, such as in the transport or drive system. For this type of measurement, the flutter signal is broken down into several different frequency bands by the use of a flutter meter as discussed in Question 22.41.

In this instrument the flutter is broken down into four bands covering 0.5 to 30 Hz, 30 to 300 Hz, 300 to 5000 Hz, and direct current to 5000 Hz. Thus, if the flutter is being generated by a capstan, sprocket, tape contact, magnetic head, gearbox, or motor, it may be isolated by noting the flutter rate.

The first instrument described in Question 22.41 is quite satisfactory for the maintenance of existing equipment, while the second instrument is used for development work.

Fig. 23-149B shows a typical fluttergram recorded on an automatic recorder. The various frequency bands in which the flutter was measured are indicated on the left-hand margin. The total rms flutter shown is 0.15 percent.

23.150 In what terms are flutter meters calibrated?—As a rule, flutter meters are calibrated to read the rms value; however, some of the older types read peak-to-peak values. The latter reading may be converted to rms by multiplying the reading by 0.707.

For many years flutter meters (or bridges) have been calibrated to read in terms of rms voltage, which is quite satisfactory for audio-frequency use. However, for instrumentation recorders and reproducers, peak-to-peak readings are generally employed. Actually, a true rms value of flutter is quite difficult to obtain, since the flutter signal contains a dc component, noise, and sine-wave components. As a rough approximation, the rms value can be assumed to be one-sixth to one-fourth the measured peak-to-peak value. (See Question 22.41.)

The standards are clearly defined in the USASI (ASA) Standard Z57.1-1954.

23.151 How is the frequency response of a light valve or galvanometer measured?—Three methods are available. The first method uses a constant-amplitude signal of different frequencies which is applied to the recording amplifier. The deflection of the light valve or galvanometer is then observed by means of a calibrated monitor card or periscope placed in the optical system of the recorder.

In the second method the peak frequency of the light modulator is found by sweeping an oscillator across the recording band. If the light modulator shows a rise at the high frequencies, the gain of the recording system is lowered an amount equal to or slightly less than the rise at the peak frequency. Frequencies of constant amplitude are then recorded, starting at 40 Hz. The negative sound track is processed, and the amplitude of the recorded frequencies is measured and then compared to a reference frequency of 1000 Hz by means of a calibrated microscope.

Lowering the gain of the recording system by an amount equivalent to the rise at the peak frequency of the light modulator will prevent the modulator from being overloaded when the peak frequency is approached.

The third system of measurement requires the use of a gain set or calibrated attenuator. A reference frequency of 1000 Hz is established for 100-percent modulation of the light modulator, by observing the deflection by means of a monitor card or periscope in the optical system. After the reference level has been established, 2000 Hz is applied. If the modulator shows a rise, loss is inserted in the gain-set attenuators to bring the modulator back to the reference frequency level. A 3000-Hz frequency is now applied and the deflection noted. If the deflection continues to rise, loss is again inserted to return the modulator to the reference level. This procedure is continued up to the highest frequency of interest.

The final frequency response is plotted by noting the increase, or decrease, in the amount of attenuator loss required to maintain a constant deflection of the light modulator.

If the loss of the attenuators increases with frequency, the light modulator has a rising characteristic. Conversely, if the loss decreases, a loss at the higher frequencies is indicated.

23.152 How is the harmonic distortion of a light modulator measured?—By deflecting the modulator to a given percent of modulation and then measuring the harmonic distortion by means of a photocell situated in such a manner that a portion of the modulated light beam is deflected to the photocell for measurement.

It is obvious that, in a measurement of this nature, a certain amount of harmonic distortion is contributed by both the photocell and its amplifier. However, if the photocell and amplifier are properly designed, the distortion added ot the measurements is generally quite small and may be ignored, as the distortion from the light modulator as a rule is on the order of 1.0 to 1.5 percent.

23.153 How is the harmonic distortion of a film-recording channel measured?—The distortion in terms of either harmonic or intermodulation distortion may be measured by applying the signal frequency (or frequencies) to the input of a recording channel. A sound track is recorded at 80-percent modulation and a print made at the optimum density. This sound track is played back from a sound head, and the distortion is read. The sound-head distortion may be measured by the use of a special harmonic-distortion measuring film.

23.154 How is the square-wave response of an amplifier measured? — The square-wave generator is connected to the amplifier as shown in Fig. 23-154A. One precaution should be observed: The output of the square-wave generator cannot be sent through the attenuator or send section of a gain set because of the wide frequency band of a square wave. Also, the gain set might include a coil which would affect the square-wave form. Interconnecting cables must be of a type that will not affect the frequency response, and any attenuator networks used to attenuate the signal before it is applied to the input of the amplifier must be designed to have a minimum of leakage to ground.

To reproduce a fundamental square wave without distortion, the amplifier must be capable of passing frequencies of $f/10$ and $f \times 10$. This means that, if a fundamental square wave of 100 Hz

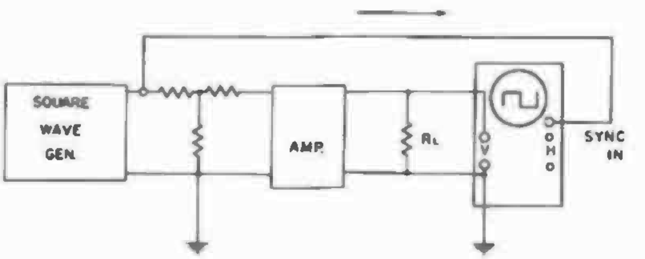


Fig. 23-154A. Measuring the response of an amplifier to a square-wave signal applied to the input.

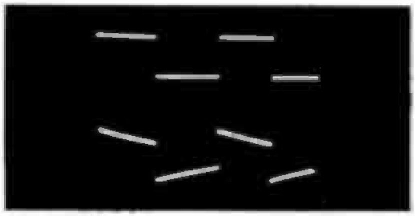


Fig. 23-154B. Reproduction of a 40-Hz fundamental square wave by a 20-watt amplifier. Slant of output display indicates good low-frequency response but with phase shift.

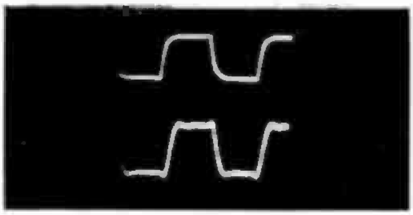


Fig. 23-154C. Reproduction of 10-kHz fundamental square wave by a 20-watt amplifier. Output waveform indicates slight ringing.

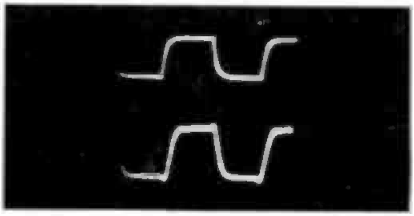


Fig. 23-154D. Reproduction of a 10-kHz fundamental square wave by a 20-watt amplifier. Shape of output waveform indicates excessive high-frequency response with slight ringing.

is applied to an amplifier, the amplifier must have a bandwidth of 10 to 1000 Hz. If a fundamental frequency of 1000 Hz is applied, the bandwidth must be 100 to 10,000 Hz. Therefore, the fundamental frequency must be selected with care.

To ensure that the square wave is not being distorted by external equipment, a noninductive resistor must be used for the load termination. The oscilloscope used for displaying the waveform must be capable of passing a bandwidth wider than the bandwidth to be measured.

Many of the older and smaller-type oscilloscopes will not satisfactorily pass a square wave without considerable distortion. Therefore, before starting the measurement, check the square-wave response of the oscilloscope.

The images in Fig. 23-154B, C, and D show square-wave images obtained at the output of a 20-watt tube amplifier. Such displays may be obtained by the use of an electronic switch, as described in Question 23.184, or a dual-trace oscilloscope. The upper trace is the input signal from the square-wave generator and the lower trace, the output of the amplifier. It will be observed the input signal shows a rounding off at the leading edge. This can be due to the frequency response of the oscilloscope or due to the shunt capacitance of the interconnecting cables. This is a good illustration of why the input signal must be monitored for comparison with the output signal of the amplifier.

Fig. 23-154E shows typical square-wave patterns with an interpretation of their shapes.

The transient characteristic of a circuit is defined as the output waveshape which results from a suddenly applied voltage at the input. The transient characteristic is limited by the ability of the

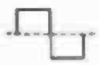
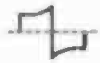
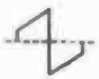
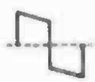
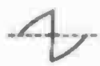
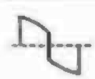
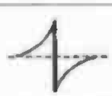
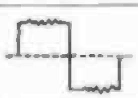
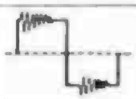
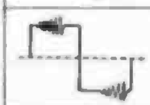
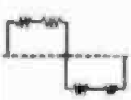
Waveform	*LF Gain*	*LF Delay*	*HF Gain*	*HF Delay*	*Damping*
	Ideal	Ideal	Ideal	Ideal	Ideal
	Insufficient	Good	Excessive	Good	High
	Excessive	Good	Insufficient	Good	High
	Good	Excessive	Good	Insufficient	High
	Good	Insufficient	Good	Excessive	High
	Excessive	Excessive	Insufficient	Insufficient	High
	Excessive	Insufficient	Insufficient	Excessive	High
	Insufficient	Excessive	Excessive	Insufficient	High
	Good	Good	Excessive	Good	Medium
	Good	Good	Excessive	Good	Low
	Good	Good	Excessive	Good	Negligent
	Good	Good	Sharp Cutoff or Peaked	Good	Low

Fig. 23-154E. Square-wave images.

circuit to transmit both high- and low-frequency signals simultaneously without distortion. The high-frequency response will determine the shape of the transient for a short time after its application, and the low-frequency response will determine the waveshape after a longer time has elapsed. In most instances, a square-wave test using two different frequencies will suffice. However, the high-frequency test should be high enough to eliminate low-frequency effects; and the low-frequency test should be low enough to include them.

Circuit damping is indicated by the shape at the top of the rise of the waveform. A circuit which is highly damped is indicated by a waveform which approaches its final maximum without overshooting, whereas an underdamped circuit will manifest itself by producing a train of oscillations after the initial rise. The natural frequency of the circuit producing the oscillations may be determined approximately by counting the number of maxima that occur in any one cycle.

Insufficient delay at the high frequencies is indicated by the rounding off of the corners of the waveshape diagonally opposite each other. If the circuit presents the same delay for all frequencies, this will be indicated by all four corners of the waveshape being the same. If the output waveshape shows signs of oscillation, it is always the result of a circuit containing frequency and delay characteristics which change sharply.

An output waveform which is symmetrical with time but whose positive and negative excursions are unsymmetrical indicates nonlinear distortion. If the amplitudes of the waveform at the top and bottom are unequal, nonlinear distortion is also indicated. Saturation of a component in the circuit is indicated by oscillation in one-half of the waveform.

If a perfectly square waveform is passed by the device under test, the device has excellent transient, frequency-response, and distortion characteristics.

23.155 ***Describe the composition of a square wave.***—A square wave is composed of a fundamental frequency and an infinite number of odd-harmonic frequencies. The even harmonics (2, 4, 6, 8, etc.) as equal to zero; only the odd harmonics are contained in the waveform as may be seen in Fig. 23-155. These illustrations have been plotted to show the different harmonic frequencies and how they combine to produce a square waveform (indicated by the straight lines).

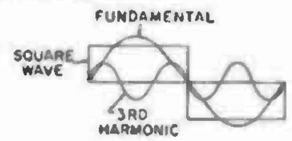
When a square wave is passed through a device, the bandwidth is the limiting factor and determines the number of harmonic frequencies that will be passed. The greater the number of odd harmonics that are passed, the more rectangular will be the waveform.

The upper frequencies depend on the fastest change occurring in the waveform. The lowest frequency depends on the repetition rate of the waveform.

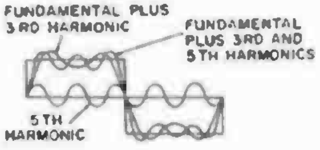
23.156 ***At what level are square wave measurements made on an amplifier?***—At a level well below the overload point of the amplifier. Due to the nature of square waves, an amplifier can be readily overloaded and the square wave response made to appear better. (See Questions 22.103 and 22.104.)

23.157 ***How should the fundamental frequency of a square wave be selected for testing an amplifier?***—As a rule, two fundamental frequencies are used. One frequency must be low enough to indicate phase shift and frequency attenuation at the low frequencies, and the second must be high enough in frequency to show the attenuation of the high frequencies. (See Fig. 23-154E.)

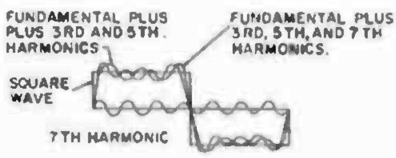
Square-wave testing is quite useful in certain types of production testing. The frequency response of a device may be checked by the application of a square wave observed on a calibrated oscilloscope graticule.



(a) *Fundamental frequency and 3rd harmonic.*



(b) *Fundamental frequency plus 3rd and 5th harmonics.*



(c) *Fundamental frequency plus 3rd, 5th, and 7th harmonics.*

Fig. 23-155. Development of a square wave from sine waves.

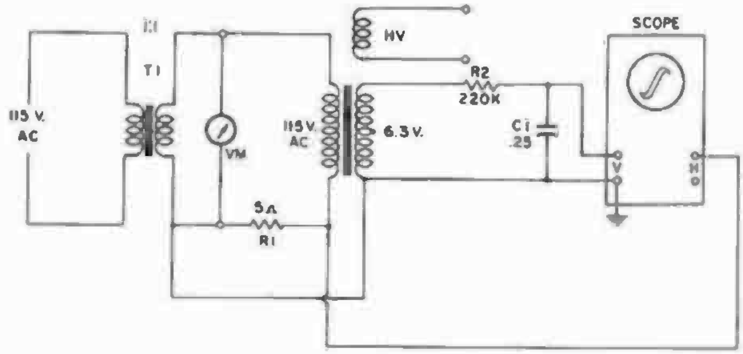


Fig. 23-158A. Circuit for obtaining a hysteresis or B-H curve of a transformer.

23.158 ***How are the hysteresis characteristics of a power transformer measured?***—A circuit suitable for measuring the hysteresis characteristics of a power transformer, using an oscilloscope to trace a B/H curve, is shown in Fig. 23-158A.

To obtain such a pattern, two voltages are necessary. One voltage is taken from across resistor R_1 connected in series with the primary of the transformer. The voltage drop across this resistor is proportional to the current in the primary and, therefore, proportional to the magnetizing force. This voltage is applied to the horizontal plates of the oscilloscope.

The second voltage is taken from the output of an integrating network consisting of resistor R2 and capacitor C1 connected across the 6.3-volt secondary of the transformer. The voltage at the output of the network is proportional to the turns ratio of the primary to the secondary as well as to the number of lines of force produced and the ratio of change of the magnetic lines of force. This voltage is applied to the vertical plates of the oscilloscope.

Because one side of the ac line is grounded, it is advisable to connect an isolating transformer between the line and the primary of the transformer under test, thus permitting the oscilloscope to be grounded.

When the controls of the oscilloscope are properly adjusted, the pattern will appear similar to that of Fig. 23-158B, which is a pattern of the transformer operating without a load. The pattern of Fig. 23-158C is with a normal load current and that in Fig. 23-158D is with an overload. The power loss under these conditions would be excessive.

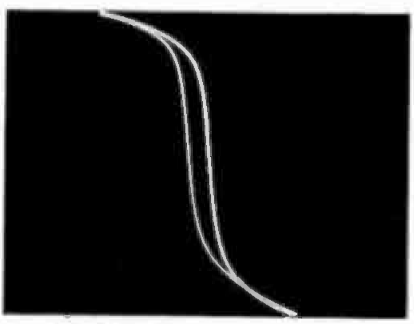


Fig. 23-158B. Hysteresis curve of an unloaded power transformer.

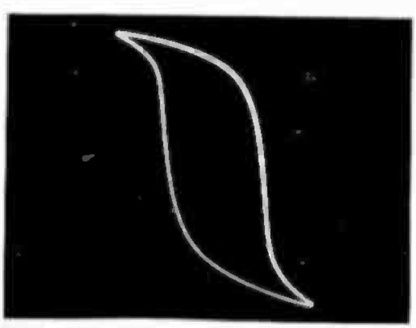


Fig. 23-158C. Hysteresis curve of a power transformer operating with a normal load.

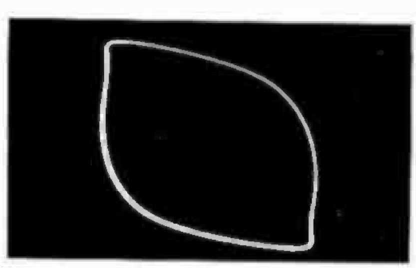


Fig. 23-158D. Hysteresis curve of an overloaded power transformer.

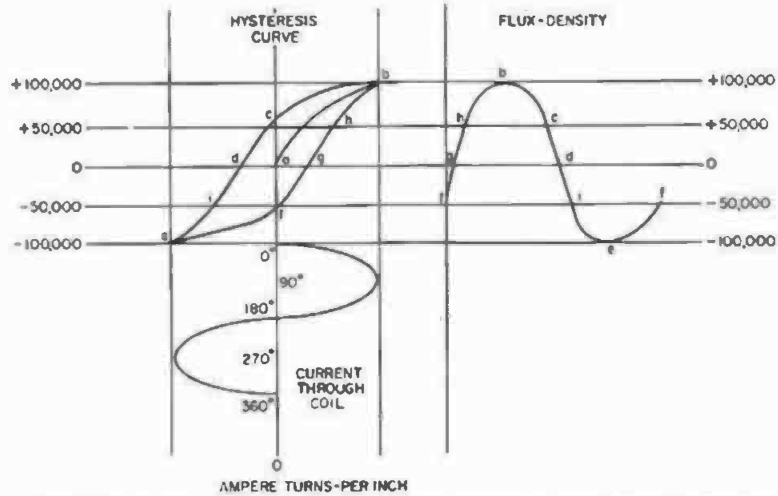


Fig. 23-158E. Hysteresis curve versus ampere turns for electric metal core iron.

The circuit shown is useful for demonstrating different methods of stacking transformer cores and, also, the characteristics of different types of core materials.

Fig. 23-158E shows a typical magnetization curve for annealed sheet steel, commonly called electric metal, the material generally used for power-transformer cores and for laminations in the construction of motors and generators. A sample of the material to be tested is placed in a coil of a given number of turns. The current through the coil is varied to obtain the required ratio of ampere turns.

The graphical plot shown is the magnetomotive force expressed in ampere turns versus the flux density (gauss) per square inch. A hysteresis loop is the graphical presentation of this phenomenon. Inspection of the curve shows the flux density to be about 95,000 lines of force per square inch. When the ampere turns for this particular metal are increased (current increased through the coil) from 20 to 40, the density is increased to 108,000 lines of force per inch. Increasing the ampere turns increases the flux density, but at a lesser rate. This is indicated by the fact that, if the ampere turns are increased four times, or to 160, the number of lines per square inch increases only to about 120,000.

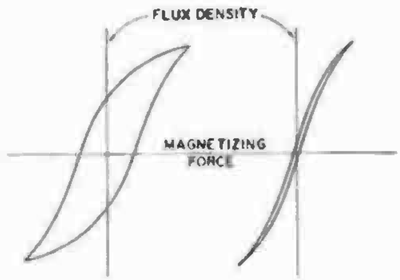


Fig. 23-158F. Comparison of a typical hysteresis loop and the narrow-band loop found in grain-oriented steel cores.

The curve shown is not indicative of all magnetic metals but varies with different alloys. The portion of the curve where it begins to flatten out is the saturation point of the metal and is the point where an increase in current produces little or no change in the lines of force produced.

Hysteresis is the lagging of the magnetizing effect behind the magnetizing force. If a given material is subjected to a magnetizing force and if the material is in a neutral state (not magnetized), the flux density will increase as shown by the upper right-hand curve a-b (0 to 90 degrees on the sine wave) on the hysteresis curve. If the magnetizing current is reduced at point b, the core will start to lose its magnetization and will follow curve b–c. At c the force becomes zero (180 degrees on the sine wave). However, a closer inspection of the curve reveals that the flux density is still 50,000 lines, even though the magnetizing force has dropped to zero. This hysteresis or lagging is caused by the magnetism retained by the core material. If the magnetizing force is reversed in polarity, the material loses its magnetism as shown by the curve c-d.

Now, if a negative-magnetomotive force is applied at d and increased to a value equal to that at b, the core material will magnetize in the opposite direction, curve d-i-e. Under these conditions, the flux density at e will equal the flux density at b. Decreasing the negative magnetomotive force to zero varies the flux density as shown by curve e-g. At g, the flux density is zero again. The minus sign on the flux value is used only to indicate that the direction of the field below point e is opposite to the direction of the field at point b.

Applying a positive magnetomotive force and gradually increasing it will cause the material to lose its magnetism along curve f-g-h, becoming completely demagnetized at point g and remagnetizing along curve g-h-b, thus completing the hysteresis-loop pattern.

The area within the loop is proportional to the amount of work done against the residual magnetism and dissipated by the transformer core material. Therefore, the hysteresis-loop characteristic of a transformer is an important factor of its design.

The magnetomotive force required to overcome the residual magnetism of the core material represents a loss of power and is referred to as the hysteresis loop. When the loop area is small, the power losses are small; when large, the power losses are large. This is illustrated by the patterns in Figs. 23-158B, C, and D. (See Question 17.142.)

In testing some of the newer metals, particularly the grain-oriented steels, the two curves are much straighter as may be seen by the comparison of a conventional electric metal and a grain-oriented steel in Fig. 23-158F.

23.159 How may core samples be tested for hysteresis characteristics?—In a manner similar to that used for measuring the hysteresis characteristics of a transformer as explained in Question 23.158.

Resistor R1 shown in series with the core sample in Fig. 23-159 will vary in value from 1 to 10 ohms, depending on the current through the primary winding. To prevent damage to the oscilloscope, a 1:1 isolation transformer is connected between the output of the autotransformer and the core sample, thus permitting the cathode-ray oscilloscope to be properly grounded.

23.160 Describe a method for measuring the internal noise of a capacitor.—The internal noise of paper, oil, mica, metalized, or any other type of capacitor except the electrolytic may be mea-

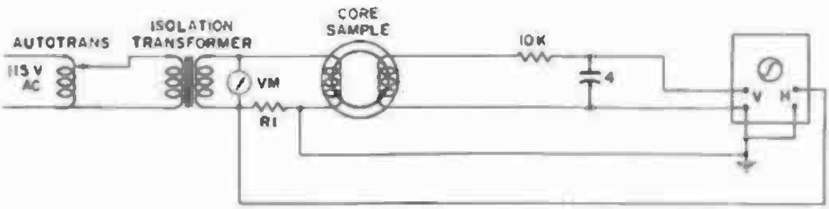


Fig. 23-159. Circuit for measuring the hysteresis characteristics of core-iron samples.

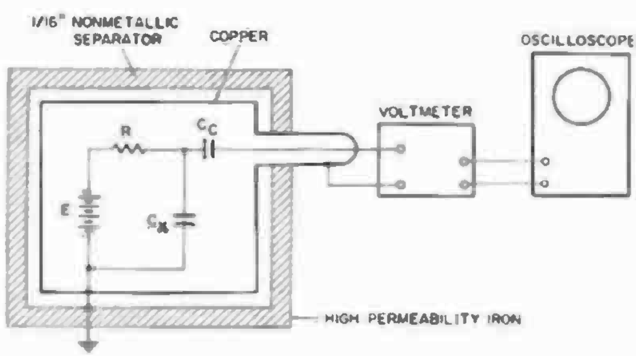


Fig. 23-160. Circuit for measuring the internal noise of a capacitor. The battery may be replaced with a noise-free and well-regulated power supply.

sured using the circuit of Fig. 23-160. The battery voltage shown in the shielded box is made equal to the normal working voltage of the capacitor. In actual practice the battery is replaced with a noise-free, well-regulated power supply with rf chokes and filtering capacitors in or near the shielded container. Series resistor R must be noise free and is approximately 1 megohm in value. Coupling capacitor C_c must also be of the noise-free variety. The capacitor to be tested for noise is connected at C_x. All elements of the test circuit must be mounted within the shield, consisting of a copper electrostatic shield and a high-permeability iron shield. The cable between the shielded box and the voltmeter is grounded at both ends as shown. The meter used must have a capability of measuring up to 1 megahertz and down to 1 millivolt, with an output connection for connecting an oscilloscope to display the characteristic of the noise.

A satisfactory capacitor will show little or no movement of the meter, even in its most sensitive position. Nested shielding is discussed in Questions 8.50 and 8.51.

23.161 ***How may a power transformer be tested for shorted turns?***—A 115-volt, 10-watt lamp is connected in series with the primary of the suspected transformer. With no load on the secondaries, the lamp will show a dull red glow. If there are no shorts in the windings, the lamp will show a considerable change in color when a short is applied to one of the secondaries. If a short is applied to one of the secondaries and shorted turns are present in one of the other secondaries, the lamp will show little or no change in color. The primary may be tested using the same procedure.

23.162 ***How is the effectiveness of transformer magnetic shielding measured?***—Audio transformers employing nested shields or plain metal cases are often rated in decibels relative to the effectiveness of the magnetic shield surrounding the coils. Such statements imply that the case surrounding the coils is capable of reducing the effect of an external magnetic field a given number of decibels relative to a given magnetic field intensity. Thus, a transformer stated to have a 90-dB magnetic shield means that the intensity of the surrounding field will be reduced 90 dB before reaching the internal windings of the transformer.

The transformer to be tested is removed from its case and placed in the center of a field coil of 8 to 10 turns to which is connected a source of 60-Hz current, as shown in Fig. 23-162. The transformer windings are terminated in their normal load impedances using noninductive resistors. A vacuum-tube voltmeter is connected across one winding.

The current through the field coil is adjusted for a given field intensity. The voltage induced in the transformer is read on the vacuum-tube voltmeter. The transformer is then returned to its case and a second measurement made at the same field intensity. The effectiveness of the shielding may then be computed in decibels by:

$$dB = 20 \operatorname{Log}_{10} \frac{E_1}{E_2}$$

where,

E_1 is the voltage measured without the shield,

E_2 is the voltage measured with the shield.

Transformers requiring extensive magnetic shielding make use of nested-shield construction as described in Question 8.50.

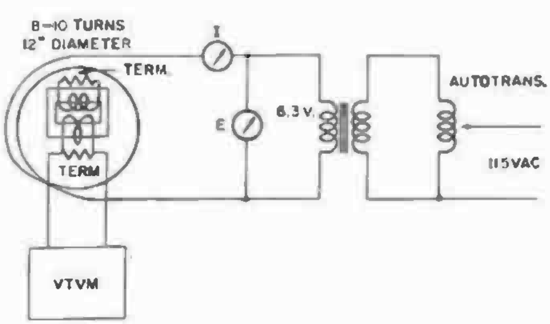


Fig. 23-162. Circuit for measuring the effectiveness of a magnetic shield, using a magnetic field of known intensity.

23.163 *What is the procedure for measuring power-supply ripple voltage?* —The measurement of the actual ripple voltage from a power supply is rather difficult because of the nonsinusoidal nature of the ripple. Generally the latter is read on a true-rms-responding voltmeter or a cathode-ray oscilloscope. In the latter instance the voltage is determined by using an internal or external voltage calibrator. As a rule, ripple voltage is measured simultaneously with the load-current measurements, as the ripple voltage will vary somewhat with the load current. Measurements are generally made at 100, 50, and 10 percent of the load current ratings and for various values of line voltage. Care must be taken when one is measuring regulated power supplies employing automatic crossover or current-limiting circuits, that they remain in the intended mode of regulation under all conditions of loading; otherwise, they appear to have poor regulation.

Ripple voltage is measured in millivolts, and for a well-designed power supply (employing feedback), it is less than 1 millivolt, the ripple frequency appearing at 120 Hz, 240 Hz, and higher. If the frequency of the ripple is unrelated to the line-voltage frequency, the ripple is, as a rule, caused by interference from nearby equipment.

The circuit in Fig. 23-163 is a typical unregulated power supply and has been selected to illustrate the various points where ripple voltage can be measured, and the equipment required to measure the ripple.

Connected across the output of the power supply is a 600-volt, 0.1-μF paper or oil capacitor, in series with a 100,000-ohm resistor. The purpose of the capacitor is to give increased protection to the voltmeter input. The resistor ties the capacitor to ground. Since the ripple voltage flows through the resistor to ground, the meter is connected across this resistor. This stabilizes the reading and is also of considerable help when ripple waveforms are photographed. Also connected across the output is a bleeder resistor, R_{BL} (normally a part of the power supply), variable load resistor R_L, milliammeter I_L, and voltmeter V_{dc} for measuring the output voltage under various load conditions.

Slow drifts and transients may be expected in this type of power supply and are the result of line-voltage fluctuations, which is perfectly normal for an unregulated power supply. For a regulated power supply the waveform images are quite steady.

Ripple voltage may be calculated:

$$\text{Percent ripple voltage} = \frac{\text{ac volts}}{\text{dc volts}} \times 100$$

where,

ac volts is the measured ripple voltage,
dc volts is the measured output voltage.

The amount of ripple voltage in decibels below the maximum dc voltage is:

$$20 \operatorname{Log}_{10} \frac{\text{dc voltage}}{\text{ac voltage}}$$

where,

dc voltage is the maximum voltage under normal load conditions,
ac voltage is the ripple voltage under the same conditions.

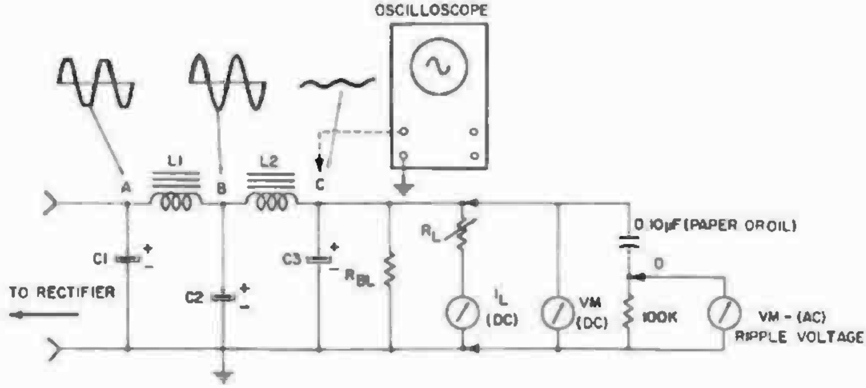


Fig. 23-163. Circuit for measuring the load characteristics and ripple voltage of a power supply. The filtering efficiency may be observed or measured at points A to C. Ripple voltage is measured at D.

Ripple voltage for a well-filtered supply is usually on the order of 80 to 120 dB below the maximum dc voltage at the normal operating current. The ripple voltage is measured at points A, B and C. The waveform shape at these various points is indicated above the filter sections. It will be observed the waveform at the output of the rectifier (full wave) has sharp peaks, with a slight flattening off at the peak. At the output of the first choke (point B), the waveform approaches a sine wave, but is rounded at the peaks. After being filtered by the second choke (point C), the ripple voltage is reduced to almost pure dc, or may have a very small pulsation. If the filtering efficiency is suspected and electrolytic capacitors are used in the filter system, they should be disconnected and others substituted, and the ripple voltage measured again.

Filtering efficiency varies with the type and number of filter sections. Different values of ripple voltage and waveform shape may differ from those indicated for this example. For electronically regulated supplies the waveform may contain spikes, which, when measured with an ordinary voltmeter, will not be indicated. Therefore, an oscilloscope must be used. Ripple voltage of this character is caused by the semiconductor rectifiers as they switch from nonconduction to conduction. For current-regulating power supplies, the current ripple rather than the voltage ripple is of primary importance.

Ripple factor is a term used to indicate the amount of ripple voltage suppression contributed by the filtering system or regulator, and it is a ratio expressed as a percent of the output ripple voltage to the input ripple voltage:

$$\text{Ripple factor} = \frac{E_{out}}{E_{in}} \times 100$$

where,

E_{out} is the peak-to-peak ripple voltage at the output of the supply,
E_{in} is the peak-to-peak ripple voltage at the output of the rectifier.

23.164 Describe a method for measuring power-supply load characteristics. —In Fig. 23-164A is shown a method for measuring the load and line regulation of a constant-voltage power supply. In setting up for measurement, several precautions must be observed. The accuracy of the meters should be 1 percent or better, and they must be connected directly to the terminals of the power supply to eliminate any error caused by voltage drop in the connecting leads. If clip leads are used, they must make good contact, as the internal impedance in a well-designed power supply may be on the order of 1 milliohm (0.001 ohm). Soldering the leads, where permissible, will eliminate high-resistance connections.

Load resistor R_L must be wirewound and operated at about one-tenth its rated power dissipation. This is necessary to prevent the surface temperature from rising to only a small amount above the ambient temperature. Failure to observe this precaution will result in a surface bubble on the resistor, with a corresponding change in the value of the resistance. This can cause short-term variations in the measured voltage drop that are not the result of output current changes.

The voltmeter and ammeter (if ac operated) must be fed from a separate circuit or from one isolated from the line feeding the power supply, to eliminate line voltage variations to the instrument. The line should be nonregu-

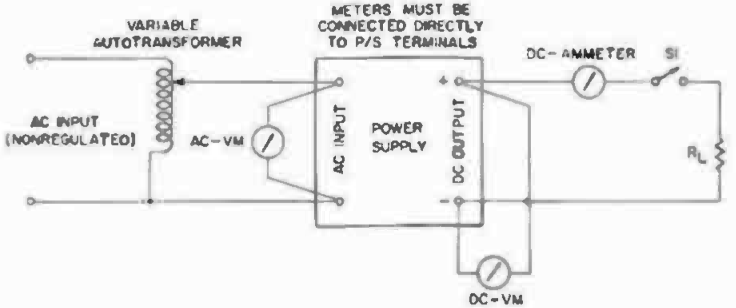


Fig. 23-164A. Circuit for measuring load and line regulation of a constant-voltage power supply.

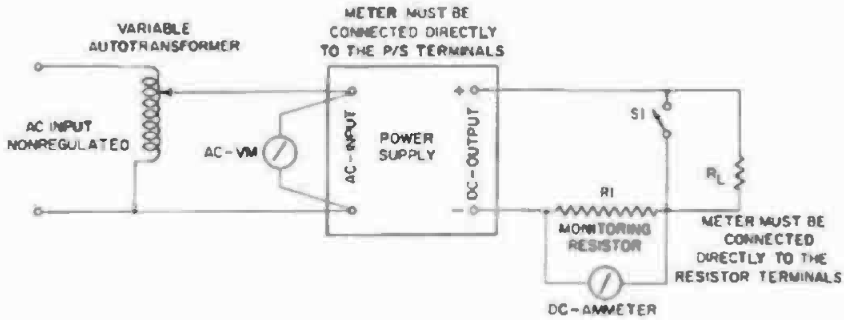


Fig. 23-164B. Circuit for measuring load and line regulation of a constant-current power supply.

lated and of good waveform. Failure to observe the latter precaution may result in distortion of the line voltage waveform fed to the power supply and thus affect the power-supply characteristics. In the instance of a power supply employing a SCR regulator, serious malfunctioning of the firing circuit is possible.

The line voltage is set to 117 volts and the dc load is adjusted for a reference current and voltage. The autotransformer is then varied from 105 Vac to 125 Vac, then back to 105 Vac again, and the variation is noted for different line voltage settings. For a well-regulated supply, the output voltage will remain almost constant.

Load regulation is measured by opening and closing switch S1 and noting the change in output voltage while the line voltage is varied over the range of interest. Notice should be taken in which direction the output voltage changes. If the output voltage drops with an increase in load current, this is indicative of a *positive* resistance at direct current, whereas an increase of output voltage with an increase of load current indicates a *negative* output resistance.

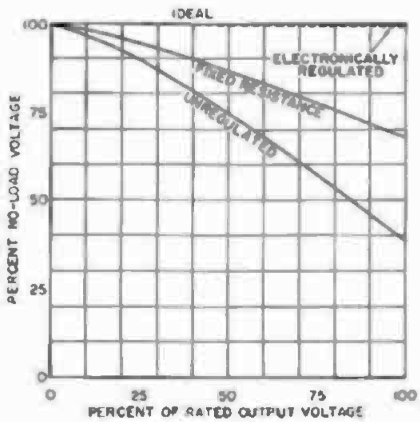


Fig. 23-164C. Comparison of regulation characteristics for power supplies.

The measurement of constant-current power supplies (Fig. 23-164B) is somewhat similar. However, series-current monitoring resistor R1 must be small in value (about 1 ohm or less) in order that the voltage drop across it is only a fraction of the total output voltage. It must be wirewound and have an accuracy of 1 percent or better.

The value of load resistance R_L is chosen so that a voltage drop across it, at the current level of the measurement, is equal to the voltage rating of the power supply minus the voltage drop across resistor R_L. S1 is connected in parallel with load resistor R_L to measure the current regulation between a low voltage (equal to that across R_L) and the maximum rated voltage of the supply.

The dc meter may be a differential or digital voltmeter, a dc oscilloscope, or any other type of instrument capable of accurate measurement of the voltage drop across R1.

Load variations are measured by closing switch S1 and noting the change in output voltage as measured across R1 while the supply is operated at various values of line voltage. Load regulation is then:

$$L_{reg} = \frac{\text{Volts across R1}}{\text{R1}}$$

Figure 23-164C compares typical regulation curves for different types of power supplies, both regulated and nonregulated, plotted percent no-load versus percent output current. It is apparent that the regulated type of power

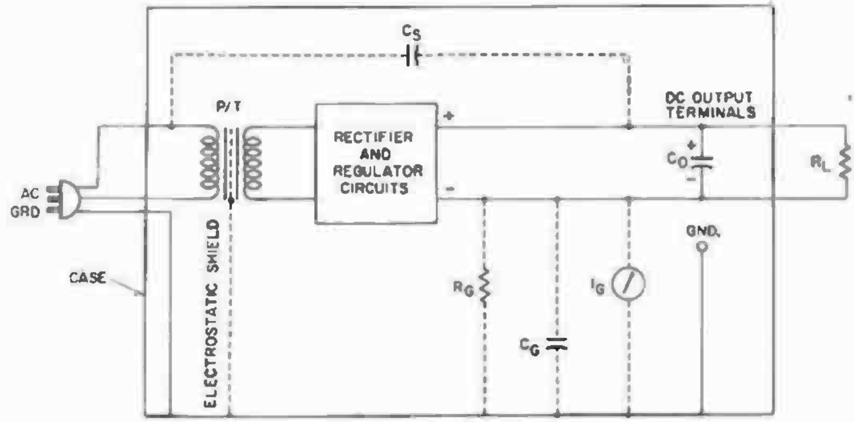


Fig. 23-165A. Leakage paths in a power supply.

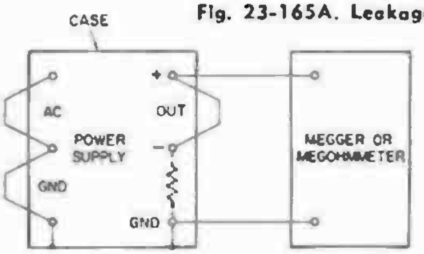


Fig. 23-165B. Measurement of leakage resistance between output and ground.

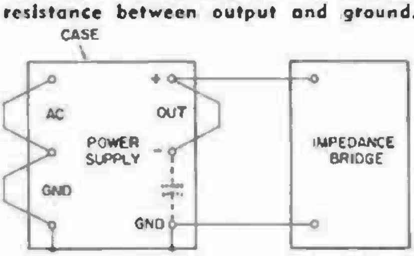


Fig. 23-165C. Measurement of shunt capacitance between output and ground. Impedance bridge must be isolated from power supply when it is activated.

supply characteristic is quite superior to that of the unregulated type.

23.165 ***Describe the procedure for measuring the isolation of a power supply.***—To properly evaluate the isolation characteristics of a power supply, five different measurements are necessary. They are: leakage resistance between output and ground, shunt capacitance between output and ground, noise between output and ground, capacitance between ac input and dc output, and the insulation breakdown voltage to ground. Isolation measurements are necessary if it is desired to evaluate the power supply output potential above ground, reduce the effects of ground currents, or when the load requires a floating power supply.

An elementary diagram of a typical power supply with its various leakage paths is given in Fig. 23-165A. Resistor R_G represents the leakage from both the positive and negative terminals to ground (case) and to other points in the circuitry. Capacitance C_s is the stray capacitance from the primary side of the power transformer to all points associated with the dc output.

Measurements of the leakage resistance between the output and ground are made with the power supply dc activated. The ac input and ground terminals are shorted, and the resistance is measured between the output terminals and ground, using a megger or megohmmeter (Fig. 23-165B). If a group of power supplies are to be compared, these measurements must be made at the same relative humidity.

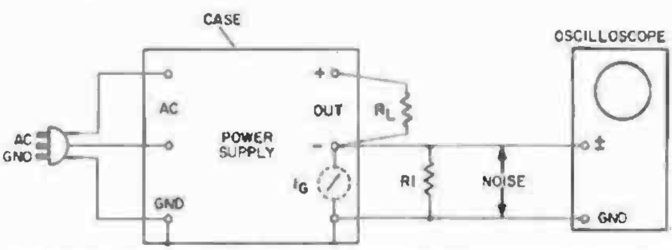


Fig. 23-165D. Measurement of noise current between output and ground.

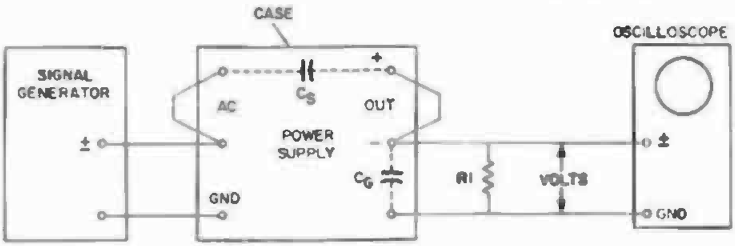


Fig. 23-165E. Measurement of capacitance between ac input and dc output.

Measurements of shunt capacitance between output and ground are made in the same manner, except that an impedance bridge is used. It is possible for the shunt capacitance to be of a different value when the power supply is activated. This may be measured also by the use of an impedance bridge; however, the bridge must be isolated in such a manner the dc output from the power supply does not damage the bridge. At this time an oscilloscope should be employed for checking the existence of grouud loop currents, which can affect the impedance bridge readings. With the supply unit activated, the bridge null may not be very sharp because of noise injected between the output terminals of the supply and ground. Connections for this test of shunt capacitance are given in Fig. 23-165C.

Noise measurements are made as shown in Fig. 23-165D. Noise is generated mainly from capacitance coupling between the chassis and points in the circuitry having a large or high-frequency potential. Because of the small equivalent capacitances and high voltages associated with them, the noise source is generally represented as a high impedance or current source which injects noise from the chassis into the output terminals. Noise may be defined as the voltage appearing across a 1000-ohm resistor R1 connected from either output terminal to ground while the power supply is activated. An ac meter or oscilloscope is connected across resistor R1 for monitoring purposes. Similar measurements are also made using a 100-ohm and 10,000-ohm resistor. If I_g is a constant-current source, the voltage measured across 10,000 ohms will be ten times greater than that across the 1000-ohm resistor. If the supply or load circuit employs capacitors between the output terminals and ground, it is likely that a linear relationship will not exist across resistor R1.

Transformer leakage is measured as shown in Fig. 23-165E. It represents the mechanism by which common-mode signals on the ac line are coupled longitudiually to the dc outpnt. Capacitance C_s cannot be measured by an impedance bridge, since it is not separable from the other existing capacitances. Capacitance C_s is measured by injecting a signal-generator voltage between the chassis ground and the shorted output terminals. Resistor R1 (1000 ohms) is connected between the output terminals and ground. The oscilloscope is used for monitoring the signal and to detect any form of signal not connected with the measurement. A tuned voltmeter is useful in making this measurement.

The final measurement consists of measuring the voltage breakdown to ground (Fig. 23-165F). The ac input terminals are shorted to ground, and a high-voltage test set is connected between the ac and the dc output terminals. The high-voltage test unit must have a high internal resistance so that when a voltage breakdown occurs in the power supply the insulation is not damaged. In most instances the mica washers separating the transistors from ground in a heat sink will withstand about 1000 volts per mil of thickness.

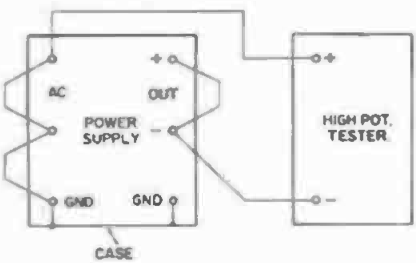


Fig. 23-165F. Measurement of breakdown voltage between ac input and dc output to ground.

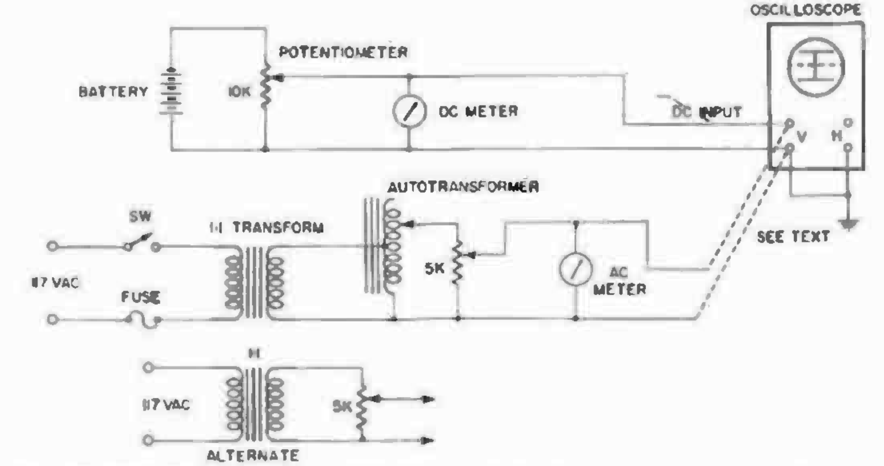


Fig. 23-166A. Test circuit for calibrating an ac meter using a dc meter and battery. All grounds must be removed from the circuit or an isolating transformer used in the ac line.

As a rule, a test voltage of 500 to 600 volts is sufficient.

23.166 ***How can an ac voltmeter be calibrated, using a dc source?***—If a dc meter of sufficient accuracy and a cathode-ray oscilloscope are available, ac voltmeters may be calibrated or rechecked without the benefit of an ac standard meter. The oscilloscope is first calibrated, using a source of battery voltage as shown in Fig. 23-166A with the ac portion of the circuit disconnected. The input control of the oscilloscope is set for dc input, and the sweep speed is adjusted for about 10 microseconds, or a sweep without a flicker. The trace is centered on the graticule-center reference line. If the oscilloscope has a dc balance control (usually a screwdriver control), it must be balanced to remove any deflection of the beam when the input attenuator is adjusted. The attenuator is now adjusted for a convenient deflection of about 2 centimeters above and below the reference line. Reversing the polarity of the battery, the deflection should be exactly the same amount in the other direction. The graticule is now ready for calibration.

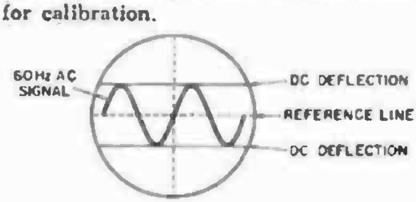


Fig. 23-166B. Oscilloscope graticule with dc calibration lines and ac signal.

Suppose that an ac voltmeter having a 10-volt scale is to be checked. Apply a *dc voltage* equal to the peak value of 10 volts ac to the oscilloscope, or 14.141 volts, using batteries or a regulated power supply. Mark both the upper and lower deflection points. Leaving all controls set, disconnect the dc circuit, and substitute the ac circuit. Adjust the autotransformer for a deflection on the oscilloscope exactly equal to that obtained with the dc input signal. The ac meter should now read 10 volts. Other scales may be checked using the same procedure, always remembering that *the dc voltage must equal the peak voltage of the desired ac deflection.* (See Question 25.149.)

Although many vacuum-tube voltmeters actually read peak-to-peak voltage, they are calibrated to read the rms value of a sine wave. (See Questions 22.98 to 22.102.) As a matter of safety, if an autotransformer is used in the ac calibrating side of the circuit, a 1:1 isolation transformer should be used to isolate the calibrating circuits. Also, the ground (if any) must be disconnected from the oscilloscope ac power cable by the use of an ac isolating plug. (The alternate method to this is to phase out the ground side to all equipment, as shown.) If an isolating transformer is used, the ground may be left on the oscilloscope.

For the calibration of ac scales of 5 volts and below, four mercury cells connected in series may be used; however, there are some disadvantages when more than three cells in series are used. (See Question 25.8.)

23.167 ***How is the voltage calculated for calibrating a VU or VI meter for a given reference power?***—The meter is connected in parallel with the source of calibrating voltage as shown in Fig. 23-167. The voltage for any reference power may be calculated by:

$$\text{Volts} = \sqrt{P \times R}$$

where,

P is the reference power,
R the impedance of the circuit for which the meter is to be calibrated.

Thus, if a meter is to be calibrated for a reference power of 6 milliwatts in a 600-ohm circuit:

$$\begin{aligned}\text{Volts} &= \sqrt{0.006 \times 600}\\ &= \sqrt{3.6}\\ &= 1.89\end{aligned}$$

The calibrating source does not have to be equal to the impedance for which the meter is to be calibrated. Any impedance will suffice if the voltage is equal to 1.89 volts. (See Question 10.42.)

23.168 ***Can a 1000-ohms-per-volt meter be used to measure the operating voltages of transistor or resistance-coupled amplifiers?***—No, because of the current drawn by the meter, which in the instance of a 1000-ohms-per-volt meter, is 1 milliampere. In measuring the voltage of a transistor or resistance-coupled amplifier stage, the increase in current due to the meter will increase the voltage drop across the circuit being measured, resulting in an erroneous reading, and in the case of transistors, possible damage.

Consider a circuit to be measured containing a 47,000-ohm resistor with a true voltage drop across it of 155 volts, using a meter with a 10-megohm input resistance. If the same voltage is measured again, using a meter of 1000 ohms per volt (300-volt scale) the indicated voltage drop will be on the order of 146 volts; and if read on a 150-volt scale, 142 volts. Operating voltages in transistor and resistance-coupled devices must be measured by a meter of at least 10 megohms of input resistance if a true reading of voltage is required. Such meters are discussed in Section 22.

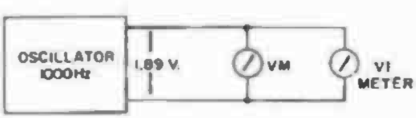


Fig. 23-167. Circuit for calibrating a VU meter for a given reference level.

23.169 ***What is the proper way to connect a voltmeter and an ammeter in the same circuit?***—Two methods may be used, as shown in Figs. 23-169A and B. The first method is used when the load current is large. The second method is used when the current load is small.

In the first instance with the load current being large, the additional current drawn by the voltmeter is of no consequence. However, if the load current is small (on the order of a few milliamperes), the current drawn by the voltmeter might result in an erroneous indication.

23.170 ***How can a simple capacitor tester be constructed using an ac voltmeter or milliammeter?***—As shown in Fig. 23-170. An ac voltmeter with either a copper-oxide or selenium rectifier is connected in series with a source of 115 volts ac and the capacitor. Capacitors of known value are connected as shown, and the meter scale is calibrated to read directly in capacitance.

A second method is to connect an ac milliammeter of 1-milliampere range in place of the voltmeter. However, this will require a series resistor similar to a voltmeter multiplier. If the meter is calibrated for 60 Hz, it can only be used on that frequency; otherwise the calibration will be in error. The value of an unknown capacitor may be calculated by reading the current through

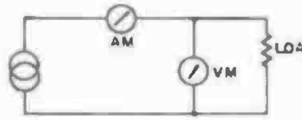


Fig. 23-169A. Method of connecting an ammeter and voltmeter when the load current is large.

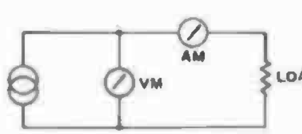
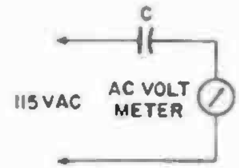
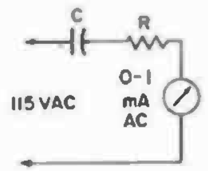


Fig. 23-169B. Method of connecting an ammeter and voltmeter when the load current is small.



(a) *Using an ac voltmeter.*



(b) *Using an ac milliammeter.*

Fig. 23-170. Circuits for simple capacitor meter.

the milliammeter. The capacitance is then equal to:

$$C = \frac{10^6 I}{2\pi f E}$$

where,

E is the line voltage,
f is the frequency,
I is the current as read on the milliammeter.

The foregoing test circuits are for oil- or paper-type capacitors only. Electrolytic capacitors have high leakage and cannot be read by this method. Capacitor testers suitable for use with electrolytic capacitors are discussed in Question 22.34.

23.171 What is the substitution method of measuring capacitance?—Because capacitance bridges have internal stray capacitance due to their design and circuitry, capacitors of 10 picofarads (0.00001 μF) or smaller are sometimes difficult to measure due to the residual capacitance of the bridge being larger than the capacitance to be measured. When this situation is encountered, the substitution method of measurement is used. As an example, if a capacitor of 5 picofarads is to be measured, connect a 10-picofarad capacitor in parallel with it. Measure the combined capacitance, remove the 5-picofarad capacitor, and measure the 10-picofarad capacitor alone. The difference between the two measurements is the value of the smaller capacitor. When these measurements are made, the capacitors must be directly connected to the terminals of the bridge to reduce the stray capacitance to an absolute minimum.

The residual or stray capacitance of a bridge may be measured by balancing the bridge with the capacitance terminals open. The value of capacitance read is the internal stray capacitance and is the factor that limits the lowest value of capacitance that can be measured with that particular bridge. In some types of measurements, the residual or stray capacitance is subtracted from the measurement; however, this would only be done in the case of an extremely small value of capacitance. Bridges such as those described in Question 22.33 as a rule have from 2 to 5 picofarads of stray capacitance.

When one is using a capacitance bridge for the first time, difficulty may be experienced when attempting to measure small values of capacitance. This difficulty can be overcome by grounding the shield of the bridge or by the use of a guard circuit which offsets the capacitance between the bridge and ground.

If there is any doubt as to the accuracy of a bridge, standard capacitors may be connected and measured using different grounding methods to eliminate the effect of stray capacitance. As a rule, the manufacturer of a particular bridge supplies information relative to grounding methods.

23.172 How may complex waveforms be generated?—A simple method of generating complex waveforms is shown in Fig. 23-172A. For the precise

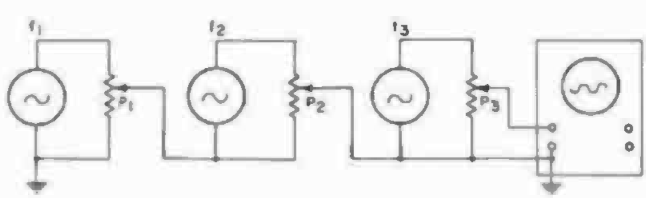


Fig. 23-172A. Three oscillators connected in series for the purpose of generating complex waveforms.

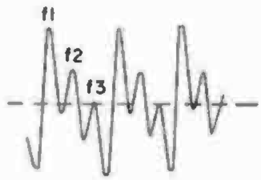


Fig. 23-172B. Resulting complex waveform when three oscillators are connected in series. Frequency f1 is the fundamental; f2 and f3 are the second and third harmonics.

production of such waveforms a rather complicated setup is required. However, for simple displays the circuit shown will suffice. The three oscillators are connected in series, with a gain control at each output. The final waveform is observed by means of an oscilloscope connected across the output of the last generator.

By setting the oscillators to different frequencies and amplitudes, complex waveforms are generated. A typical waveform appears in Fig. 23-172B.

23.173 Do oscillators generate subharmonics?—No, not unless they are used in conjunction with a frequency divider which will generate subharmonics of the oscillator fundamental frequency. Frequency dividers are often used with frequency standards as described in Question 22.40.

23.174 How can the internal distortion in an oscillator be reduced?—By the use of an external bandpass or low-pass filter. However, as this method requires a separate filter for each frequency, it becomes rather impractical.

If the oscillator is of the Wien-bridge type, the distortion can generally be reduced by adjusting the feedback loop to the bridge circuit and by the selection of tubes. If the instrument uses a lamp in the cathode circuit, a selected lamp may help to reduce the distortion. Also, the more accurate the balance between the two sides of the bridge, the lower will be the harmonic distortion. (See Questions 22.47 and 22.48.)

23.175 How may two audio oscillators be connected in parallel for calibration without pull-in? — By connecting them in parallel using a parallel-T network as shown in Fig. 23-175. The outputs of the standard oscillator and the uncalibrated oscillator are set to the same output level by observation of the level on the null indicator. When the uncalibrated oscillator is tuned through the same frequency as the standard, the pointer of the meter will swing back and forth over a wide range. As the frequency of the uncalibrated oscillator is brought nearer to the frequency of the calibrated oscillator, the swing is reduced. When they are both at the same frequency, the pointer will be at a minimum deflection or at zero beat. (See Question 23.178.)

Beat indications will be obtained at frequencies one-half and twice the frequency of the standard. However, frequencies beyond these will not be indicated. Submultiple or harmonic frequencies will be of considerably less amplitude than the fundamental frequencies.

If an oscilloscope is connected as shown, the beat will be indicated by a winking or a change in amplitude of the observed signal. Zero beat is the point where the signal on the oscilloscope screen is at a standstill and of minimum amplitude.

A much more satisfactory method of calibrating oscillators is explained in Question 23.176.

23.176 How are Lissajous figures used for frequency calibration?—By con-

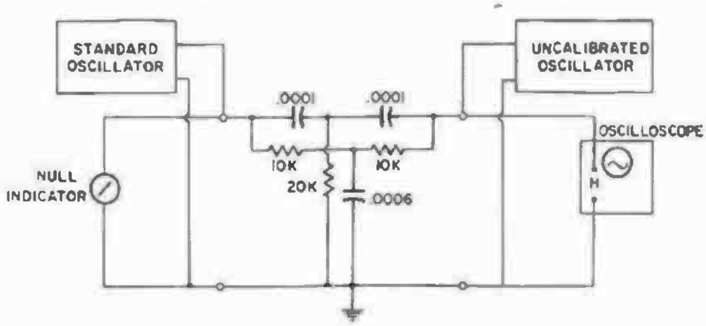


Fig. 23-175. Parallel-T network connected in the outputs of two oscillators, to prevent pull-in during calibration.

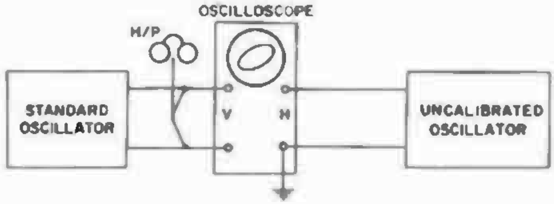


Fig. 23-176A. Method of calibrating an oscillator against a frequency standard.

necting a frequency standard or a calibrated oscillator to the vertical input of the oscilloscope and the oscillator to be calibrated to the horizontal input as shown in Fig. 23-176A.

Suppose an oscillator is to be calibrated against a frequency standard similar to that described in Question 22.40. The frequency standard is set to 1000 Hz and the dial of the uncalibrated oscillator is rotated until a circular pattern as shown at (a) of Fig. 23-176B is obtained. This pattern indicates a ratio of 1:1 and is obtained when both the standard and the uncalibrated oscillator are set to exactly the same frequency. It is the starting point for future calibrations.

If the circular pattern slowly drifts toward an oval, then a diagonal line, and back again to an oval and then a circle, it has passed through one complete cycle. The number of cycles the pattern passes through in 1 second is the number of cycles per second the uncalibrated oscillator is off calibration from the standard. If calibration is held to within 2 percent of standard frequency, this is generally satisfactory. However, if oscillator is stable, it can be calibrated to within less than 1 percent without too much difficulty.

Assume the next frequency to be calibrated is 2000 Hz. If the frequency standard does not supply this frequency, a Lissajous figure indicating a frequency ratio of 2:1 will be used, as in (b) of Fig. 23-176B. The pattern now has one loop in the vertical plane and two in the horizontal plane. For 3000 Hz, a 3:1 configuration is used as shown in (c). For even multiples of the 1000-Hz standard freqnency, the ratios will be 4:1, 5:1, and so on. For frequencies that are not multiples of the standard ratios, patterns such as 3:2 and 4:3 are employed. The frequency of the uncalibrated oscillator is determined from the

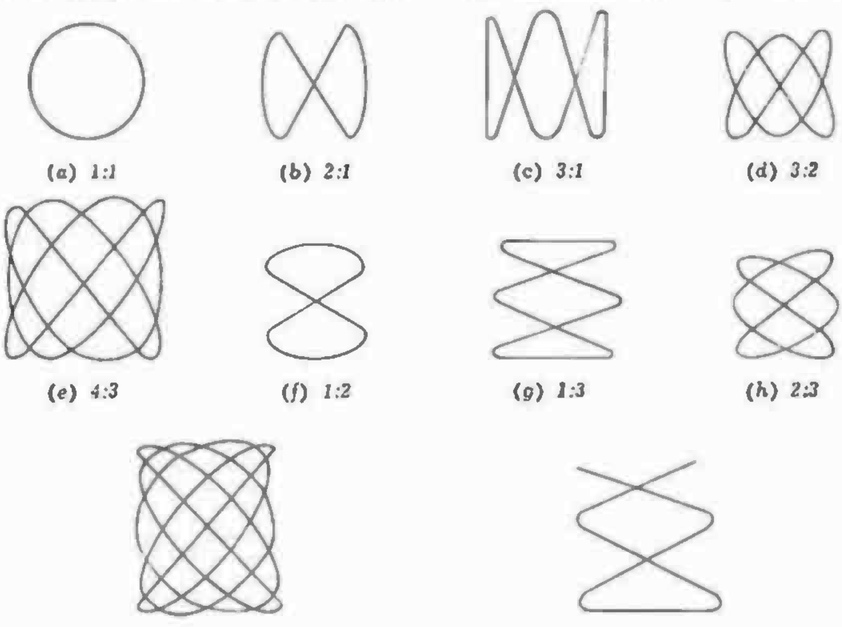


Fig. 23-176B. Lissajous figures.

equations in Question 23.177. As the higher frequencies are approached, a higher frequency is employed for the standard to reduce the difficulty of counting the loops.

For frequencies below the lowest standard frequency, say 500 Hz, a configuration with two loops will be observed, except the loops are now in the vertical plane rather than the horizontal, as in (f) of Fig. 23-176B.

Several patterns that are used for frequencies below the standard frequency are shown in (g) to (i) of Fig. 23-176B. If available, a low-frequency standard should be used to aid in counting the loops.

An incomplete loop is shown in (j) of Fig. 23-176B and may be used when calibrating, if the ratio is definitely known. However, because incomplete loops are rather difficult to identify, the complete loop is generally used.

23.177 What is the equation for calculating the frequency ratio from a Lissajous figure?—For frequencies greater than that of the standard frequency:

$$F_x = \frac{L_h}{L_v} \times F_s$$

where,

- L_h is the number of loops in the horizontal plane,
- L_v is the number of loops in the vertical plane,
- F_s is the frequency of the standard,
- F_x is the unknown.

For frequencies below the standard:

$$F_x = \frac{F_s}{L_h/L_v}$$

A typical example is the 4:5 ratio at (i) of Fig. 23-176B. If the standard is 1000 Hz, the unknown frequency is 1000/1.25, or 800 Hz.

If the standard oscillator is replaced by a secondary frequency standard similar to those described in Question 22.40, a different method is used for frequencies other than those normally supplied by the standard. As a rule, a secondary frequency standard supplies output frequencies which are multiples of a master crystal-controlled oscillator contained within the standard. For calibration of frequencies not supplied by the standard, Lissajous figures are employed in conjunction with an oscilloscope.

Lissajous figures are not dependent on the applied frequency or frequencies. The pattern indicates only the ratio existing between the applied frequency and that of the standard. By proper interpretation of the patterns, various frequencies may be obtained using a single standard frequency. Headphones may be connected across the output of the uncalibrated oscillator to determine whether the output frequency is above or below that of the standard. This will be of help in the interpretation of the patterns.

23.178 What does the term "zero beat" mean?—It is a well-known fact that when two frequencies are combined or beat against each other, a third frequency called the difference or beat frequency is produced. The third frequency will vary if one oscillator is held constant and the frequency of the other is varied.

Fig. 23-178A is a plot of the beat frequency against the audible range of frequencies. When the difference frequency is above audibility, no sound is heard, as indicated by the shaded area of the graph.

If the variable oscillator is varied from a low frequency, starting at the left and indicated by the letter (a), no sound is heard because the beat frequency is above the range of audibility. Increasing the variable oscillator frequency causes the difference or beat frequency to decrease until it reaches point (b). Here it is quite low in frequency and sounds more like a series of clicks than a smooth tone. As the variable oscillator is continued toward the higher frequencies, the beat frequency becomes lower and lower until it becomes inaudible. This is the point, called *zero beat*, at which the frequen-

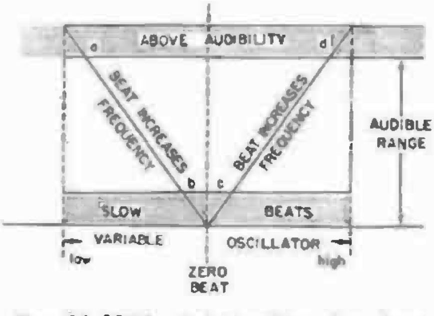


Fig. 23-178A. Relationship of a beat frequency to one fixed and one variable oscillator.

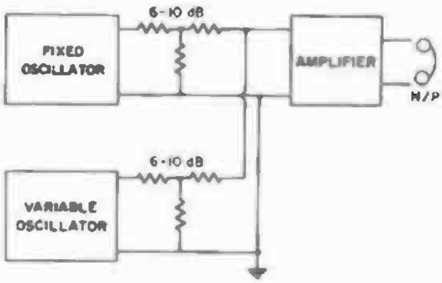


Fig. 23-178B. Two audio oscillators connected in parallel to produce a difference frequency. The pads prevent the oscillators from pulling into the same frequency.

cies of the two oscillators are exactly the same.

If the variable oscillator is passed through the zero point and its frequency continues to be increased, the beat will be heard again and will increase in frequency until it reaches point (d) and passes into the inaudible-frequency area again.

Oscillators may be calibrated using the beat method described by connecting them in parallel and listening to their outputs with headphones. To prevent the oscillators from being pulled in by each other, it is desirable to connect a pad of 6- to 10-dB loss in the output of each oscillator, as shown in Fig. 23-178B.

Beat-frequency oscillators are constructed using the previously described method of producing a difference frequency, using radio-frequency oscillators. Such oscillators are described in Question 22.51.

23.179 What is an A/B test?—A comparison of two similar devices under the same electrical, acoustical, or optical conditions, When such tests are made, as in the instance of speaker systems, the two speaker systems are connected to a switching arrangement so that they may be switched while listening to program material using the same electrical or acoustical output levels.

Generally, such tests are made under closely controlled conditions so that a direct comparison can be made and conclusions drawn.

23.180 How can the null indication of an ac bridge be sharpened?—A poorly defined null indication is caused by the presence of a considerable amount of harmonics in the signal source being used to balance the bridge. The null indication can be sharpened by the use of a bandpass or low-pass filter in the signal source.

23.181 How may an oscilloscope be used for measuring current?—By measuring the voltage drop across a shunt resistor connected in series with the circuit carrying the current to be measured.

The screen of the scope is calibrated in voltage and the current calculated using the equation:

$$I = \frac{E}{R}$$

where,

E is the voltage drop across the shunt,
R is the resistance of the shunt in ohms,
I is the current through the shunt.

23.182 How may the sensitivity of a dc meter movement be measured?—By connecting the meter movement in series with a source of current, a variable resistance and a milliammeter as shown in Fig. 23-182. The current is measured for a full-scale deflection of the unknown meter.

23.183 Show a simplified method for measuring the impedance of a capacitor.—The capacitor to be measured is connected in parallel with a decade resistance box, a vacuum-tube voltmeter, and an oscillator as shown in Fig. 23-183. Resistor R_s of approxi-

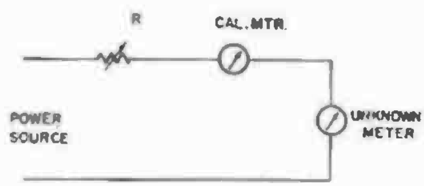


Fig. 23-182. Circuit for measuring the sensitivity of a dc meter movement. A milliammeter is connected in series with the unknown meter. The resistor R is adjusted for a full-scale deflection of the unknown movement.

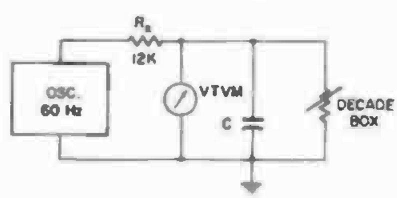


Fig. 23-183. Simplified circuit for measuring the impedance of a capacitor.

mately 10,000 ohms is connected in series with the output of the oscillator to raise its internal impedance.

The measurement is made by first disconnecting the decade resistance box and setting the oscillator to 60 Hz for a convenient reading on the meter. The decade box is then connected and adjusted for a shunt value that will reduce the reading on the meter 3 dB below the reference point. The impedance of the capacitor is then equal to that of the decade box.

The impedance measured in this manuer is only at a frequency of 60 Hz. The impedance of the capacitor increases with a decrease in frequency. If a number of capacitors are to be measured, the decade box must be disconnected, and the reference voltage reestablished for each capacitor. It will be noted that capacitors labeled to have the same capacitance will have slightly different impedances at the same frequency. This is caused by variations in manufacturing tolerances.

Electrolytic, paper, oil, and other types of capacitors also can be measured with this method. Approximate values of impedance at 60 Hz for capacitors of different sizes are given below.

Value (μF)	*Impedance* (*ohms*)	*Value* (μF)	*Impedance* (*ohms*)
0.10	26,500	40	66.5
1.0	2655	100	26.5
5.0	532	500	5.32
10.0	265	2,000	1.33
20.0	133		

23.184 How is an electronic switch connected to an oscilloscope for displaying two signals simultaneously?—In the older-type oscilloscopes, dual-trace display was the exception rather than the rule. However, older instruments may be made to display multiple signals by the use of an electronic switch connected to the vertical amplifier.

If the input and output signals of an amplifier (or other device) are to be observed simultaneously, an electronic switch is connected to the input of the oscilloscope vertical amplifier, with a lead taken from the output of the signal uuder observation to the external trigger circuit of the oscilloscope, to synchronize and stabilize the display (Fig. 23-184A).

By proper adjustment of the electronic switch controls, the two waveforms can be separated for comparison and individual study. Three waveforms may be displayed by the use of two electronic switches, as shown in Fig. 23-184B.

Fig. 23-184C shows the appearance of a single waveform with a base line. This display is obtained by using one electronic switch. The two input gain controls are set to zero, and the two horizontal traces are superimposed by means of the position control. The gain control of the input signal to be analyzed is advanced for desired image size. The second gain control is left closed, providing the base line or trace.

Fig. 23-184D shows two waveforms separated for comparisou by adjusting the position control for the desired separation.

Fig. 23-184E shows the appearance of two waveforms superimposed for direct comparison, while Fig. 23-184F shows two traces adjusted for phase difference measurement. The two traces are superimposed by means of the position control, and then both gain controls are advanced until both of the wave-

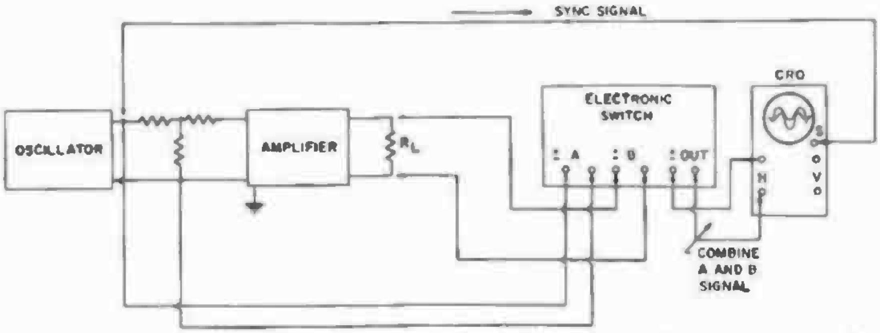


Fig. 23-184A. Method of connecting an electronic switch for the simultaneous observation of two waveforms.

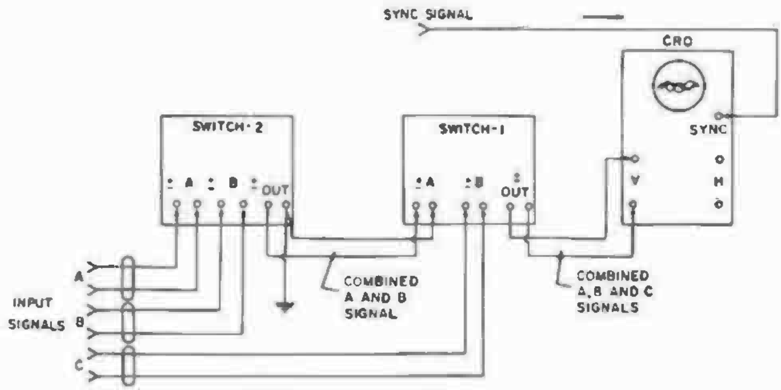


Fig. 23-184B. Method of connecting two electronic switches for the simultaneous observation of three waveforms.

forms are clearly visible. The phase angle may be computed:

$$\phi = 180 \frac{X_1}{X_2} = \text{degrees}$$

where,

X_1 is the distance between the leading edges of the two waveforms,
X_2 is the distance between the two waveforms at their base line.

Fig. 23-184G shows a pattern obtained by the use of two electronic switches connected in tandem for displaying three waveforms simultaneously. Two of the waveforms are superimposed by means of the position control of one instrument, while the other waveform is displayed by means of the position control of the second instrument.

23.185 How may a cross-modulation oscillator used for photographic-film recording be tested for proper functioning?—The oscillator panel is adjusted for various output signal levels in a manner normally employed for recording cross-modulation tests on film. The cross-modulation test panel containing a 400-Hz bandpass filter is patched to the output of the oscillator

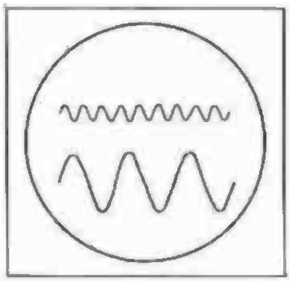


Fig. 23-184D. Separation of two traces when using an electronic switch.

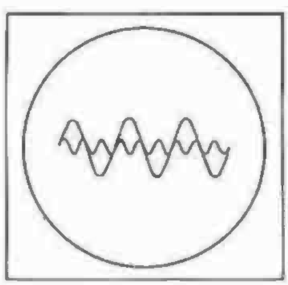


Fig. 23-184E. Superimposition of two waveforms for direct comparison.

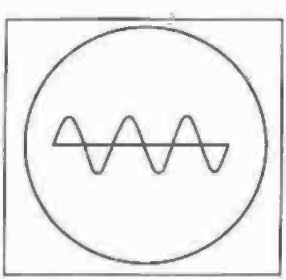


Fig. 23-184C. Waveform with base line when using an electronic switch.

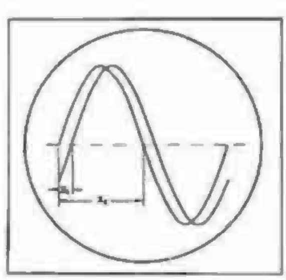


Fig. 23-184F. Measuring the phase angle between two waveforms.

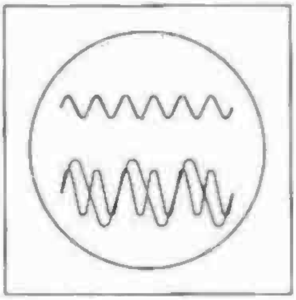


Fig. 23-184G. Appearance of images when two electronic switches are connected in tandem for displaying three waveforms simultaneously.

with a repeat coil between as shown in Fig. 23-185A. The repeat coil is a necessity to prevent leakage and the formation of ground loops.

The cross-modulation products of the oscillator section are measured in the same manner prescribed for the measurement of sound tracks in Question 18.233. A cross-modulation oscillator measured in this manner should show a cancellation of at least 60 dB or better. The harmonic distortion of the individual oscillators should not exceed 2 percent. If the oscillator panel employs a fixed ratio of modulation between the oscillators, the percent of modulation may be measured by the use of an oscilloscope. The image on the oscilloscope screen, as in Fig. 23-185B, is adjusted for a ten-line deflection above and below the center line of

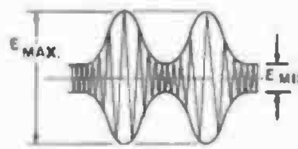


Fig. 23-185B. Method used for the calculation of percentage modulation of cross-modulation oscillator. This method of measuring percentage modulation may be applied to any amplitude-modulated system.

the graticule, or a total spread of 20 lines. The percent of modulation may be computed:

$$\%\ \text{modulation} = \frac{E_{max} + E_{min}}{E_{max} - E_{min}} \times 100$$

where,

E_{max} is the maximum swing of the carrier voltage,

E_{min} is the minimum swing of the carrier voltage.

If the percent of modulation is 75, the minimum swing will cover 2.86 lines, or a swing of 1.43 lines on each side of the center line.

This method of measuring the percent of modulation may be used for computing the percentage modulation for any amplitude-modulated system.

23.186 How is the percent of modulation of a radio transmitter measured? —The percent of modulation of a radio transmitter may be measured in two different ways. The first is to connect a coupling coil to the vertical amplifier of an oscilloscope and place it near the tank coil of the radio transmitter, as in Fig. 23-186A. A pattern of the modu-

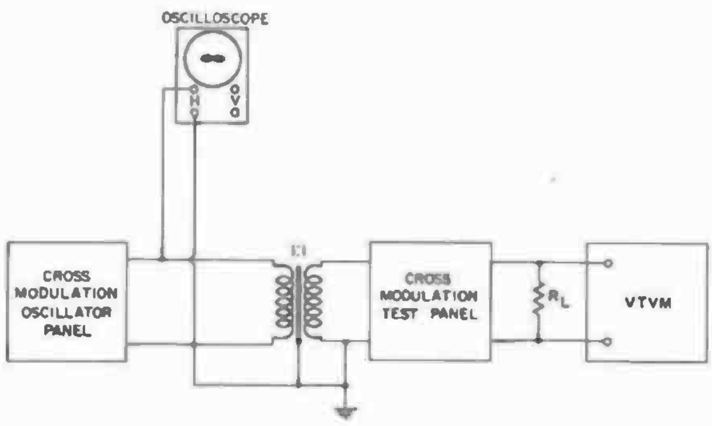


Fig. 23-185A. Connections for measuring the cross-modulation products and percentage of modulation of a cross-modulation oscillator used for photographic film recording.

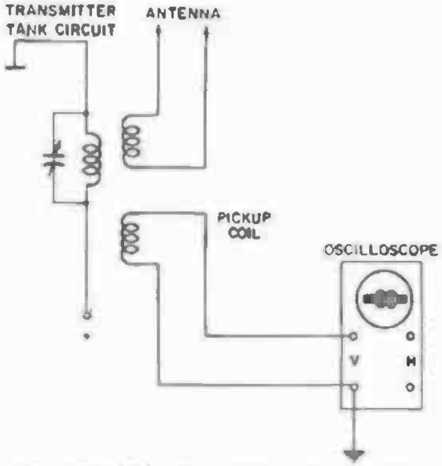


Fig. 23-186A. Connections for measuring the percentage modulation of a radio transmitter using a pickup coil near the tank circuit.

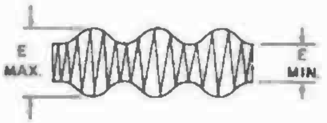


Fig. 23-186B. An rf carrier with modulation from setup of Fig. 23-186A.

lated radio-frequency carrier will appear as shown in Fig. 23-186B.

The second method consists of connecting the pickup coil to the vertical plates of the oscilloscope and connecting the horizontal plates to the modulation transformer of the radio transmitter to obtain an audio-frequency signal. (See

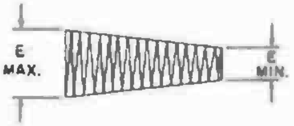


Fig. 23-186D. Trapezoidal modulation pattern from setup of Fig. 23-186C.

Fig. 23-186C.) This results in a trapezoidal pattern as shown in Fig. 23-186D.

23.187 ***How may the sound-transmission loss of a motion picture screen be measured?***—By placing the screen in front of a theater-type loudspeaker system and measuring the acoustical sound level at a distance of 10 feet from the center of the screen, using a sound-level meter such as described in Question 22.94.

After the level has been measured and noted, the screen is removed from in front of the speaker system and the acoustical level again measured. The difference between the two acoustical levels is the transmission loss of the screen at a particular frequency.

Although there is no standard at the present time for such measurements, good engineering practice would indicate that the loss at 6000 Hz should not exceed 2 dB with not more than 4 dB at 10,000 Hz, with reference to 1000 Hz. When such measurements are made, precautions must be taken to prevent the formation of standing waveforms in the area of measurement.

23.188 ***How may the frequency response of a loudspeaker be measured***

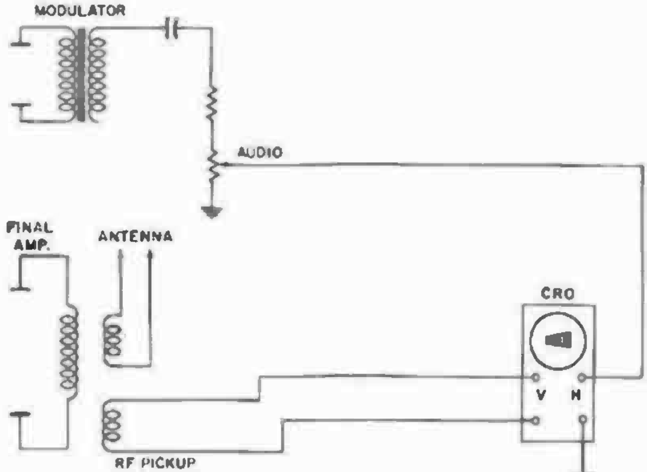


Fig. 23-186C. Connections for measuring the percentage modulation of a radio transmitter using a trapezoidal pattern.

without an anechoic chamber?—In the absence of an anechoic chamber, the loudspeaker to be used as a standard of comparison is set up in a room fairly free from reflections. (See Fig. 23-188.)

A warbled tone from a warble oscillator is applied to the amplifier driving the loudspeaker. The acoustic output of the speaker is picked up by a microphone, amplified, and read on a VU meter or vacuum-tube voltmeter. The microphone should be of good quality with a wide frequency response, preferably a ribbon-velocity or capacitor type. The microphone is mounted on the center axis of the speaker, approximately 2 feet from the diaphragm.

The amplifier system must be of uniform frequency response over a frequency range greater than that to be measured from the loudspeaker. The acoustic level from the loudspeaker must be set to a value which does not overload the microphone acoustically or the preamplifier electrically.

Frequencies of interest are applied at a constant level to the loudspeaker, as indicated by the meter across the output of the amplifier. The variations in the frequency response of the loudspeaker are picked up by the microphone and read on the meter connected across the output of the amplifier. The frequency response obtained thus is plotted with reference to 1000 Hz.

The standard speaker is then removed, the one to be compared is connected in its place, and the same measurement made again. The two response measurements are then compared by centering them on the 1000-Hz reference frequency. It should be borne in mind that a frequency-response measurement obtained in this manner is not a true frequency response of either londspeaker because of reflections from the walls of the room, variations in the frequency response of the microphone, and other factors. However, if a loudspeaker of known quality is available, other loudspeakers may be compared to it and conclusions drawn.

Somewhat better results may be obtained if the measurements are made outdoors on a roof, pointing the speakers upward to prevent reflections. Distortion and sensitivity measurements may be made with the same setup. Before attempting a measurement as described, the reader is referred to Question 23.80, which describes how the frequency response of a microphone may be measured using the comparison method, as both have several factors in common.

A relatively simple method, devised by D. E. L. Shorter and G. A. Briggs, is to place a microphone about 12 inches in front of the speaker or enclosure. The microphone is then covered with heavy woolen or blankets of *Fiberglas*. Frequencies of interest are applied to the speaker and the response measured at the output of the microphone preamplifier. If the characteristics of the microphone are known, they may be taken into consideration when plotting the results.

For frequencies of 1000 Hz or higher, the microphone is left at 12 inches. However, below this frequency the microphone is placed 36 inches away from the enclosure. It is claimed by the suggesters that measurements made in this manner are within 1 dB above 1000 Hz and below this frequency within 2 dB of free-field measurements.

23.189 What is the effect of making frequency, distortion, and power measurements with an inductive termination?—Whenever an inductive termination is used for the preceding measurements (such as a speaker), the frequency characteristics will show a rising characteristic, because of the increase of the impedance of the termination with freqnency. If such terminations are used when a distortion measurement is made, it is possible to have

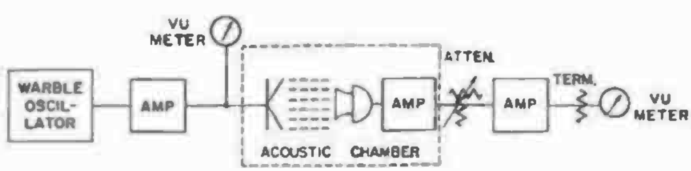


Fig. 23-188. Test setup for comparing one loudspeaker against another. This measurement is a comparison only and should not be construed as a true measurement of the characteristics of either speaker.

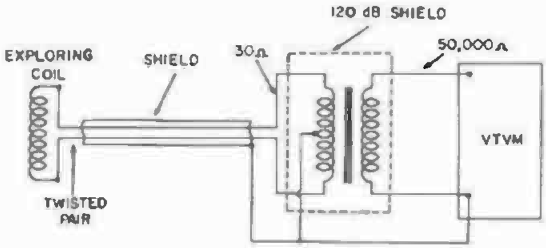
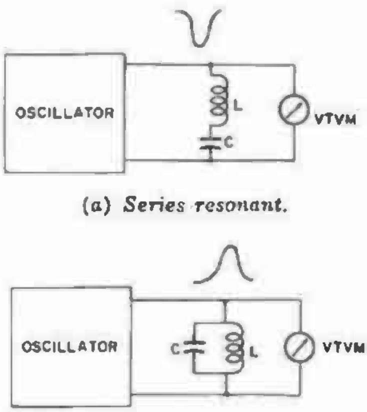


Fig. 23-190. Connections for exploring coil and meter for measuring magnetic fields.

erroneous measurements. If a capacitive load is used, the terminating impedance will show a decreasing output as the frequency is increased. Terminating resistors must be noninductive and capable of carrying the full power output of the device under test. (See Question 23-56.)

23.190 ***How can extraneous magnetic fields be explored?***—By the use of an exploring coil and vacuum-tube voltmeter connected as shown in Fig. 23-190. The exploring coil should consist of a fairly large number of turns wound on an air core. The output from the coil is carried over a twisted pair in a shield which is grounded at the receiving end as shown. The step-up coil is required to bring the voltage from the coil to a level suitable for measurement by the vacuum-tube voltmeter. A 30-ohm-to-50,000-ohm input transformer with 120-dB shielding makes a quite satisfactory transformer.

23.191 ***How may the resonant frequency of a parallel- or series-resonant circuit be measured?***—As shown in Fig. 23-191. The circuit to be measured is connected in parallel with the output of an oscillator and a vacuum-tube voltmeter. The frequency of the oscillator is varied until a dip or rise is noted on the meter. If the circuit is series resonant, the meter will dip, and if it is parallel resonant, the meter will rise to a peak. The resonant frequency is that frequency where the maximum rise or dip is obtained.



(a) *Series resonant.*

(b) *Parallel resonant.*

Fig. 23-191. Method of measuring the resonant frequency of a series- or parallel-resonant circuit.

23.192 ***How can it be determined which peak of a sine wave is being indicated on an oscilloscope?***—By connecting a battery to the input terminals of the vertical amplifier and noting the deflection of the beam. If the beam is deflected upward when the positive pole of the battery is connected to the high side of the input, the instrument indicates the positive peak of the sine wave when the trace is upwards. If the trace is deflected downward, the negative peak is indicated.

23.193 ***If a sensitive meter is not available, how may the residual hum level of an amplifier be measured?***—By measuring the hum voltage between the plate and ground of the final amplifier stage, with the secondary of the transformer properly terminated.

The hum voltage may be measured using a rectifier-type ac voltmeter, with a large capacitor connected in series with the meter to prevent it being damaged by the high dc voltage of the plate circuit. The hum power may then be calculated:

$$P_{HUM} = \frac{E^2}{Z_t \times Z_r}$$

where,

E is the voltage measured between the high side of the primary and ground,
Z_t is the primary impedance,
Z_r is the impedance ratio of the output transformer.

The result of the preceding measurement is the hum level in milliwatts at the output-transformer primary winding. The hum power is then converted to decibels:

$$dB = 10 \operatorname{Log}_{10} \frac{P_1}{P_2}$$

where,

P_1 is the hum level in milliwatts,
P_2 is a reference level (generally 1 milliwatt).

23.194 How may the internal noise of a vacuum tube be measured? — As shown in Fig. 23-194. Normal operating voltages and a resistive load are applied to the tube. The measurements may be made using either alternating or direct current on the heater. However, a higher noise level may be expected when alternating current is used.

The vacuum-tube voltmeter is connected to the plate circuit through a large coupling capacitor C to isolate the meter from the effects of the dc plate potential. The plate-load resistance R_L is equal to the load resistance recommended by the manufacturer. The grid resistance R_g must not exceed that specified for the tube in question. The noise is measured in millivolts.

23.195 Describe a standard voltage cell and its use.—The primary voltage standard is the Weston standard voltage cell, developed by Edward Weston in 1892 and manufactured by the Weston Instrument Co. (Figs. 23-195A and B). The cell is of the saturated type, intended for use under controlled conditions as a primary standard of reference for the volt. Such cells are generally used in banks of five or more to permit crosschecking. The average voltage for a single Weston saturated cadimum cell is 1.01863. Since the saturated-type cell has a temperature coefficient of approximately 50 microvolts per degree centigrade, it is customary to hold the temperature to a selected value around 30 degrees centigrade by means of an air or oil bath. Under these conditions the cell may be used as a precise standard with a high degree of stability. Voltage-correction tables for

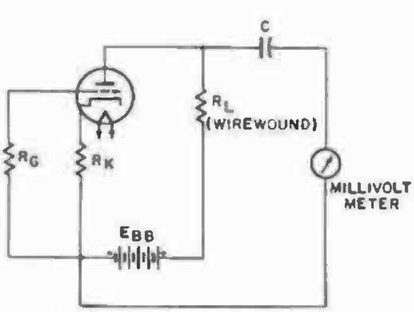


Fig. 23-194. Circuit for measuring the internal noise level of a vacuum tube.

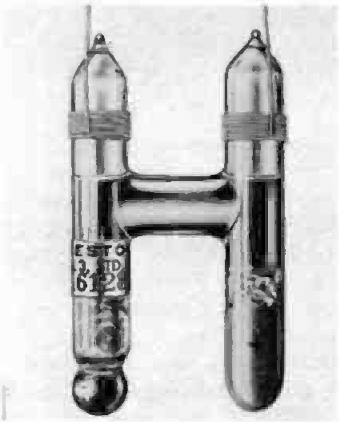


Fig. 23-195A. Weston Standard voltage cell, saturated cadmium type.

Fig. 23-195B. Weston Standard voltage cell, unsaturated cadmium type.

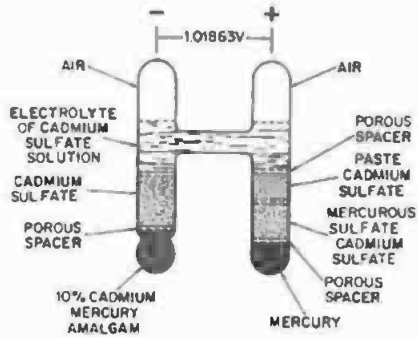


Fig. 23-195C. Cross-sectional view of a Weston Standard cell showing the chemical elements used in its construction.

temperatures between 20 degrees and 40 degrees centigrade are furnished with each cell.

Standard cells are labeled with their voltage, as determined in the laboratory before shipment, by comparison with a bank of similar cells, cross checked with a bank of cells at the United States National Bureau of Standards. By international agreement, the 1.000330 is in absolute volts.

In Fig. 23-195C is shown a cross-sectional view of a cell, calling out the chemical elements used in its construction. When the cell is in use, it is balanced against a variable voltage to prevent the drawing of current from the cell. If an appreciable amount of current is drawn from the cell (not more than 50 microamperes), it is left standing idle for several hours. The voltage will then return to normal. Also, if the cell is moved around, it must stand idle before being used. The internal resistance of a single cell is 100 to 500 ohms. Any device calibrated from these standard cells is termed a *secondary standard*. (See Question 23.196.)

23.196 How may a standard voltage be obtained for comparison purposes?—In applications involving the standardization of ac and dc voltages, two methods are available: the nse of a standard voltage cell to which an unknown voltage is compared, or the generation of a standard voltage accurate enough for the direct calibration of a meter.

Calibration laboratories generally make use of a precision potentiometer device similar to that of Fig. 23-196 for voltage comparisons. By international agreement the absolute volt equals 1.00030. Precision potentiometer boxes used in the United States indicate international volts.

To use the decade voltage box illustrated, the voltage across resistor R2 is first set to a value of exactly 1 volt by the calibration control R1. The unknown voltage is applied to the dc input terminals and compared to the standard voltage by adjusting the decade resistances R3 to R7 for a zero indication on the zero-center galvanometer M. The unknown voltage at null may be computed by the formnla:

$$E = \frac{R_2 + R_3 + R_4 + R_5 + R_6 + R_7}{R_2}.$$

The push button in series with the meter is used only when the decades are adjusted for a null indication on the meter. Calibration resistor R1 is a screwdriver adjustment and requires only infrequency adjustment. This control is set for exactly 1-volt drop across resistor R2 with the dc input terminals open. Input voltages up to 100 volts may be measured using the circuit shown. The accuracy of the measurement depends on the accuracy of the decade box and the voltage across R2.

A mercury cell may be used as a source of standard voltage, since the voltage of 1.345 volts remains fairly constant throughout its useful life. (See Question 24.96.)

Extremely accurate reference-voltage power supplies are also available for calibration purposes, having an accuracy of 0.01 percent, with a stability of 0.001 percent per day, temperature coefficient of 3 ppm/°C from 0°C to +50°C, and with a hum and noise level of minus 100 dB.

23.197 How may the characteristics of a vacuum tube be displayed on the

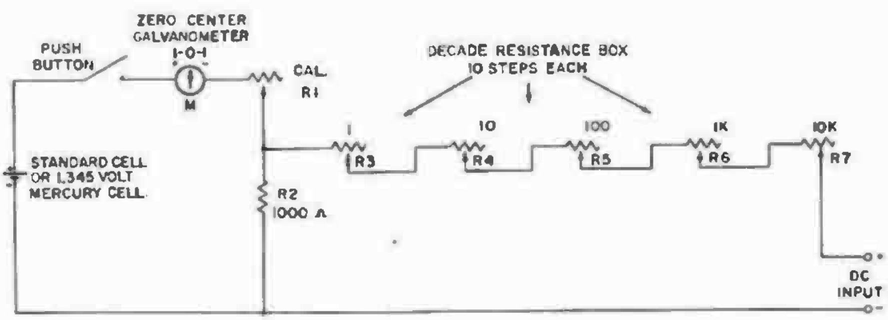


Fig. 23-196. Standard voltage potentiometer box for determining an unknown voltage by comparing it against a standard source of voltage.

screen of a cathode-ray oscilloscope?—By the circuit shown in Figs. 23-197A and B. In Fig. 23-197A, the tube is self biased by means of fixed resistor R_k in the cathode circuit, with the correct value for the applied plate voltage.

A source of variable bias voltage is applied to the control grid by means of a potentiometer, P. A small transformer with a secondary voltage of approximately 6 volts is connected in the ground side of potentiometer P. The internal sweep oscillator in the oscilloscope is turned to its off position. By holding the plate and control-grid voltages constant and swinging the control grid to cutoff by injecting an ac voltage in series with the control grid, the tube characteristic is traced on the oscilloscope.

Setting the dc control-grid bias voltage to different values will permit a family of curves to be traced. Plate-voltage/plate-current characteristic curves may be traced by holding the grid bias voltage constant and varying the plate voltage by means of an ac voltage, as shown in Fig. 23-197B.

The ac voltage injected at the plate of the tube must be of such a value that the plate current is driven to zero on the negative swings of the ac voltage. Pentode characteristics may be traced in a similar manner, except that a steady source of screen-grid voltage must be supplied.

If a long-persistence type of cathode-ray tube is used in the oscilloscope, several traces may be made at different control-grid voltages and the traces held for several minutes for photographing.

23.198 How is the frequency response of a photocell measured using a chopper wheel?—As shown in Fig. 23-198. A chopper wheel consists of a disc pierced near its periphery by a number of small holes. The wheel is placed between a source of light and the photocell which is connected to a preamplifier and a vacuum-tube voltmeter. The chopper wheel is driven by a variable-speed motor. The number of holes in the wheel depends on the speed range of the motor and the frequency hands to be covered. The light source should be fed from a source of regulated direct current.

Precaution must be taken that the light source does not overload the photocell and that the voltage applied to the photocell is of the correct value.

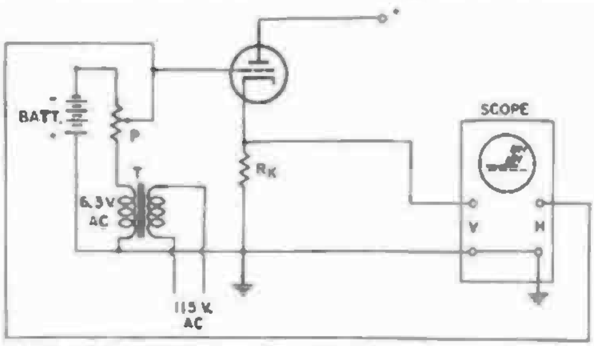


Fig. 23-197A. Circuit for displaying the grid-voltage versus plate-current characteristics of a triode tube on an oscilloscope.

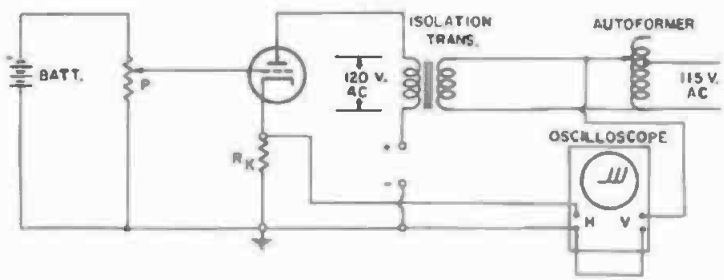


Fig. 23-197B. Circuit for displaying the plate-voltage versus plate-current characteristics of a triode vacuum tube on an oscilloscope.

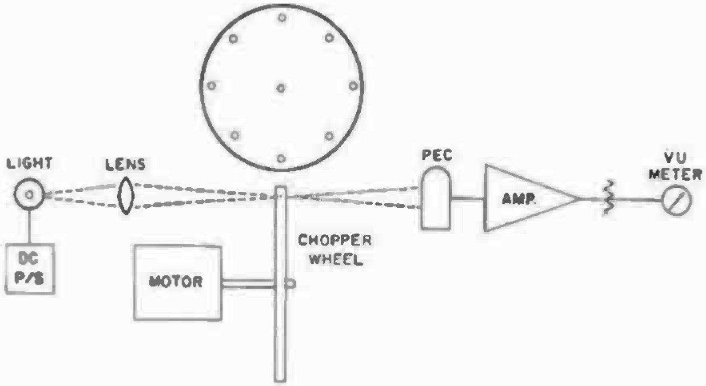


Fig. 23-198. Measuring the frequency response of a photocell by the use of a chopper wheel.

The frequency measurements are made by establishing a reference frequency by regulating the speed of the motor to produce a given frequency. The balance of the measurements are made in a similar manuer. The output voltage of a photocell is directly proportional to the amount of light falling on its elements.

23.199 Give the design of an adapter for measuring the frequency characteristics of a phototube preamplifier.—Fig. 23-199 shows a circuit suitable for measuring the frequency response of a preamplifier. The adapter consists of the two resistors and a capacitor contained within the dotted lines. The input of the adapter is fed from a 1:1 repeat coil. The output of the adapter plugs into the normal phototube connections to the amplifier as shown.

Because of the high impedance and gain of a phototube amplifier, it is necessary to place the resistors and capacitor of the adapter in a shield. If the amplifier employs an input plug, the circuit elements may be mounted inside the shell of a plug.

The gain of an amplifier measured using the adapter with the values shown is not the true gain because of the voltage divider circuit formed by the two resistors R1 and R2. To obtain the true gain, a correction factor is used which may be calculated:

$$dB = 10 \operatorname{Log}_{10} \frac{R_1}{R_2}$$

where,
R_1 is the shunt resistor,
R_2 is the series resistor.

The value of R1 is always such that it supplies the correct terminatiou for the repeat coil; thus, for a 250- or 600-ohm coil, the terminating resistor would be one of these values. (See (g) of Fig. 23-14D.) Resistor R2 is generally between 1 and 2 megohms.

The true gain of the amplifier is the measured gain using the adapter, plus the insertion loss of the adapter. The circuit shown may also be used for

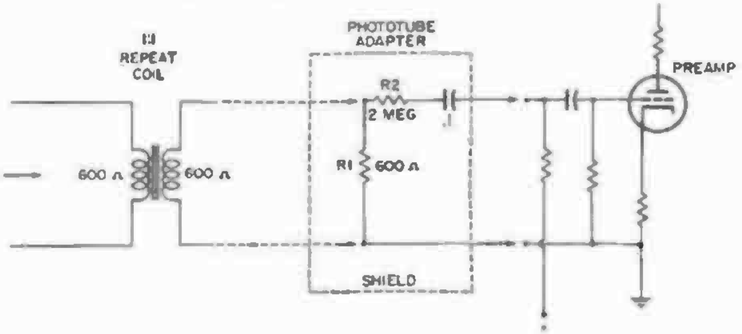


Fig. 23-199. Input adaptor used for measuring the frequency characteristics of a phototube preamplifier.

making harmonic or intermodulation measurements.

23.200 Describe the procedure for adjusting the frequency response of a record-reproducing channel using a frequency test record.—Several different makes of test records are available for adjusting the frequency response of a reproducer chain. During the original recording of program material, the recording channel is adjusted for a given recording characteristic known as the RIAA recording characteristic. (See Fig. 13-95 and Fig. 13-110.) Test records also use this characteristic. When this characteristic is reproduced, the reproducer must have a frequency response inverse to the recording characteristic to produce a flat frequency response.

The lateral velocity of a recording stylus may be expressed as:

$$2\pi fa$$

where,

f is the frequency,
a is the amplitude swing of the recording stylus.

Therefore a constant-velocity recording has increasing groove amplitude with decreasing frequency. Phonograph records are normally recorded with the low frequencies reduced in velocity (constant-amplitude region) and the high frequencies increased in velocity (constant-velocity region). Decreasing the velocity of the lower frequencies during recording is done to limit excessive lateral excursions of the recording stylus and to make more efficient use of the recorded area of the record. When such a record is reproduced, the low frequencies are restored to their original amplitudes by a low-frequency post-equalizer.

The increase of the velocity of the higher frequencies during recording increases the signal-to-noise ratio during reproduction. When the record is reproduced, the amplitude of the high frequencies is reduced by a high-frequency post-equalizer in the reproducer circuit.

To properly equalize a reproducing channel, a VU meter or vacuum-tube voltmeter is connected across a resistive termination at the output of the reproducing power amplifier (the termination must be resistive, not a speaker). The output level is adjusted for a normal listening level at the 1000-Hz reference frequency of the test record.

As the record is played back, the equalizer controls of the reproducer system are adjusted for a flat response as read on the meter. If the overall frequency variation is within plus or minus 2 dB from the lowest to the highest frequency of the record, the response may be considered to be satisfactory. If a compromise must be made, the variation between 100 and 8000 Hz should, if possible, be held to within plus or minus 1 dB. After the correct frequency response has been obtained, the only deviation should be that required to compensate for room acoustics and speaker characteristics.

For monitoring in commercial installations, the reproducer equalizers are fixed and are complementary to the recording characteristic. Corrections required for room acoustics or speaker characteristics are connected in the monitor system. This leaves the equalization fixed and supplies a standard of listening quality.

In home reproducing equipment the equalization is included in the pickup preamplifier and is therefore fixed. However, if the test record is played back with the variable controls set for a flat frequency response, the measured response should be well within the plus or minus 2-dB limits.

23.201 What are the techniques for measuring the characteristics of equipment located at a point remote from the transmission-measuring equipment? —If the power capabilities or distortion characteristics are to be measured on an amplifier located at a point some distance from the transmission equipment, the output of the amplifier must be terminated in a noninductive load resistor at the output terminals of the amplifier. This will eliminate the dc resistance of the transmission line and prevent a power loss or a mismatch of impedances.

Transmission lines are, as a rule, composed of number 20 or 22 wire and are not designed to carry any amount of power. When the amplifier is terminated as stated, only the voltage across the output termination appears on the transmission line and is measured by the VU meter of the transmission set. In some instances, the termination in the transmission set will cause the amplifier to oscillate because of the long line. Feedback between the transmission-line pairs will also cause oscillation.

If the input of the amplifier is grounded, a repeat coil should be connected between the send terminals of the gain set and the amplifier (if the gain set does not contain one permanently). If the amplifier is designed for bridging, a termination is connected at the input terminals of the amplifier. Connecting the termination at the amplifier is not always necessary if the amplifier is measured frequently. It may be terminated at the transmission set after an initial measurement has been made to determine if the frequency characteristics are affected.

Before making any measurement at a remote point, a turnover test (see Question 23.58) should be made to determine if any unbalance in the test setup exists. A good point to remember in making remote transmission measurements is that the ground at the remote point must be separated from any grounds at the transmission equipment. This is accomplished by inserting a repeat coil in the send circuit.

Noise measurements are sometimes rather difficult to make on equipment at a distance from the noise-measuring set. If difficulty is experienced, the best way is to terminate the output of the equipment at its terminals and measure the noise with a vacuum-tube voltmeter.

Equalizers and filters are measured in the manner described in other parts of this section. Precautions must be taken to eliminate ground loops and to insert repeat coils where necessary, because such devices are prone to pick up a considerable amount of noise.

The measurement of a complete recording channel light modulator or a similar device generally does not present any particular problems and may be accomplished using the techniques described elsewhere in this section.

23.202 ***How is the frequency response of a wave filter affected if the test oscillator has a high percent of harmonic distortion?***—If an 80-Hz high-pass filter is under test and the test oscillator has strong harmonics at 40Hz, the frequency response below 80 Hz will be affected. If the oscillator harmonics at 40 Hz are appreciable, the response at 40 Hz will never be down more than the amplitude of the harmonics at the 40-Hz frequency.

To ensure the correct frequency response being measured, the test oscillator should not have more than 0.25 percent total rms harmonic distortion.

23.203 ***Describe a visual method for checking diode rectifier units.*** — The characteristics of zener diode aud regulator diode rectifiers may be displayed on an oscilloscope as shown in Fig. 23-203. The horizontal plane of the oscilloscope represents the voltage developed across the diode, while the junction voltage is displayed in the vertical plane. Typical displays for good and bad diodes are given below the circuitry. The forward conduction trace may be eliminated by connecting a gate diode in series with the 50-ohm resistor.

23.204 ***Describe the procedure for measuring the dc breakdown voltage of a zener diode.***—The diode is connected as shown in Fig. 23-204. The current through the diode is adjusted by means of resistors R1 and R2, which are fed from a source of constant-current. The diode current is read on meter M1. As the voltage is increased beyond the specified breakdown, the enrrent in-

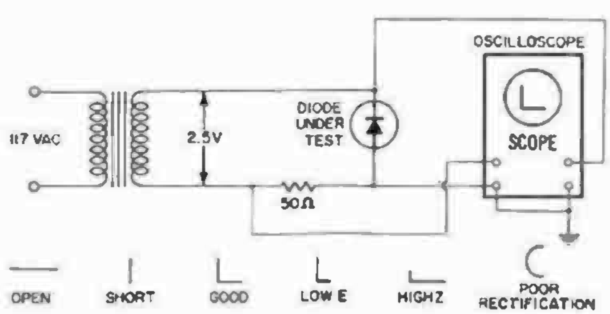


Fig. 23-203. Circuit for measuring the rectification qualities of a semiconductor diode rectifier.

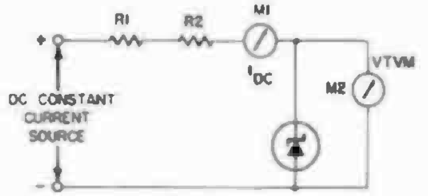


Fig. 23-204. Circuit for measuring the dc breakdown voltage of a zener diode.

creases rapidly with very small increases of voltage. Therefore the breakdown voltage is nearly constant over a considerable current range. The current is maintained through the diode at its rated value, and breakdown voltage is read on meter M2 across the diode.

23.205 How is dynamic breakdown impedance of a zener diode measured?—The breakdown or dynamic impedance of a zener diode is the small-signal impedance at a specified value of direct current in the breakdown region of the diode. A small value of rms current is superimposed on the dc current, and the resulting voltage drop across the diode is measured, as in Fig. 23-205. Dynamic impedance is an important consideration in diode regulators, since it establishes the reverse volt-ampere characteristics of a diode. A low impedance ensures better regulating action.

The value of the dynamic impedance varies with junction current and the diode rating. For example, a 27-volt, ¾-watt zener has a breakdown impedance of 50 ohms, while one for the same voltage rating but with a 1-watt power rating has a breakdown impedance of 28 ohms.

In dynamic impedance measurements, the diode current is set for 20 percent of the maximum diode current rating as read on meter M1, by adjusting the resistors R1 and R2 in conjunction with the dc supply voltage E_{in}. An ac voltage from transformer T1 is applied through R3, C1, and the impedance of the zener diode. The ac current is adjusted for a value of 10 percent of the dc current, and is read on meter M2.

When these conditions have been established, the ac voltage developed across the junction is read on ac voltmeter M3. The dynamic impedance may now be computed:

$$\text{Dynamic Impedance} = \frac{V_{ac}}{I_{ac}}$$

where,

V_{ac} is the ac voltage across the diode,
I_{ac} is the ac current through the diode.

23.206 How is an oscillator employing a cathode-follower output coupled to an external circuit?—Oscillators employing a cathode follower and voltage divider network in the output are generally of the economical type and offer several difficulties in coupling to an external load circuit, where an input transformer is concerned.

In (a) of Fig. 23-206 is shown the output circuit of an audio oscillator, using a cathode-follower tube in the output stage and a voltage divider network for adjusting the output voltage in conjunction with the variable control of the oscillator-amplifier generator section. Output circuits of this type generally develop about 10 volts of signal voltage and are designed primarily for use with equipment employing a high-impedance input, such as resistance-coupled amplifiers.

Three effects are noted when this output circuit is used. First, if the voltage divider control is set to a value below maximum, and an input transformer of 150 to 600 ohms of impedance is connected across the output, the shunt resistance of the voltage divider is 100 ohms in parallel with the transformer primary. This can have serious effects on both the frequency and distortion characteristics of the transformer.

Second, when a circuit of low impedance is connected in parallel with the voltage divider, the harmonic distortion of the cathode-follower stage is

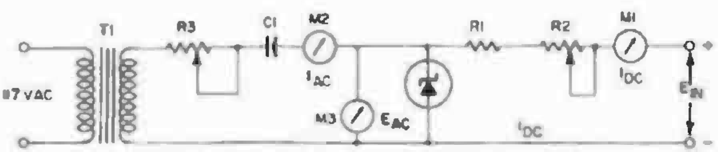


Fig. 23-205. Circuit for measuring the dynamic breakdown impedance of a zener diode at a specified value of direct current.

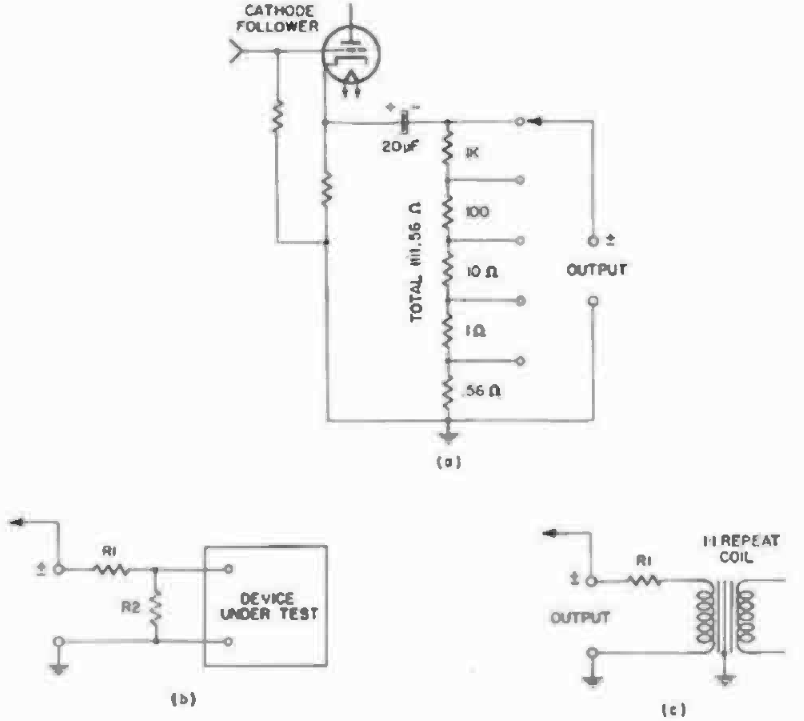


Fig. 23-206. Output stage of an oscillator using a cathode follower. (a) Normal circuit. (b) Circuit for increasing the output impedance or presenting a given impedance. (c) Use of a repeat or isolation coil.

increased because of the loading effect of the voltage divider, with the low impedance of the input circuit (in parallel) being reflected back to the cathode circuit. This is most noticeable in the 1000-ohm maximum output position.

A third effect, though not always serious, but one that must be taken into consideration, is leakage current through the electrolytic capacitor to ground through the voltage divider. Although the current is quite small its presence can affect the frequency and core material characteristics of certain types of transformers. Only an electrolytic capacitor of low-leakage characteristics can be used for coupling between the cathode circuit and the upper end of the voltage divider.

To avoid the effects just mentioned, a 600-ohm resistor R1 in (b) of Fig. 23-206 is connected in series with the output to prevent the external circuit from seeing less than 600 ohms as the switch is moved from the top of the divider network to lower values. If a true 600-ohm source impedance is required, an additional resistor R2 is connected in shunt with the input circuit. It should be remembered that when these additional resistors are connected, the available signal voltage from the signal generator is reduced.

Another effect sometimes encountered is the increase of oscillator distortion when the output control of the oscillator is turned toward maximum output. This is caused by possible overloading of the output stage, and it can be avoided to some extent by operating the output control at its lowest position in keeping with the required output voltage.

When the output of the oscillator is connected directly to a resistance-coupled amplifier or one having a ground on one side of the input circuit, the low potential or ground side of the oscillator must be connected to the grounded side of the input circuit. If hum or noise is encountered, it can sometimes be eliminated by reversing the ac plug, either to the oscillator, amplifier, or both. Noise and hum frequencies can also be eliminated by connecting a 1:1 repeat coil between the oscillator output and the input of the equipment under test, as shown in (c) of Fig. 23-206.

The output circuit discussed should not be confused with oscillators using *ladder-type output attenuator networks*, as such networks are designed to present a constant output impedance. Networks of this design have little or no effect on the oscillator distortion, regardless of the setting of the output control.

23.207 How is instantaneous peak power of an amplifier related to sine-wave power?—If the amplifier output signal is of sine-wave character (0.5 to 2 percent), the instantaneous peak power is 1.414 times the rms value of the sine wave. As power is proportional to voltage squared, the instantaneous peak power is 1.414^2, or twice the sine-wave power. Thus, an amplifier with a power output of 50 watts has an instantaneous peak power of 100 watts. This calculation is often used by manufacturers to rate their amplifiers in peak power.

Although the preceding discussion from a theoretical standpoint is correct, in practice it is not generally correct. To measure the actual peak power of an amplifier, the internal power supply voltage would have to remain at its no-signal value without variation, regardless of the demand for additional current as the power output is increased. If this condition can be satisfied (and it never is), then the peak-power rating may be stated by doubling the power output.

In the original Institute of High Fidelity (IHF) Standard A-200-1958, the constant-voltage method was recommended for the measurement of peak power. This method is still in the new Standard A-201-1966 and may be used if desired.

A much better method is given in the Standard using tone-burst techniques, which results in a measurement of the amplifier actual peak power, using its internal power supply. (See Questions 23.208 and 23.209.)

23.208 Describe the method used for measuring music power output of an amplifier.—Music power-output amplifier measurements are made under somewhat similar conditions occurring when the amplifier is playing complex waveforms, as might be found when reproducing music and speech waveforms. Such measurements are made by applying short bursts of a 1000-Hz sine-wave input signal of low distortion (0.10 percent or lower) to the amplifier. The rise time of the tone burst or modulator should approximate that of the envelope rise time of speech and music (for 10 to 90 percent of input voltage), which is on the order of 10 to 20 milliseconds. (See Question 22.74.)

A distortion-factor meter (DFM) is connected across the amplifier load resistor and an oscilloscope to the output terminals of the DFM. The sine-wave output of the DFM (100-percent calibrate position) is used to calibrate the peak-to-peak deflection of the oscilloscope in terms of steady-state DFM indication. For convenience, the oscilloscope graticule may be calibrated to read directly in terms of power output. The turn-on voltage (dc) of the modulator is used to trigger a single horizontal sweep of the oscilloscope. The test circuit is given in Fig. 23-208A.

The input signal is modulated by means of a simple device consisting of a light source and a photoresistive element assembled in a light-tight case. The photocell is a polycrystalline semi-

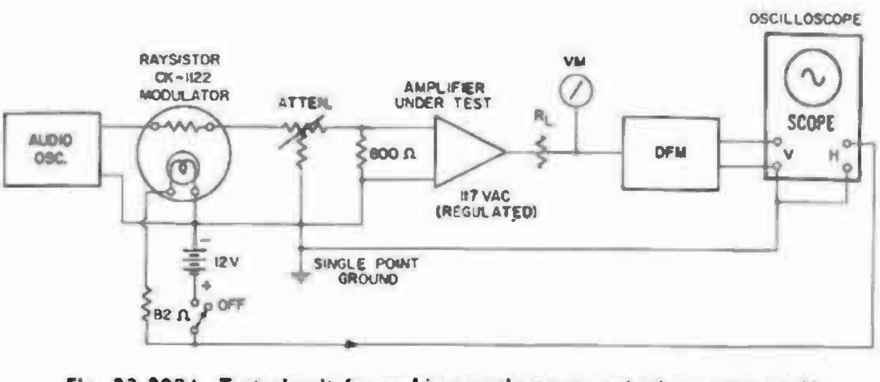


Fig. 23-208A. Test circuit for making music power output measurements.

conductor with a light source consisting of a small incandescent lamp. Advantage is taken of the time delay offered by the lamp filament in coming up to full brilliancy, to acquire the 10- to 20-millisecond rise time corresponding to an average speech or music rise-time envelope. In its dark condition the photocell has about 2-megohm resistance and in the light about 200 ohms, thus providing a range of approximately 60 dB from dark to light. (Some leakage occurs through the 2-megohm resistance in the dark condition.) This device connected in series with the input provides a convenient method of control for the input signal. The appearance of the input signal at the output of the modulating device appears as shown in Fig. 23-208B. The assembled photocell device is shown in Fig. 23-208C. It will be observed that the rise time (Fig. 23-208D) is rapid, while the decay time trails out. A tone-burst generator may be substituted for the modulator, provided it can be adjusted for the 10- to 20-millisecond rise time. (See Question 22.139.)

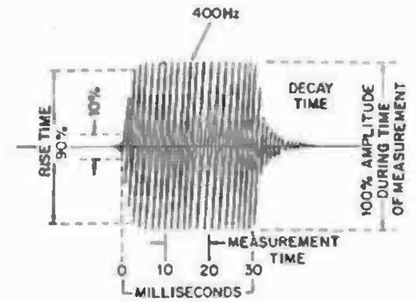


Fig. 23-208B. Appearance of signal at output of modulator.

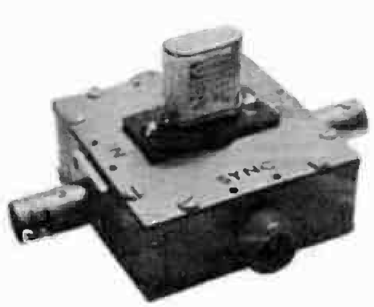


Fig. 23-208C. Tone-burst unit used for measuring dynamic amplifier distortion.

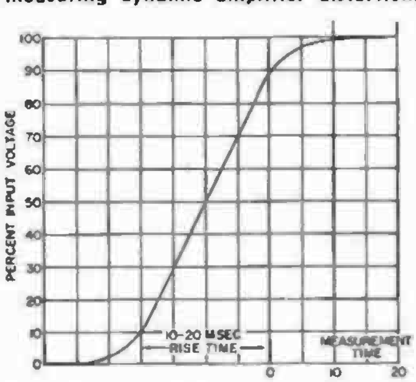


Fig. 23-208D. Envelope shape of modulated sine-wave signal. The distortion and power output measurements are made during the last 10 milliseconds.

The oscillator signal, after passing through the modulator unit, is applied to an attenuator and then to the amplifier. The output of the amplifier is terminated in its specified load resistance with a noninductive resistance capable of carrying the full power output of the amplifier. A voltmeter connected across the load resistance indicates the output level, and a DFM and oscilloscope complete the circuit. Line voltage must be supplied from a well-regulated source with not more than 2 percent of THD. (See Question 23.210.)

After establishing the sine-wave output level (this is accomplished by turning on the pulse for a constant output) with the DFM set to calibrate (sine-wave 100 percent), the oscilloscope input attenuator is adjusted for a reference output power. The DFM is then set to distortiou and a measurement is made in accordance with the normal procedure for the particular DFM being used.

The modulator switch is now closed and the distortion is measured using the 10- to 20-millisecond period of the pulse as indicated (Fig. 23-208B). Observing the character of the oscilloscope display will reveal several factors not generally known about amplifiers, such as the effect on the output signal for a heavy surge. Here the signal may overshoot and then drop and rise again, or it may have the appearance as in Fig. 23-208E. At periods where the power supply fails to deliver the required voltages with a consequent drop in output power, the storage and recovery times of the filter capacitors may be readily observed.

Since a DFM may generate a low-frequency transient with a rapidly ris-

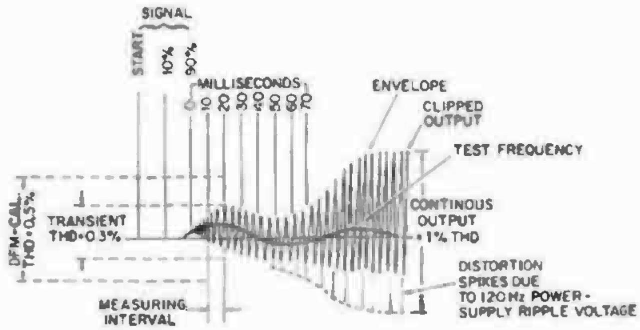


Fig. 23-208E. Distortion waveform appearance at output of distortion meter, as viewed on an oscilloscope.

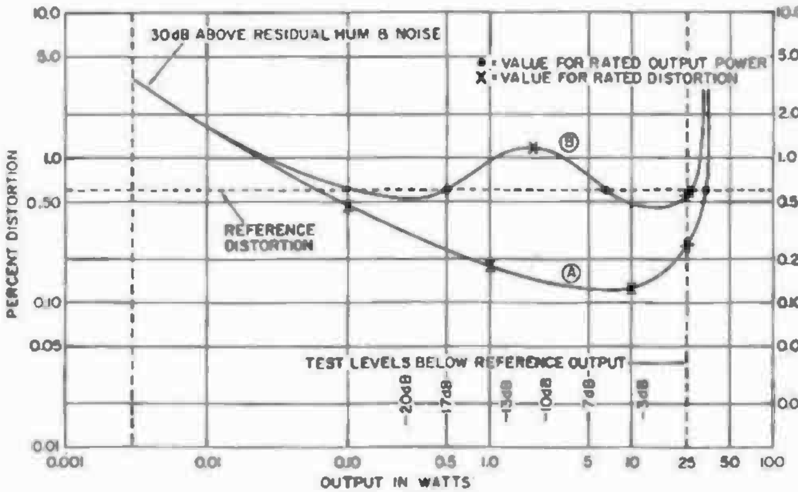


Fig. 23-208F. Power output and distortion curves for two imaginary amplifiers. The dashed lines are the reference power and distortion at which the two amplifiers will be tested. The values have been plotted in accordance with procedures outlined in IFH Standard A-201-1966.

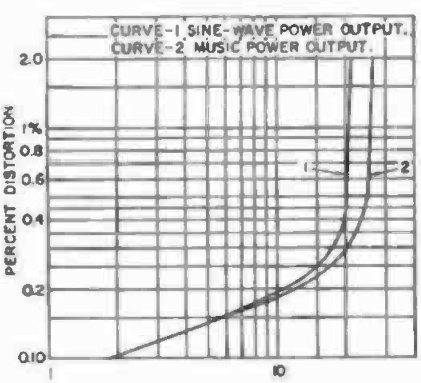


Fig. 23-208G. Harmonic distortion versus power output. Curve 1 is the steady-state output using a sine wave. Curve 2 is the output in terms of music power.

ing input signal and may also indicate the modulated signal sidebands as distortion, the vertical peak-to-peak deflection of the oscilloscope between 10 and 20 milliseconds after reaching 90 percent of the maximum input voltage is taken as the *measured distortion*, provided that the THD of the input signal is sufficiently low. The output signal level of the amplifier is similarly measured in the same time interval. Care must be taken that the transients generated due to turn off of the signal by the modulator have decayed to a negligible amount before proceeding to the next measurement. This method of measurement assumes that primarily only one harmonic distortion component is created by the amplifier and therefore a peak-to-peak reading is equivalent to a DFM reading.

Four terms used frequently with music power measurements are reference distortion, reference output distortion, rated distortion, and rated output power. Reference power and reference distortion are values claimed by the manufacturer, and rated power and rated distortion, all the values found from continuous power or music power measurements, are set forth in Institute of High Fidelity Standard (IHF) A-201-1966.

Two terms used in conjunction with determining the rated output of an amplifier are continuous power output, and transient distortion. Continuous power is the output power an amplifier is capable of delivering for 30 seconds with a sine-wave input signal. Transient distortion is a measurement of power output, taking into account the change in power-supply voltages under heavy signal demands and its recovery time. It has been found that amplifiers can deliver from 20 to 30 percent more power under music-power tests than for a continuous sine-wave signal. It is not unusual for an amplifier to deliver considerably more music power for a short time, with less distortion; however, a low-frequency transient is created by the power-supply ripple or modulation and is measured as distortion. Thus, the dynamic rating of an amplifier depends on its continuous power-output capabilities and transient distortion measurements. The measurement that results in higher distortion or lower power output determines the final rating of the amplifier. (See Question 23.209.)

The rated power *bandwidth* of an amplifier is defined by the two frequencies (low and high) where the curve of distortion versus frequency taken at a level 3 dB below the reference output crosses the reference distortion line. (See Fig. 23-7J.)

In plotting the results of music power output measurements graphically, a somewhat different method of presentation is used (Fig. 23-208F). Here are shown two imaginary amplifiers A and B with identical reference values (claimed by the manufacturer) and the characteristic found after measurements. Both amplifiers were rated 0.6 percent of THD at 25 watts of output. The manufacturer's values are entered on the graph as shown by the dotted lines. It will be noted that 3×5-cycle log-log paper is used for plotting rather than the conventional semilog paper. Using log-log paper allows the distortion characteristic to be plotted at the lower operating levels and permits the effects of crossover distortion in transistor amplifiers to be more easily presented. Measurements are made at 3, 7, 10, 13, 17, and 20 dB (and lower) below the reference power output level (for this example 25 watts), at frequencies between 20 and 20,000 Hz.

For continuous power-outpnt measurements (constant sine wave), the signal is to be applied for not less than 30 seconds. Resulting measurements for both continuous output and transient output are plotted in the same manner as given in Fig. 23-208F. The measurements are to extend from a point 30 dB above the residual hum and noise to five times the reference distortion.

The rated continuous output is taken at the intersection of the continuous output-versus-distortion curve with the reference distortion line. Referring to the plot of amplifier A, the distortion and power output claimed by the manufacturer meet the reference distortion and power output; therefore the amplifier is rated at the point where the reference and the measured distortion cross. However, for amplifier B the measurements indicate it failed to meet the manufacturer's claim, and that *crossover distortion* is introduced by the class-B output stage. This is indicated by the rise in distortion between 1 and 7 watts of output.

It is not unusual to find in a well-designed amplifier that the music power output is 10 to 30 percent higher than for the sine-wave power condition Fig. 23-208G illustrates the difference in the power output under *transient* conditions and *continuous* output conditions. Curve 2 indicates about 20 percent more power output is obtained at 0.5 percent of distortion than for curve 1, the same distortion using a steady-state sine wave. (See Questions 23.209 and 23.210.)

23.209 Why does an amplifier tested for music power output develop greater power than when tested using a continuous sine wave?—In the operation of any amplifier, the power developed in the various stages is supplied by a power transformer whose voltage is rectified, filtered, and then applied to the transistors or vacuum tubes in the amplifier. Since all components in the power-supply circuits have resistance and resistor voltage-dividing networks are used, there is a considerable voltage drop in the power supply before the operating voltages are applied to the amplifier stages.

When a signal is applied to the input of an amplifier and the various stages increase their current demands, the no-signal dc voltage drops in the voltage divider circuits are increased, thus decreasing the operating voltages from their no-signal parameters. Therefore the amplifier stages do not always operate in the most desirable portion of their characteristics, and as a result lower power is developed in the output stage, and distortion is increased.

The ideal type of power supply would be one that has no dc voltage drop and in which the operating voltages remained at the no-signal value, regardless of the current demands. In practical amplifiers this situation is not possible. However, some of the difficulty is overcome by the fact the filter capacitors (if large enough) during the periods of no- or low-current demands are charged to almost the operating voltage, and as the signal changes in amplitude and the current demands increase, the filter capacitors supply the additional current, thus reducing the voltage variations to some extent.

If a constant-amplitude sine wave of low distortion is applied to the input of an amplifier, and the input level is increased to the point of maximum power output, the dc operating voltages will decrease and the distortion will increase. However, if the same signal is applied in short bursts to the input (tone burst), the current demands are of short duration, and the demand for additional current is supplied by the filter capacitors in the power supply. Therefore the amplifier power output is affected very little.

The fact is that amplifiers used for the reproduction of speech and music do not operate with steady-state sine waves, but with complex waveforms of speech and music. The current demands are for very short periods, and the filter capacitors can generally supply the additional current demands. Therefore amplifiers tested using a steady-state signal (continuous sine wave) do not result in a true picture of the amplifier operating capabilities. This is why the IHF Standard specifies amplifiers are to be rated as to both their *music power output and continuous power output.*

Because of the difficulty in making distortion measurements on amplifiers, considerable research has been done in the art of tone-burst measurements. Amplifiers with a given power rating for continuous output with a sine wave, will often show from 10 to 30 percent more music power output with tone-burst techniques. It is the policy of most manufacturers to rate their amplifiers for both continuous and music power output. The subject of music power measurements is discussed in Question 23.208. (See Question 12.230.)

23.210 Describe the effect of a changing line voltage on the power output, harmonic and intermodulation distortion of an amplifier.—Amplifiers as a rule are designed for a given power output at a given line voltage. The ef-

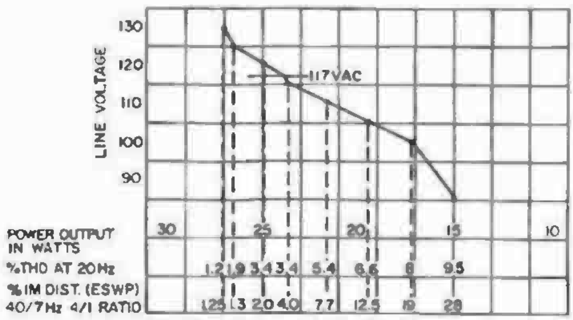


Fig. 23-210A. The effect of a changing line voltage on the power output of an amplifier at 20 Hz. The harmonic and intermodulation distortion were measured simultaneously with the change in line voltage.

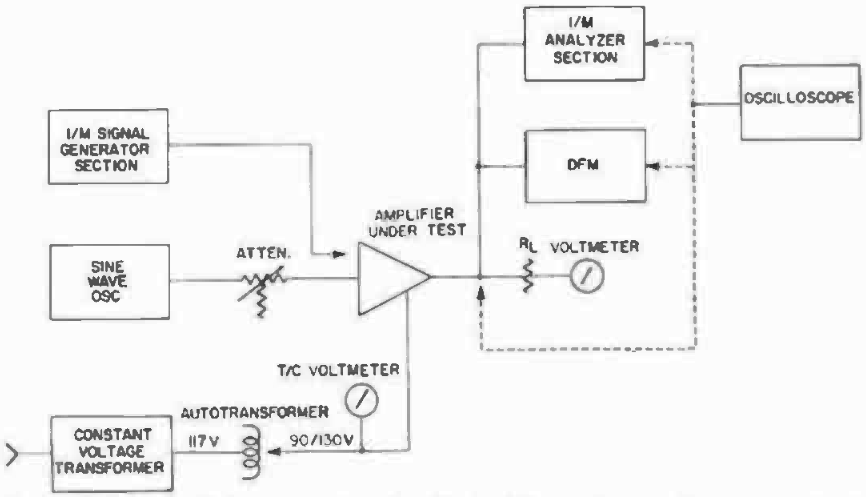


Fig. 23-210B. Circuit for measuring the effect of line voltage on the power output distortion of an amplifier.

fects of a changing line voltage are not too noticeable above a frequency of 100 Hz. However, below this frequency if the line voltage is reduced, the power output of an amplifier can drop off at a surprisingly rapid rate and the distortion is increased as evidenced by the curve shown in Fig. 23-210A.

To illustrate the procedure for making such a measurement, it will be assumed that a 25-watt amplifier is to be measured. The setup for this discussion will be as shown in Fig. 23-210B. The line voltage is set to 117 volts, using a meter of at least 1-percent accuracy and of the thermocouple type if available. The line voltage is adjusted by means of a variable-voltage transformer fed from a constant-voltage transformer. This latter transformer should be of the low-harmonic type. The constant-voltage transformer is not an absolute necessity, but it does help to smooth out small variations in the line voltage. If it is used, its waveform distortion should not exceed 2-percent THD. A 1000-Hz signal is applied to the input of the amplifier and the level is adjusted for a 25-watt output. At this point, a turnover test is made to determine if any unbalance in the test circuit exists, as discussed in Question 23.58.

The output termination must be within plus or minus 1 percent of the rated output load resistance and must be capable of carrying the full output of the amplifier without changing its resistance when hot. This subject is discussed further in Question 23.56. The voltage across the termination for 25 watts may be computed:

$$E = \sqrt{P \times R} \quad \text{or} \quad P = E^2/R$$

where,

P is the output power in watts,
R is the load resistance in ohms.

At a line voltage of 117 volts and an output of 25 watts, the harmonic distortion is measured at 20 Hz. The line voltage is now reduced to zero, and the amplifier is allowed to cool for about 5 minutes. The line voltage is then *slowly* brought up to 90 volts and the amplifier is permitted to warm up thoroughly. The output level is then computed, and the harmonic distortion is measured.

If external equipment such as an oscilloscope, distortion-factor meter, or a voltmeter are left connected permanently across load termination R_L, the actual load resistance is the impedance of the various instruments in parallel with the load resistance. The exact value may be calculated by the use of Ohm's law for parallel resistances. (See Question 25.127.)

Throughout this series of measurements, the line voltage should never be allowed to overshoot, but should be brought up to its correct value slowly so as not to heat the circuit or charge the filter capacitors above that normally produced by a low line voltage. If there is any doubt as to the accuracy of the measurement, the line voltage should be reduced to zero, the amplifier permitted to cool off, then the voltage again brought up to the desired value and the amplifier allowed to warm up until the output reaches a maximum. The line voltage may then be slowly increased to the next higher value.

After the harmonic-distortion measurements have been completed, the same procedure is followed as for making intermodulation tests. Precautions must be taken to use equivalent sine-wave voltage for setting the output power when intermodulation measurements are made, as the amplifier can be badly overloaded. A table of equivalent sine-wave power voltages is given in Fig. 23-113B. A study of the results obtained in this manner indicates that the output cannot be obtained from an amplifier operating under conditions where the line voltage is an appreciable amount below that specified by the manufacturer.

The drop-off in power at the lower frequencies is due to the lower plate and heater voltages. In a transistor amplifier this is also true, and although there are no heaters, the voltage to the collector will be lower and this affects the overall gain.

The problem of voltage regulation is not an easy one to overcome, and to do so increases the cost of manufacturing, plus adding weight and components not in the normal amplifier. To correct for wide voltage variation a plate-filament transformer of constant-voltage design may be substituted for the regular power transformer. However, even this does not completely satisfy the situation. Voltage regulator tubes or zener diodes and thermistors are required to complete the regulation.

Although the previous discussion paints a rather black picture for varying line voltage unless the amplifier is being used to supply a constant output, it may not be of too much importance, particularly in the home. Where the amplifier is used in a critical position, steps must be taken to regulate the incoming line voltage and the internal circuitry of the amplifier. (See Question 23.7.)

23.211 How are amplifier stability tests made?—Stability tests are to determine how stable an amplifier is under normal operating conditions. The amplifier is first operated without an input signal. The output load resistance and the input resistance are varied from 1/100 of the normal values to open circuit and monitored with an oscilloscope across the output load termination for signs of oscillation or an increase in hum and noise. Similar tests are made using a capacitance ranging from 100 pF to 10 μF, and with inductive values from 10 μH to 1 H. The oscilloscope used for these tests must have a bandwidth of at least 2 megahertz or greater. If the amplifier passes these tests, it may be rated to be stable under a no-signal condition. The same tests are then repeated, with an input signal ranging from 20 to 20,000 Hz.

Generally, oscillations (if any) will be observed above 20,000 Hz. This may be checked by the connection of a 20,000-Hz high-pass filter between the output of the amplifier and the input to the oscilloscope while a low-frequency signal is applied to the input. Low-frequency oscillation is checked by the use of a low-pass filter in the output, with a high-frequency signal applied to the input. Caution must be observed that power supply ripple or distortion components of the input signal are not mistaken for oscillation. The input impedance of the filter network is considered to be a part of the load impedance. If the amplifier passes the latter test, it may be rated as unconditionally stable under signal conditions.

23.212 Describe the measurement of damping factor.—Damping factor is a measurement of the regulation of an amplifier, and it is an indirect measurement of the output impedance. It is the ratio of the output voltage (under standard conditions) to the measured output voltage change when the output load is removed. Damping factor is measured between 20 and 20,000 Hz to determine the effect of frequency. It is measured at a given reference output level and then at several successive 10-dB steps below the reference output level. This latter test will bring to light any effect due to change in operating levels. The results are plotted on 3×5 log-log paper to show effects of frequency with respect to damping factor.

The rated damping factor is that value measured at 1000 Hz at a reference output level. Regulation is the in-

verse of damping factor, expressed in percent. (See Question 20.103.)

23.213 How are amplifier sensitivity measurements made?—The sensitivity of an amplifier is a measure of its ability to produce a given power output for a given signal voltage at the input. The measurement is made by applying a signal to the input and measuring the power output for each increase of input voltage up to the rated power output. The result is plotted as input voltage versus power output, as shown in Fig. 23-213. The plot in Fig. 23-213 is for an amplifier capable of developing 100 watts of power output with a total harmonic distortion of less than 1 percent.

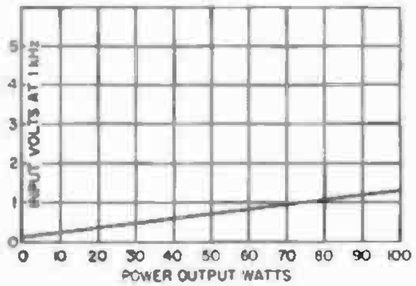


Fig. 23-213. Sensitivity for a 100-watt power amplifier, plotted input volts at 1000 Hz versus power output.

23.214 How are the effects of line noise measured?—The method specified in IHF Standard A-201-1966 is shown in Fig. 23-214. The amplifier is connected as for a normal noise measurement. An audio oscillator is connected by means of a step-down transformer in series with the low-potential side of the ac line. A signal of 2 volts from 20 to 20,000 Hz is induced into the line, while the noise level is observed on the voltmeter across the output. The signal-to-noise ratio is computed with and without the induced signal.

The step-down transformer secondary must be capable of carrying the full load current of the amplifier when it is developing full output power. This same method of measuring line noise may be used for any type of instrument and at frequencies above the audio-frequency spectrum. Line filters are discussed in Question 7.101.

23.215 Describe how separation tests are made on stereophonic amplifiers.—The term "separation" is generally associated with stereophonic radio receivers and phonograph pickups. However, this term is also used to rate the degree of separation (leakage) of signals between the two sides of a stereophonic reproducing system. Separation may be defined as the ratio of a wanted signal to an unwanted signal. This would be analogous to applying a signal to one side of the system and measuring the leak-through at the output of the other side of the system. The ratio of the two signals is the measure of separation and is expressed in decibels. Such measurements are generally made at frequencies of 20 to 20,000 Hz, and commonly called crosstalk.

As an example, consider a stereophonic preamplifier and two 40-watt power amplifiers, using a common power supply (Fig. 23-215A) are to be

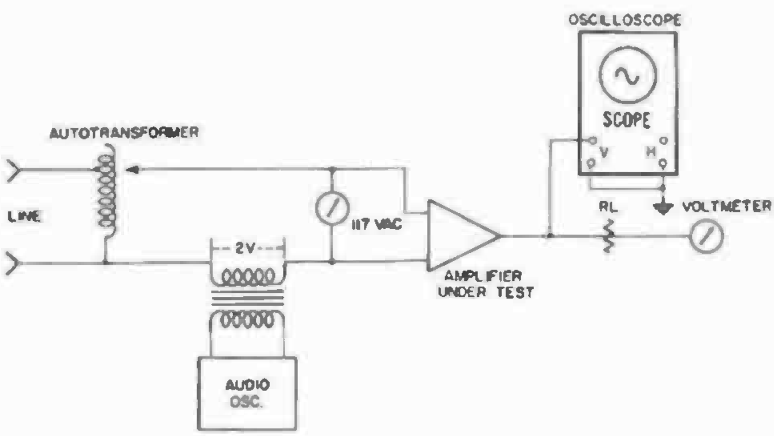


Fig. 23-214. Circuit for measuring the effect of line noise as specified by IHF Standard A-201-1966. The transformer for inducing the signal must be in the grounded side of the line.

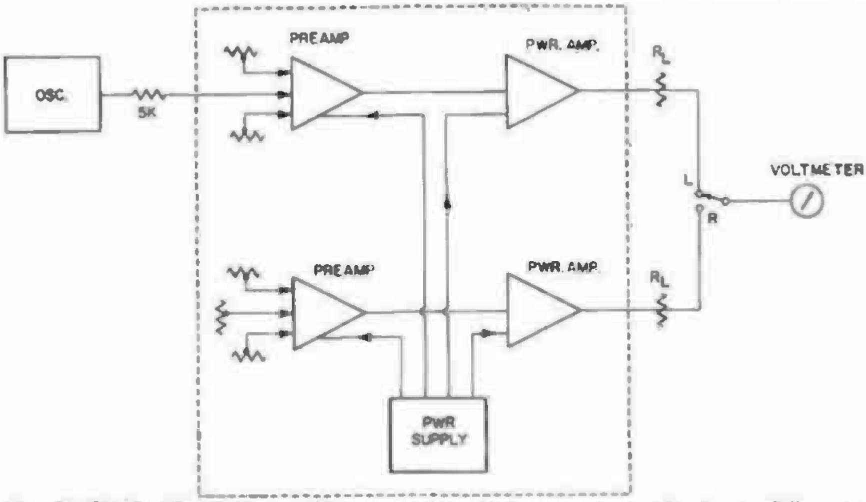


Fig. 23-215A. Circuit for making separation tests on a stereophonic amplifier employing two preamplifiers and two power amplifiers fed from a common power supply.

measured for their separation characteristics. A signal (1000 Hz) from a low-distortion oscillator is applied to the input of the left preamplifier and the output at the power amplifier is set to the reference output power of the amplifier (reference power output specified by the manufactnrer; see Question 23.208). After the reference output is set, the oscillator output is reduced 10 dB. This second output is termed *reference level.* The voltage across a 16-ohm load resistor for a power of 40 watts is 25.3 volts. Reducing the input signal 10 dB reduces the power output level to 4.0 watts, or 8.0 volts.

The voltmeter is now transferred to the output of the right-hand amplifier and the leakage voltage is read. Assuming this voltage to be 0.10 volt, the ratio of the two voltages is 8.0/0.01, or 800/1. The ratio in dB is

$$dB = 20 \, Log_{10} \frac{8.0}{0.01}$$

$$= 20 \times 2.904 = 58.08 \, dB$$

The connections are now reversed between the two amplifiers, and the leakage for the other side is measured. The results are plotted in a manner similar to that shown in Fig. 23-215B. The rated separation as taken at 1000 Hz for this particular measurement is 58 dB, which includes the leakage in the preamplifiers. The irregularity in the separation curve is due to internal leakage caused by common coupling through the impedance of the power supply, stray capacitance, wiring, and ground connections.

A portion of the leakage indication may also be hum and noise. Therefore,

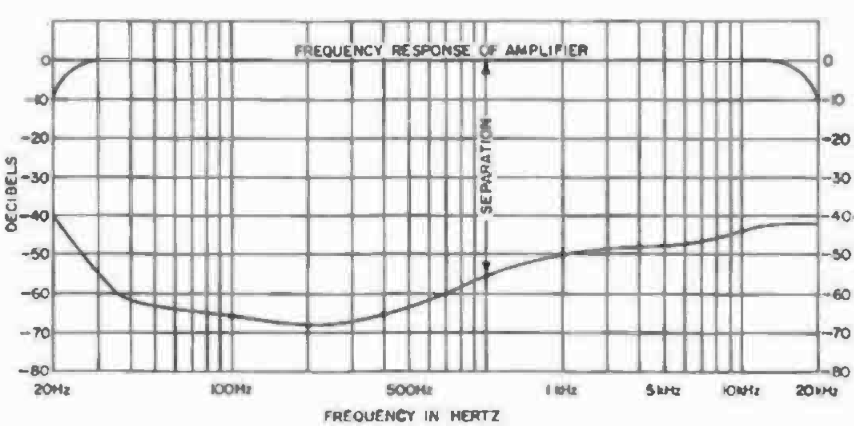


Fig. 23-215B. Typical separation characteristic for a stereophonic amplifier assembly.

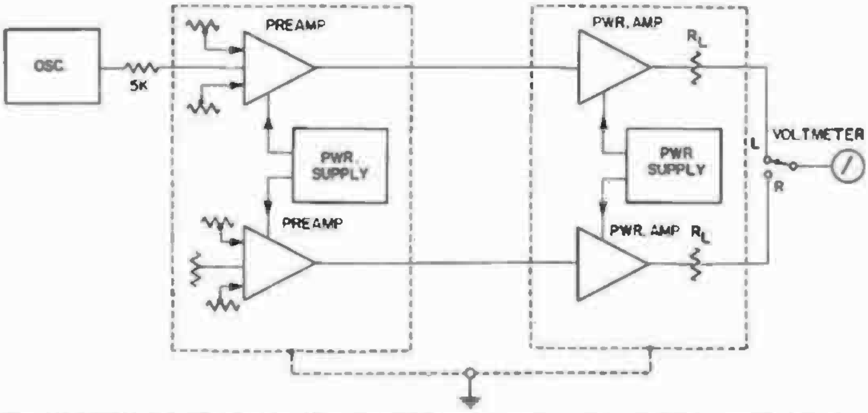


Fig. 23-215C. Circuit for making separation tests on a stereophonic amplifier employing a common power supply for the preamplifiers, and a separate power supply for the power amplifiers.

the noise should be measured before a separation measurement is made. Deflection of the voltmeter because of internal noise may be checked by turning off the inpnt signal and noting if any change occurs in the leakage signal. An amplifier indicating a separation of 50 dB or better is considered satisfactory.

Stereophonic systems using separate preamplifiers and power supplies and two power amplifiers employing a common power supply are shown in Fig. 23-215C, and a dual power-amplifier assembly using a common power supply is shown in Fig. 23-215D.

Separate preamplifiers are measured by terminating their several inputs in 5000 ohms and the output circuits in 100,000 ohms, shunted by a capacitor of 1000 pF to simulate the cable capacitance (Fig. 23-215E). The signal from the oscillator is applied to the left input through a 5000-ohm series resistor. The output level is set to the voltage specified by the manufacturer, and then reduced 10 dB. The input to which the signal has been applied is then made inoperative, and the leakage measured at the ontput of the same channel by opening each of the gain controls in turn for each input of that particular side. This test may also be made using the power amplifiers and measuring the leakage in their outputs. However, in this instance any leakage encountered in the power amplifiers will be included in the overall measurement. In this latter measurement the terminating resistors and capacitors are omitted. The results of the test for a separate preamplifier are plotted as shown in Fig. 23-215B. (See Question 25.212.)

In the instance of two power amplifiers each having their own supply, the leakage would be negligible. Where two power amplifiers are fed from a

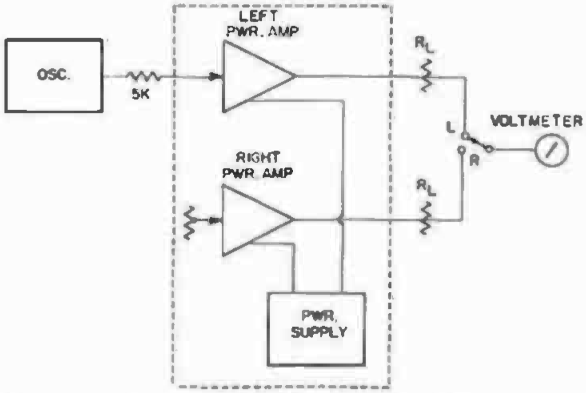


Fig. 23-215D. Test circuit for measuring the separation between two stereophonic amplifiers, operated from a common power supply.

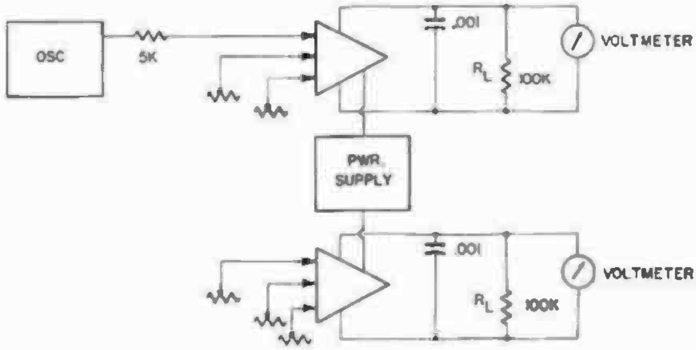


Fig. 23-215E. Test circuit for measuring the separation between two stereophonic preamplifiers using a common power supply.

common power supply, leakage may be encountered. Amplifier systems employing a common power supply for both the preamplifier and power amplifiers are more prone to leakage than those using separate supplies.

The lack of proper separation in a stereophonic reproducing system has little effect on the quality of reproduction. However, it does cause a shift of the stereophonic image. What is most important is the leakage of distortion products from one channel to the other.

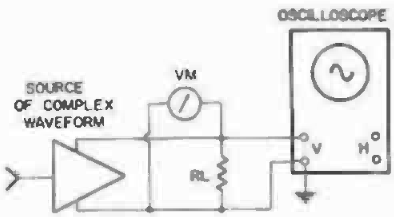


Fig. 23-216. Circuit for determining a multiplying factor for a voltmeter when measuring complex waveforms.

23.216 How may an rms-responding meter be calibrated to read true rms voltage?—When using a voltmeter (vacuum-tube or transistor) calibrated to read the rms value of a sine wave for measuring complex waveforms, the reading thus obtained is not a true measure of the voltage. To obtain a true reading a voltmeter (as discussed in Question 22.98) is used, or the meter in question may be connected in combination with an oscilloscope as shown in Fig. 23-216.

The complex waveform is applied to the vertical input of the oscilloscope, with the rms voltmeter connected in parallel with the source of the complex waveform. The meter is set for a reference voltage, and the oscilloscope adjusted for a reference deflection. The complex waveform is then removed and the internal oscilloscope calibrator (or external one) adjusted for the same deflection as the complex waveform. A ratio for the particular waveform under measurement can then be established from the calibration voltage. It should be realized that the ratio thus obtained is valid only for the particular type of waveform under observation, and the voltage measured is the peak-to-peak value.

23.217 How are ac-line ground loops avoided between test equipment and devices under measurement?—Because of the Underwriters' requirements, all portable electrical equipment in the United States and Canada must be equipped with a three-pin ac plug, with the center terminal connected by means of a third wire to the electrical ground system, and the other end of the third wire connected to the equipment being served. When such a ground is used with electronic test equipment it is possible for a ground loop to be created between equipment under test and the test equipment, as shown in Fig. 23-217.

Two ground connections shown can cause ground currents through two pieces of equipment or more, as indicated by the dashed lines. Since electrical-system grounds always have considerable noise currents through them, this noise is induced in both the equipment under test and the test equipment. Also, such ground connections are often the cause of oscillations and unstable operations. (See Questions 24.31 to 24.34.)

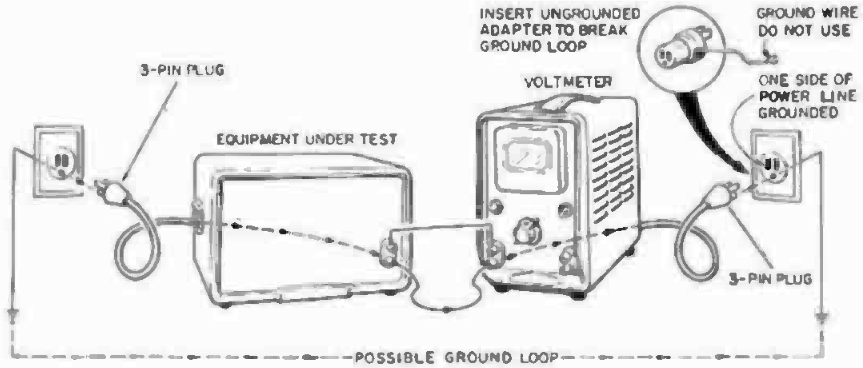


Fig. 23-217. Procedure for avoiding ground loops caused by ac line grounds.

If the test equipment happens to be a millivoltmeter, although the ground current may be only a few microamperes, this small current can result in considerable error in the meter readings. Ground currents can be avoided by the use of an ac adapter plug, shown in the illustration. The use of these plugs also avoids the possibility of a short circuit to the hot side of the ac line when instruments with the circuit ground tied to the chassis are being used for measuring the line voltage.

When any type of measurement is conducted, the ac line plugs should be reversed for the lowest indication of line leakage. Although the ground loops may be eliminated, there is still the possibility of pickup from stray magnetic fields by the leads of the test equipment. The leads should be carefully shielded and connected to the *ground of the instrument that they are used with*. Another convenient way of eliminating ground loops is to feed the whole test setup from the secondary of an ac line-isolation transformer. This will permit the grounds for several pieces of equipment to be brought to a central point, where they are then grounded to a good earth ground system.

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Installation Techniques

Installation techniques not only involve the physical installation of equipment, but also take into consideration such factors as power requirements, separation of transmission lines, grounding, isolation from magnetic fields, the reduction of noise in the system, effects of temperature, ventilation, and humidity control in certain areas. Of equal importance is the laying out of patch bays or jack fields, designing of bridging buses, relay rack and gutter design, placement of lines for power, high- and low-level transmission lines, and intercommunication and talkback systems, to mention but a few of the subjects dealt with in this section.

24.1 What is a patch bay?—In large sound installations, it is the practice to make the inputs and outputs of the principal components of the system available on jacks in a patch-bay area. This system affords a means of testing, maintenance, substitution of equipment, and special circuitry to be set np from the jacks through patch cords. In addition, transmission lines are run to various parts of the plant for remote operation of equipment, as well as to a central testing area, termed "circuit laboratory," where various types of test equipment are installed for making routine tests.

It is not uncommon for a large system to have several thousand jacks. For the most part, these jacks are of the normal type; that is, they permanently connect various pieces of equipment together to form a basic system. (See Fig. 24-9B.) Inserting a patch cord into a given jack permits any piece of equipment to be picked np and connected into another part of the system, or to be removed entirely. Typical patch bays for recording consoles are shown in Figs. 9-22D, and 9-46B and E.

24.2 What is a jack field? — The same as a patch bay. (See Question 24.1.)

24.3 What is a single-circuit cord plug?—A plug, such as the one shown in Fig. 24-3A, consisting of a brass sleeve and tip. A small rod connects to the tip and is carried back through an insulated bushing to the body of the plug, where a terminal screw is provided for connection to the cord conductor. The tip is always the high-potential side of

Fig. 24-3A. Interior view of Western Electric 47A tip and sleeve cord plug.

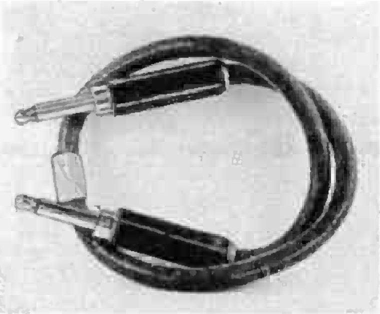


Fig. 24-3B. Single-circuit patch cord.

the circuit. If the sleeve is used, it is connected to the low-potential side of the circuit or to ground.

An insulated sleeve fits over the body of the plug and protects the inner terminals, and acts as a grip for the plug. A patch cord appears in Fig. 24-3B.

24.4 What is a double-circuit patch cord plug? — The double-circuit patch cord is practically the standard of the sound industry, particularly in large installations such as broadcast, recording, and motion picture studios, Fig. 24-4A.

The double-circuit plug was originated by Western Electric and is more commonly known as a 241A plug. An interior view of this plug is shown in Fig. 24-4B. The component parts of the plug consist of two single-circuit plugs held in a dual mounting, except the plugs float in the body to allow for variations in jack strip mountings. The insulated body of the plug carries a series of notches along one edge, to permit the plug to be polarized.

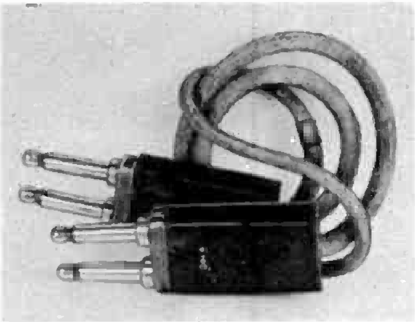


Fig. 24-4A. Western Electric 241A double-circuit patch cord.

Double-circuit patch cords are made up of two plugs, one at each end of the cord. The conductors of one end are connected to the plug tips. The opposite ends of the cord are connected to the corresponding tips, resulting in a polarized circuit. For easy identification of the polarization, the notched edge of the plug is turned to the ground or low-potential side of the circuit. An exploded view of the component parts of a double-circuit plug is shown in Fig. 24-4C.

If a shield is provided in the cord, it is connected to the frame of one plug at *one* end only. The other end of the shield is left unconnected, to prevent the formation of ground loops when the cord is used between grounded jack rows.

24.5 What is a tip, ring, and sleeve-cord plug?—This plug is similar in design to the single-circuit plug of Fig. 24-3A, except that a ring and separator are interposed between the tip and the sleeve. An interior view of a tip, ring, and sleeve plug is shown in Fig. 24-5A. The components are identified as: A, the tip; B, the ring; and C, the sleeve. A separator used only to lift the top spring of a jack assembly when the plug is inserted in the jack is iden-

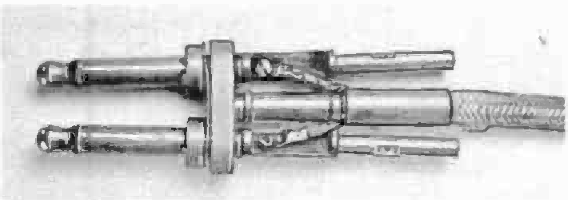


Fig. 24-4B. Interior view of Western Electric 241A double-circuit patch cord plug.

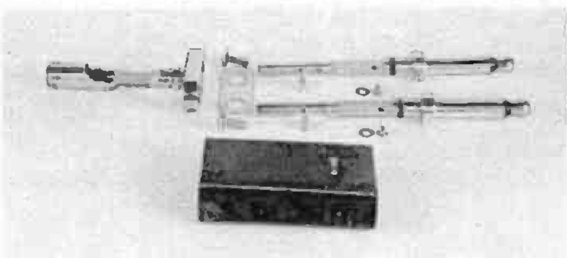


Fig. 24-4C. Exploded view of the parts of a Western Electric 241A double-circuit patch cord plug.

tified as D, and E are the terminals in the body of the plug. The insulated sleeve covering the interior also acts as a grip for inserting the plug. The tip is connected to the high-potential side of the circuit, the ring to the low potential, and the sleeve to ground.

Tip, ring, and sleeve plugs are used extensively in telephone switchboards, intercommunication equipment, telephone hand sets, and similar equipment.

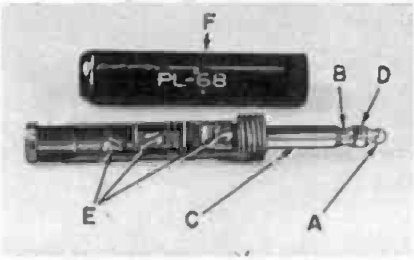


Fig. 24-5A. Interior view of a tip, ring, and sleeve cord plug: A—Tip, B—Ring, C—Sleeve, D—Spacer, E—Terminal screws, F—Insulated grip.

The tip, ring, and sleeve plug is used extensively in large recording and broadcast installations, particularly in the mixer console patch bays to reduce the number of jacks required and provide a third circuit for grounding a shielded cord. The plug used for this service is the Western Electric 310 plug or its equivalent. This plug is slightly larger than the conventional tip, ring, and sleeve plug, as may be seen in Fig. 24-5B.

Some newer as well as older installations use the double patch cord (Fig. 24-4A) as it permits the circuit to be turned over (polarity reversed), which cannot be done with the tip, ring and sleeve type without special turnover strap jacks (Fig. 24-7C) which are not always available.

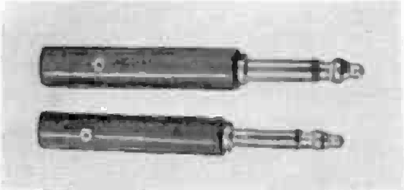


Fig. 24-5B. Western Electric 310 and 309 tip, ring, and sleeve cord plugs. The Type 310 is used in broadcast installations at the present time.

24.6 How is a jack strip constructed?—A typical jack strip is shown in Fig. 24-6A. Jack strips may be obtained in either single or double rows. As a rule, the mounting strip is made of an insulating material; however, the one pictured employs a metal strip designed for standard rack mounting, one multiple high. Several different types of mounting strips are shown in Fig. 24-6B, together with a designation strip for holding identification tabs over the jacks.

A single row of jacks consists of 24 jacks. A double row has 48 jacks. A single-row jack strip using single-circuit jacks will provide 24 circuits. If used with a double-circuit plug, it provides only 12 circuits.

24.7 What is a strap jack and how is it connected?—Strap jacks are used for connecting a number of patch cords in parallel. The manner of connecting a strap, and the type of jacks used, will be governed by the type of patch cords to be paralleled. A strap jack for paralleling a group of double-circuit patch cords is shown in Fig. 24-7A. The number of jacks in a strap should be sufficient to provide the strapping of at least four circuits, or it will be of little value.

The tip spring of the right-hand jack of a pair (facing the rack) is always connected to the high-potential side of

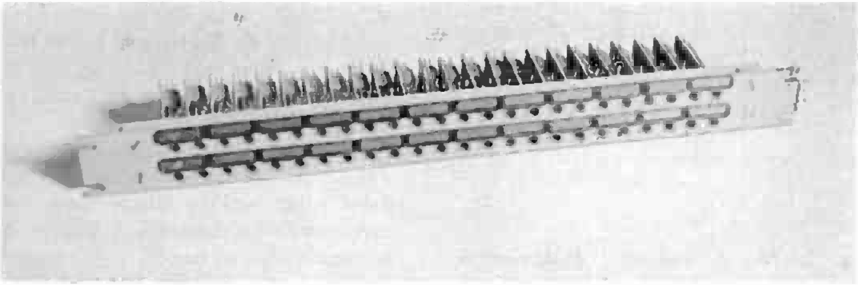


Fig. 24-6A. Double row jack strip. Mounting strip one multiple (1¾" × 19") high.

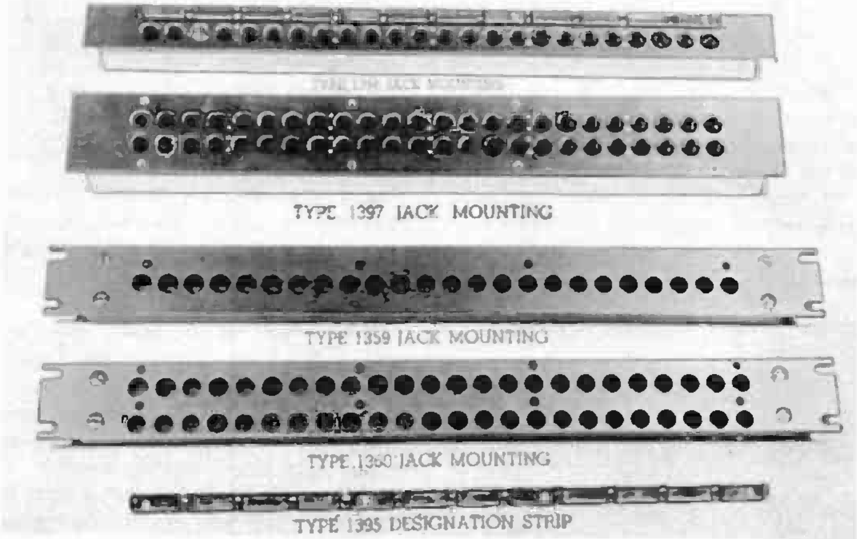


Fig. 24-6B. Circuit jack mounting strips. The two upper mountings are made of insulating material, the lower, dural, notched for standard rack mounting (19" × 1¾"). (Courtesy, Cinema Engineering Co.)

the circuit and the tip spring of the left-hand jack to the low-potential side or ground. The frame of the jack may be grounded to provide a ground for shielded patch cords. Many installations ground the jack frame even though the patch cords are not shielded. This will

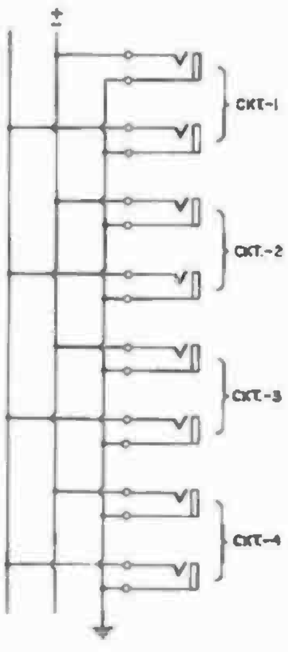


Fig. 24-7A. A 4-circuit, double-patch cord strap jack. Frames of jacks may be left ungrounded if desired.

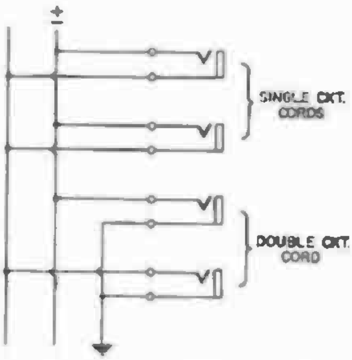


Fig. 24-7B. Strap jack for combining a single-circuit patch cord with a double-circuit cord.

provide a grounded barrier between jacks (supplied by the jack frame).

A strap jack for combining single-circuit and double-circuit cords is shown in Fig. 24-7B. Fig. 24-7C shows a strap jack for reversing the polarity of a single-circuit patch cord. It will be noted the tip connection to the lower jacks has been reversed. Fig. 24-7D is a strap connection for tip, ring, and sleeve patch cords.

Strap jacks are also used to provide a ready means of interconnecting equipment when making transmission measurements. An example might be the terminating of an amplifier in a resistive load and connecting a VU and distortion-factor meter across the output. A

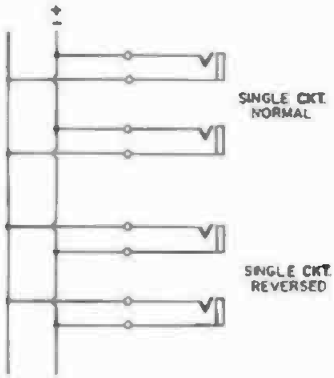


Fig. 24-7C. Strap jack for reversing the polarity of single-circuit patch cords.

four-circuit strap would be required for this type of connection.

24.8 What is an open-circuit jack and how is it used?—A schematic diagram of an open-circuit jack is shown in Fig. 24-8A. Such jacks consist only of a frame and a single spring. The spring is mounted so that only the tip of the cord plug makes contact when inserted in the jack. The frame of the jack is not used with a double-circuit plug. When used with a single-circuit plug, the frame of the jack is connected to the low-potential or grounded side of the circuit.

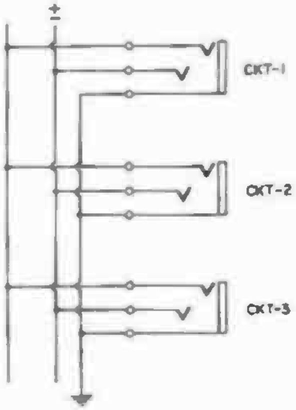


Fig. 24-7D. Strap jack for tip, ring, and sleeve patch cords.

Open-circuit jacks are generally used for connection to equipment to be patched to other parts of a system, and are not normalled to other equipment. Open-circuit jacks are employed with terminating resistors, strap-jack assemblies, multiple jacks, and similar de-

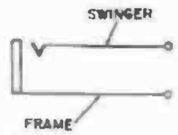


Fig. 24-8A. Open-circuit jack (tip and sleeve).

vices. An open-circuit jack and plug are shown in Fig. 24-8B.

24.9 What is a normal jack and how is it used?—The majority of jacks used in large installations such as recording or broadcasting studios are of the normal type. These jacks are employed to interconnect permanently several pieces of equipment which are always used in conjunction with each other.

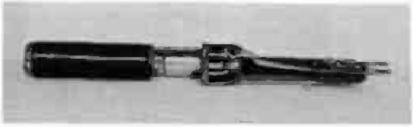


Fig. 24-8B. Western Electric 47A single-circuit plug, and 221A open-circuit jack.

Equipment connected by normal jacks may be used as a complete unit without the necessity of using patch cords, yet any one of the individual pieces connected by normal jacks may be lifted by insertion of a patch cord in the normal jack.

Fig. 24-9A shows a typical schematic for a normal jack. It will be noted the jack consists of a frame, a large spring called a swinger, and a smaller spring under the swinger called the normal spring. The smaller spring is in permanent contact with the swinger spring above. If a plug is inserted in the jack, the tip of the plug will make contact with the upper spring, raising it, and breaking the contact of the smaller spring below.

As the upper spring is connected directly to the input or output of the device the jack feeds, a direct connection is supplied to the plug for patching

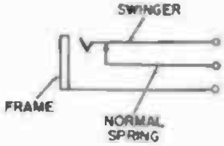


Fig. 24-9A. Normal jack (closed circuit).

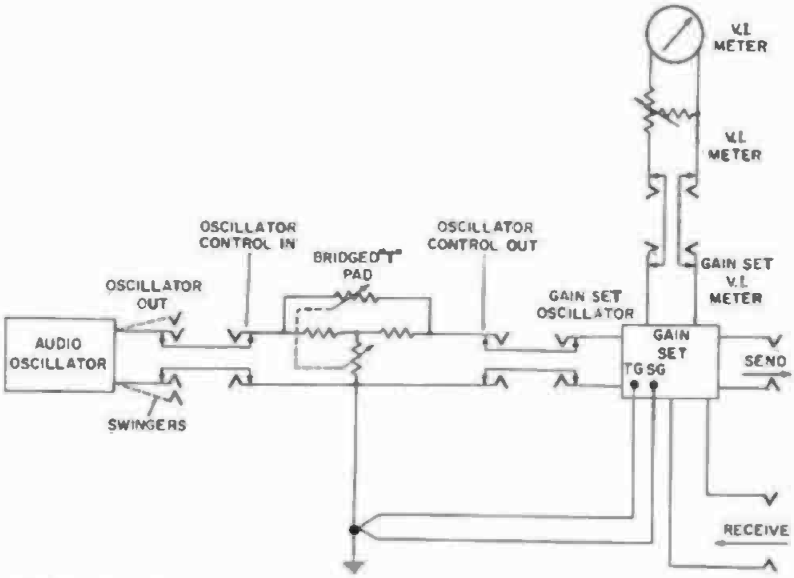


Fig. 24-9B. A typical transmission measuring setup using normal jacks for interconnecting an oscillator, oscillator control, gain set, and VU meter.

elsewhere. This feature is illustrated in Fig. 24-9B. Here, several pieces of transmission measuring equipment are shown normalled together. If a plug is inserted in the output of the oscillator jacks, the normalled circuit to the oscillator control is broken and the output of the oscillator may be patched elsewhere. A normal jack with a plug inserted to show the lifting action of the contact springs is shown in Fig. 24-9C.

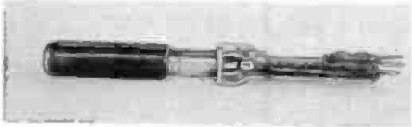


Fig. 24-9C. Western Electric 47A plug and 218A normal circuit jack.

24.10 How is a tip, ring, and sleeve jack constructed?—The spring arrangement of a tip, ring, and sleeve jack is shown in Fig. 24-10A. It will be noted this jack has two large springs, the

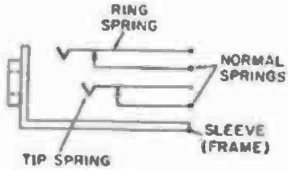


Fig. 24-10A. Spring arrangement of a tip, ring, and sleeve jack.

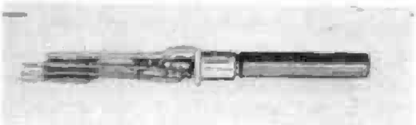


Fig. 24-10B. Western Electric 310 plug and 238A tip, ring, and sleeve jack.

upper member making contact with the ring of the plug and the lower with the tip of the plug. The frame grounds the sleeve of the plug. If the jack is to be used for normalling purposes, smaller springs make contact with the upper and lower springs in a manner similar to the normal jack described in Question 24.9. A tip, ring, and sleeve jack with a plug inserted to show how contact is made with the springs is shown in Fig. 24-10B.

24.11 What is a multiple jack?—An open-circuit jack connected perma-

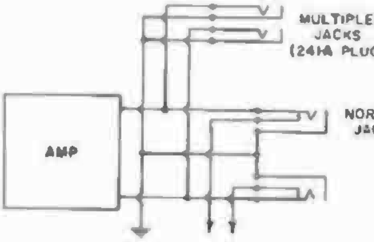


Fig. 24-11. Multiple jacks connected across two double-circuit normal jacks.

nently with a normalled circuit, as shown in Fig. 24-11.

24.12 What is the recommended method of cleaning patch cord plugs?—It is highly important that the contact surfaces of a cord plug be maintained bright and clean if it is to render trouble-free service. The contact component parts of a plug are made of brass and corrode quite easily, which increases their surface contact resistance.

Although jacks offer an automatic cleaning motion, the plug must still be cleaned frequently. It is the practice in large installations to polish patch cords at least once each month. The contact surface is polished using Bell Metal Polish, a noncorrosive, chemically neutral, polishing paste developed especially for the telephone industry, or Doe's Plug Burnishing Paste. Abrasives such as sandpaper, crocus cloth, and emery paper must *never* be used on a patch cord plug.

During the cleaning period the cord is also inspected for electrical turnovers in the conductors, using a patch cord and cable tester as described in Question 24.59. A device for clamping a double- or single-type plug in a vise while polishing the tip and sleeve is shown in Fig. 24-12.

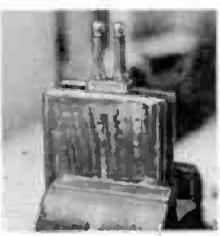


Fig. 24-12. Jig for holding patch-cord plug while cleaning.

24.13 How is the interior surface of a jack sleeve cleaned?—By use of a special brush as shown in Fig. 24-13. This brush is inserted in a drill motor and rotated while inserted in the jack sleeve. As a rule, no polishing paste is used. If it is necessary to use a polishing paste, it should be used sparingly to prevent it from getting on the surface of the jack springs.

A small metal brush is also used at times; however, this has a tendency to remove small amounts of material from the interior surface of the sleeve every time it is used and, in time, the sleeve hole in the jack is enlarged, causing a poor fit for the plug.

24.14 How are jacks connected for shielded patch cord use?—Shielded patch cords are used only with double-circuit plugs (Fig. 24-4A) and tip, ring, and sleeve plugs (Fig. 24-5A). The ground connection for the shield is obtained by grounding the frame of the jack. This is illustrated in the connection shown for strap jacks in Figs. 24-7A and D.

24.15 How is the internal ground connection to a shielded patch cord made?—Shielded patch cords are only used with a double-circuit plug (Fig. 24-4A) and a tip, ring, and sleeve plug (Fig. 24-5A). Connection to the shield is made by connecting to the frame of one plug at *one* end of the cord only. If the shield is connected at both ends of the cord, a ground loop is created when the cord is used between two rows of grounded jacks.

24.16 Which side of a rack is the low-level side?—The left-hand side when facing the front of the rack.

24.17 Which side of a rack is the high-level side?—The right-hand side when facing the front of the rack.

24.18 Which side of a jack pair is the low-level side?—The left-hand side when facing the pair from the front of the rack. Shielded pairs for making the cableforms used in racks generally consist of a red/green and red pair. The red/green conductor is always the grounded or low-potential side of the circuit (left) and is connected to the low-level jack of the pair. (See Question 24.38.)

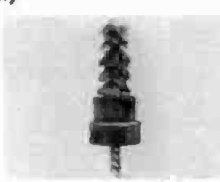


Fig. 24-13. Western Electric bristle brush for polishing the interior surfaces of a jack sleeve.

24.19 How are the operating levels designated in a cableform?—The levels carried by a cableform are designated high, low, and intermediate, and are so designated relative to a reference level of 1 milliwatt, or 0 dBm. Thus all cableforms carrying signal levels of 0 dBm and lower are called low-level, and those above 0 dBm, high-level. The intermediate cableform carries levels overlapping into the low level and the low-high levels.

Operating levels falling into this category are run in an intermediate cableform separated from both the low- and high-level cableforms as described in Question 24.21.

24.20 How are cableforms secured and placed at the rear of a relay rack? —The rear view of a typical rack used for the installation of sound equipment is shown in Fig. 24-20A. At the left is the high-level cableform and at the right the low-level cableform. Note that the positions of the high- and low-level cableforms are reversed when viewed from the rear. As a rule, cableforms are supported by tying them to short metal rods projecting from the side members of the rack, as shown at the upper right side of the rack.

Cableforms running to jack strips may be handled in two different ways. In Fig. 24-20A, the forms running to the jack strips break out and run across the strip taking in the whole group of jacks, making the strip either a high- or low-level group. In some installations it might be more desirable to split

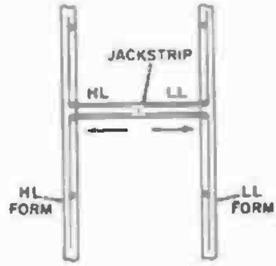


Fig. 24-20B. Rear view of high- and low-level cableforms running to a jack strip split into high- and low-level circuits.

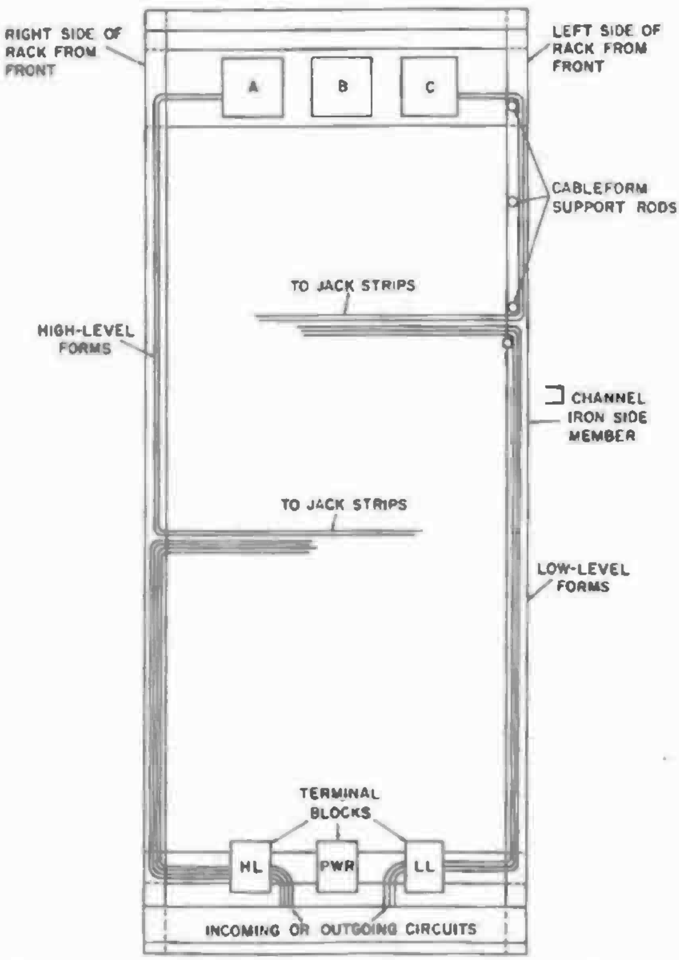


Fig. 24-20A. Rear view of a typical rack showing how the cableforms are placed.

the jack strips into two parts, making oue-half high level and the other half low level, as shown in Fig. 24-20B. This is common practice and werks quite well as the high- and low-level forms are not running parallel with each other.

It is not good practice to make one row of a double-row jack strip (Fig. 24-6A) high level and the other row low level, because of the parallel positions of the two cableforms. However, this is permissible if the level differences between the high- and low-level forms are relatively small.

24.21 What is an intermediate-level cableform and where is it placed in a rack? — Intermediate cableforms generally carry operating levels falling between the high-level and low-level forms. The intermediate form may be placed on the low-level side, provided it is laced as a separate form and is supported on the tie rods at least ½ inch away from the low-level form.

Intermediate cableforms may also be laid on the high-level side if the difference in level existing between the two forms is not too great.

24.22 Where are alternating current power circuits placed when they are run to the top of the rack?—They are brought to the top of the rack in conduit or, better yet, brought over the top of the rack in the gutter and then dropped downward in conduit to the equipment concerned.

24.23 Where are high-level lines, such as those from a power amplifier feeding a loudspeaker system, placed in a rack?—If possible, they should be encased in conduit from the output of the amplifier to the point where they are to terminate. If this is not practical, the lines are twisted and run inside the rack on the high-level side of the rack. This will separate them from the low-level circuits and provide a certain amount of shielding. Alternating-current heater lines are also run in the side members, on the high-level side.

24.24 Where are ac lines run in a cabinet-type rack?—All ac lines are run inside the rack, at the rear, on the high-level side, in a small 1½-inch gutter with ac outlets wherever required. This gutter is available in 10-foot lengths and is cut to size to fit the rack.

24.25 What is the reason for twisting wires carrying an alternating current?—The radiation of the usual ac field around a conductor carrying an alteruating current is reduced because the field from one wire is 180 degrees out of phase with the other. This causes a reduction in the strength of the magnetic field, reducing its effect on surrounding circuitry. All ac wiring in equipment or lying in a metal gutter should be tightly twisted. This will ensure a minimum of interference from clicks and stray fields.

24.26 What size wire is recommended for transmission circuits?—If the audio power is 1 watt or less, shielded pairs of No. 19 wire similar to those shown in Fig. 24-44 are recommended. Circuits carrying the output of high-power amplifiers must have low dc resistance to prevent the loss of power when transmitted over a distance. As a rule, No. 10 to 14 plastic-covered wire will suffice for this service.

The wire size may be determined by referring to the wire table in Questions 20.147 and 20.152.

24.27 Why has 600 ohms been standardized for input and output impedances?—In past years, several different values of impedance, such as 150, 250, and 500 ohms, have been more or less standard for input and output circuits for sound equipment. During the early days of fm and television sound transmission, because of the equipment available at the time, it was easier to meet the frequency requirements of the FCC relative to equalization, using a 150-ohm impedance. Also, this was the first attempt to standardize on a given impedance for both input and output circuits. However, as the VU meter was based on a 600-ohm circuit, special input transformers were required for these meters, as discussed in Question 10.34.

Equipment manufactured within the last few years use 600 ohms for both input and output impedance, except in the instance of microphone preamplifiers, which use 50 ohms (in older equipment this was 30 ohms). Some European microphones specify a load impedance of 200 ohms, others 1000 ohms. Power amplifiers employ both bridging and 600-ohm input impedances, with output impedances ranging from 4 ohms to 600 ohms, depending on the service requirements of the amplifier.

24.28 What is the formula for calculating the circular-mil area of a wire?

$$CM = \text{Diameter in mils squared}$$

$$\text{Diameter} = \sqrt{CM}$$

where,

M is the diameter of the wire in mils.

One circular mil equals 0.7854 square mils.

24.29 How may the correct size wire for a particular load be calculated?

$$CM = \frac{2D \times I \times R}{E}$$

where,

D is the loop resistance of the run in feet,
I the current in amperes,
R is a constant which has a numerical value of 10.8 for annealed copper wire and 11.06 for hard-drawn copper wire.

24.30 What are the average operating levels encountered in a large sound installation? — The approximate levels are:

Microphones	−10 dBm to 0 dBm
Capacitor with preamplifier.	
Microphones	−65 dBm to −50 dBm
Ribbon and dynamic.	
Preamplifiers	−30 dBm to 0 dBm
Mixers	−40 dBm to − 6 dBm
Booster Amplifiers	−40 dBm to − 6 dBm
Line amplifiers	−30 dBm to − 4 dBm
Record. amplifiers	+ 4 dBm to +18 dBm
Mixer console.	
Recording power amplifiers	0 dBm to +50 dBm
Disc recording.	
Monitor amplifiers	+26 dBm to +46 dBm

The foregoing data do not represent any particular installation, but show the wide range of levels encountered in a large and varied installation. If a block diagram of an installation similar to that of Fig. 24-30 is made beforehand, showing the expected operating levels, it will aid in the laying out of racks, cable runs, and cableforms.

24.31 What is a ground loop and how is it caused?—The formation of a ground loop can best be explained by referring to Fig. 24-31. Here are shown two pieces of equipment, A and B, connected by means of a shielded pair. For the sake of illustration, it will be assumed that the pair is shielded, but the shield is bare, so it can make contact with the ground. It is actually grounded at the points X and Y.

It is known that many stray electrical currents are flowing through the ground, as the result of grounding electrical power systems, trolley cars, and other electrical equipment. (See Question 25.124.) Consequently, if both ends of the shield are grounded at points X and Y, electrical currents in the vicinity of the ground connections will take the path of least resistance to the other end, which is over the shield rather than through the highest resistance of the ground. When an electrical current flows through a conductor, magnetic lines of force are caused to expand and contract around the conductor. This is especially true if the current is an alternating one. If a second conductor is placed in the field of the first, the lines of force cut the second conductor and induce a current in the second conductor.

This is the condition that prevails when ground currents are permitted to flow over a shield, or when a shield is grounded in several places. The induced current is generally the cause of noise or high-frequency oscillations,

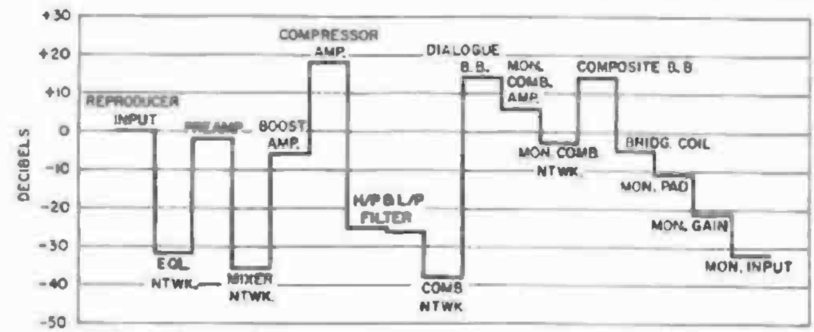


Fig. 24-30. Circuit-level diagram for dubbing console in Fig. 9-47A. The levels are predicted on a zero dBm signal level from the reproducers.

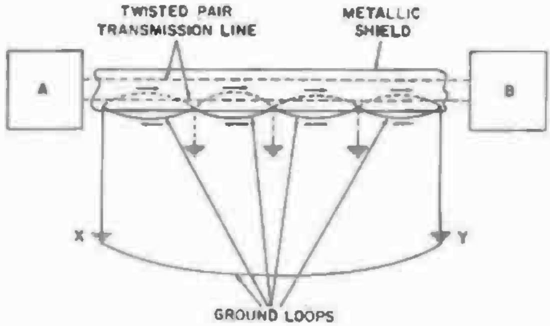


Fig. 24-31. Ground loops caused by multiple grounds on bare shielded pair.

the latter being particularly true of systems having high values of amplification.

In large installations where heavy radio-frequency current may be flowing, it is the practice to use bare shielding and bond the shields together every few inches, to form individual shields into a solid shield. This method is satisfactory only if the bond is securely made and tied to a ground bus every few inches.

Large sound installations use shielded wire with a cloth braid over the outside of the shield to prevent it from coming into contact with ground. The shield is grounded at one end only, as described in Question 24.33.

24.32 ***What is a common-ground system?***—A single or common-ground wire connecting several different pieces of equipment. Single-wire ground systems are not adaptable to large sound installations because of the many different operating levels and the formation of ground loops. (See Question 24.31.) Common-ground systems are also a cause of high noise levels, particularly when low-level, high-gain equipment is connected at the end of the ground connection. Fig. 24-32 shows four units of a sound installation consisting of two low-level voltage amplifiers and two power amplifiers.

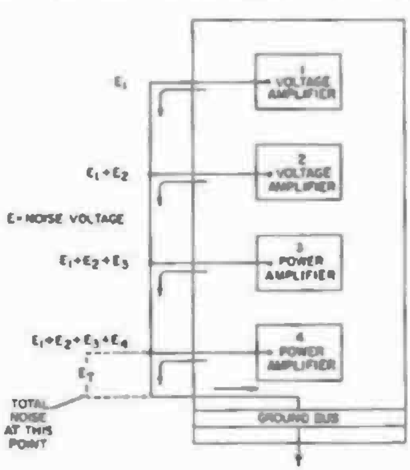


Fig. 24-32. Common or single-ground wire system.

Amplifiers for low-level operation, such as microphone and photocell preamplifiers, are designed for a very low internal noise level and require careful installation and a low-resistance ground connection. On the other hand, a power amplifier operates at a considerably higher level and can tolerate a somewhat higher internal noise level. However, a power amplifier must also be well grounded for stability and quietness of operation. If the four units shown are grounded by means of a common ground wire, the internal noise of units 3 and 4 would be induced in series with the ground return of voltage amplifiers 1 and 2.

If the situation is analyzed, any noise generated in the ground circuit of unit 1 is added to the ground noise of unit 2. Combined noise of units 1 and 2 is then added to the ground noise of unit 3. Likewise, the three previous units add their noise to that of unit 4. The noise on the lower portion of the ground wire between unit 4 and the point where it connects to the ground bus is E_t, the noise of all four units combined.

Although the actual ohmic resistance of the ground wire may be on the order of a fraction of an ohm, a small voltage drop (because of the noise current) will exist on the ground wire. Therefore, high-gain, low-level units will be affected by the amount of noise current in the ground wire.

The common-ground wire system is not recommended for large sound installations for the reasons explained previously. However, in a small installation such as a public address system where the number of pieces of equipment is small, it may be used to a good advantage and is easy to install.

24.33 What is a transmission-ground system?—A method of grounding used in large commercial sound installations. The ground system is split into two parts, one section being called the transmission ground and the second the chassis or frame ground.

The transmission ground connects to the grounds of filters, equalizers, amplifiers, and the negative side of power supplies (this is the portion of the ground system carrying speech currents) and is run directly to the main ground bus of a given equipment rack. The frame or chassis ground connects

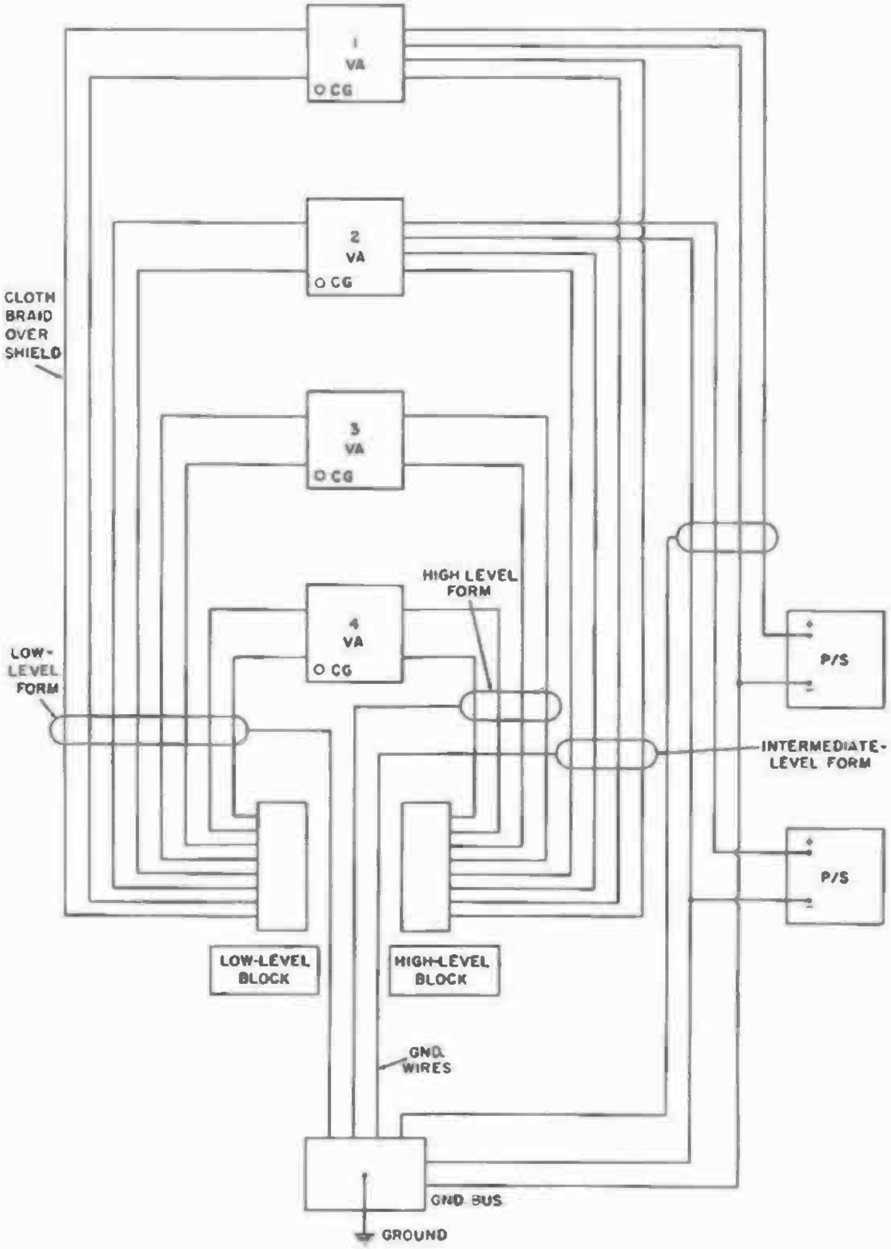


Fig. 24-33A. Transmission ground system applied to a rack containing four units, with separate power supplies. (Front view of rack.)

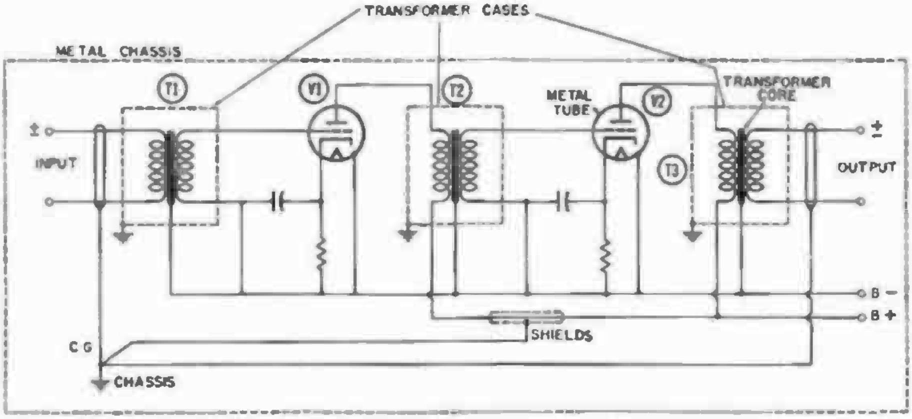


Fig. 24-33B. Transmission ground system applied to an amplifier. The internal wiring shields are connected to the chassis ground at one point only. The cores of the transformers are connected to the B− circuit. Transformer cases are connected to the chassis. The B− circuit is brought to a separate terminal, and is completely clear of any grounds in the amplifier.

to shields, transformer cases, conduit, and racks. Such a grounding system, if carried out to the letter, eliminates the formation of ground loops and reduces the noise level of the system. A transmission-ground system will also permit the ground of any given rack to be lifted without disturbing other parts of the installation.

A typical transmission-ground system is shown in Fig. 24-33A. It consists of several pieces of equipment supplied by external power supplies (a common practice in the motion picture industry). Each piece of equipment has individual input and output terminals, power supply connections, and a terminal indicated CG (case ground). This latter terminal is fastened directly to the chassis of the equipment concerned.

In a transmission-ground system, the B-minus terminal of a particular piece of equipment is not connected to the chassis as would be the case of a common ground system such as that described in Question 24.32. The B-minus terminal is left floating as shown in Fig. 24-33B. This latter diagram is a two-stage amplifier, transformer-coupled at the input and output, as well as interstage. It will be observed the transformer cases are grounded to the

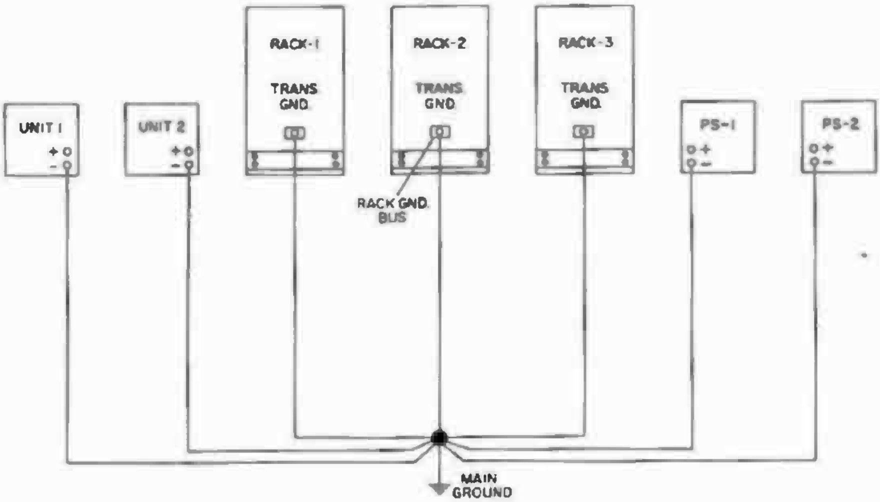


Fig. 24-33C. Transmission ground system. The racks and cases of the power supplies receive their ground from the normal conduit ground.

chassis but the core of the transformer is connected to the B-minus circuit. Also, the metal envelopes of the tubes are connected to this circuit. Any internal shields surrounding the wiring are connected to the chassis at the CG terminal. This method of wiring will reduce the internal noise of an amplifier to a minimum and also prevent noise from being introduced into the transmission circuits via the transformer cores and power supply. In a true transmission-ground system, there is no continuity between B-minus and chassis ground. Again referring to Fig. 24-33A, each unit requiring a separate power supply has a shielded pair run between the power supply and the unit in the rack. It will be particularly noted that an individual insulated ground wire is run from the B-minus terminal of the power supply directly to the ground bus at the bottom of the rack.

In some installations a ground wire is also run from each individual CG terminal on the chassis of each unit to the gronnd bus. This connection is generally dispensed with when the equipment is mounted in a rack, as the rack grounds the chassis. However, if the unit is not mounted where it can be properly grounded, a ground wire is run from this terminal to a ground.

Each circuit feeding an input or output is run in shielded pairs with a cloth braid over the shield as shown in Fig. 24-33A. Although the drawing shows the pairs as individual lines, they are actually formed into a single cableform and securely laced together. The lower ends of the shields are connected together and run directly to the ground bus at the bottom of the rack. This is the *only* point at which the shields are grounded. Under no circumstances must a second ground be connected to the shields.

If a shielded pair has an internal ground wire, as shown in Fig. 24-44A, the ground wires are brought to a common terminal on the connecting block at the bottom of the rack and from there to the ground bus. Ground wires running from a cableform must be insulated and not smaller than number 16 to 18 gauge.

Fig. 24-33C illustrates how the transmission-ground bus of several racks is connected to the main ground of the installation. The wire between main ground and rack must not be less than number 6 gauge and, if practical, larger. In installations where a number of racks are placed side by side, a copper bus bar 1¾ inches by ¼ inch thick may be run across the fronts of the racks just above the foot plate. This bus bar is securely bolted to each rack to bring all the racks to as near the same ground potential as possible.

24.34 If a jack strip uses a metal mounting strip to ground the jack frames, is it necessary to ground the jack frames in addition?—If the jack frames are provided a low-resistance path to ground through the rack members, no additional ground is required. However, in some types of jack strips, the jack frames are insulated from ground although the mounting strip is metal. In this instance, the frames of the jacks are strapped together and run to ground.

It should be pointed out that in the manufacture of jack strips using a metal mounting strip, the strip is often anodized. The manufacturer depends on the insulation offered by the anodizing process to provide the insulation between the jack sleeve and the mounting strip. In some instances, this is on the order of 1 megohm; however, if the anodizing is scratched during the assembly of the jacks in the strip, a direct contact between the strip and the jack frame results. In this latter case, a ground loop could be created between the frames of the jacks (if they are grounded separately) and the frame of the rack.

In racks housing low-level equipment, it is recommended that jack strips using insulated mounting strips be employed, and the jack frames be tied together and grounded to the rack ground bus.

24.35 What is the procedure for calculating the impedance of a bridging bus and its terminating resistor?—Fig. 24-35A shows a typical bridging bus as might be found in a recording studio. It will be noted the equipment bridged across the bus acts as the load impedance for the line amplifier driving the bridging bus. The first step in the design of the bus circuit is to determine what and how many pieces of equipment will be permanently connected (normalled) across the bus.

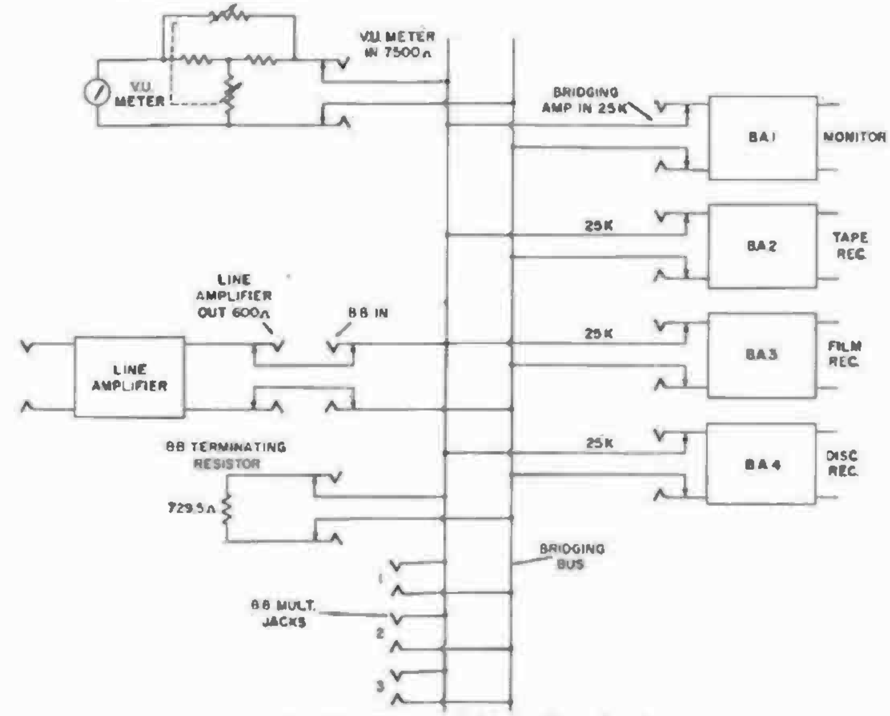


Fig. 24-35A. Typical bridging-bus circuit.

In the drawing shown, the bus has normalled across it four bridging amplifiers each having a 25,000-ohm bridging impedance and a 7500-ohm VU meter. In parallel with the equipment is a bridging-bus terminating resistor and three sets of bridging-bus multiple jacks. It will be assumed that, for the sake of illustration, any additional equipment bridged across the bus will have a bridging impedance of not less than 25,000 ohms. The output impedance of the line amplifier feeding the bus is 600 ohms. If the total resistance presented by the several pieces of equipment bridging the bus is computed (exclusive of the bridging-bus terminating resistor), it will be the equivalent of 3416 ohms. If only this value of resistance were used for terminating the output of the line amplifier, the amplifier would not see its proper load impedance of 600 ohms. For an amplifier to see 600 ohms it is necessary to connect a resistance, in parallel with the bridging bus load, of such value that the correct load impedance will be presented to the amplifier output. Carrying this a step further, the four 25,000-ohm bridging inputs present a load impedance of 6250 ohms. In parallel with this is the 7500 ohms of the VU meter. Using Ohm's law for parallel circuits:

$$BB = \frac{6250 \times 7500}{6250 + 7500} = \frac{4.68 \times 10^7}{1.37 \times 10^4}$$

$$= \frac{4.68 \times 10^3}{1.37} = 3416 \text{ ohms.}$$

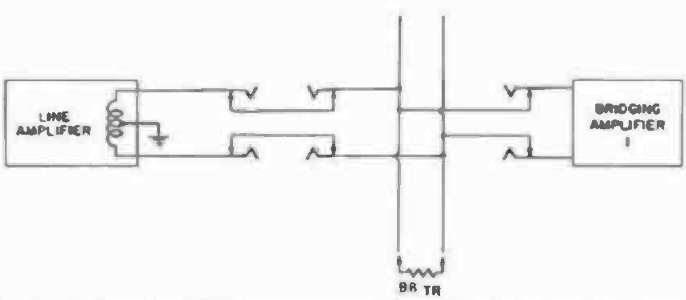


Fig. 24-35B. Bridging bus fed from an amplifier with a balanced to ground output.

Now, the question arises, what value of resistance connected in parallel with 3416 ohms will provide a resistance of 600 ohms? Again using Ohm's law for parallel resistors:

$$BB_{tr} = \frac{3416 \times 600}{3416 - 600} = 728.84 \text{ ohms.}$$

Therefore, if a resistance of 728.84 ohms is connected in parallel with the bus and the foregoing load impedances, the amplifier will see a load impedance of 600 ohms. As a rule, it is safe to load the output of a well-designed amplifier having an output impedance of 600 ohms to about 540 ohms, or 10 percent below 600 ohms. This means that for the foregoing bridging bus several more 25,000-ohm inputs could be paralleled across the bus without affecting the operating characteristics of the line amplifier. If the bridging impedances are greater than 25,000 ohms, more equipment can be fed from the bus.

As a rule, a bridging input impedance greater than 50,000 ohms is not used. Standard bridging impedances are: 7500, 25,000, and 30,000 ohms. Although the bridging bus in this example is ungrounded, one side may be grounded, if required. Generally, large installations operate with an ungrounded bridging bus to provide a greater degree of flexibility. If the amplifier feeding the bridging bus is balanced to ground, (output center tapped) Fig. 24-35B, the ground is applied by the line amplifier.

Two reference levels are in use for bridging-bus operation: 1 milliwatt and 6 milliwatts. The first is the present-day standard, while the second may still be found in older installations. Because 6 milliwattts corresponds almost to a plus 8 dBm, some installations operate with a plus 8-dBm bus which is, in reality, the old 6-milliwatt reference level. Usually broadcast installations use a bridging-bus level ranging from plus 10 to 18 dBm, while in the motion picture industry the bus level ranges from plus 8 to 14 dBm.

24.36 Describe a relay rack and cabinet-type equipment rack. — Relay racks are used to support various types of electronic equipment. The telephone industry, early in its development stage, recognized the need for standardization in mounting equipment to permit its manufacture by anyone and yet be able to mount it in a rack without the necessity for drilling holes for each individual piece of equipment, and, also, to permit the equipment to be placed in a rack with a minimum of effort. This thought resulted in the design of a metal rack called a *relay rack*, shown in Fig. 24-36A. Relay racks were originally designed for holding telephone relays—hence the name.

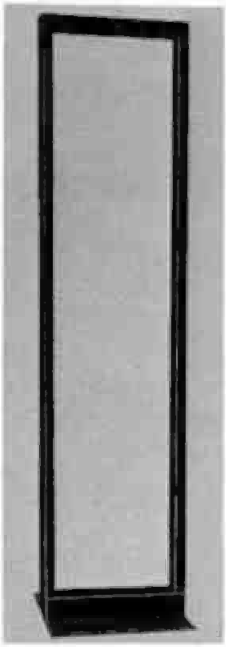


Fig. 24-36A. A 7-ft. relay rack. The side members are "U" shaped to permit cable to be laid inside. The base is of heavy angle iron, which is used to secure the base to the floor.

The rack consists of two side members constructed of channel iron held at the top by steel straps, and at the bottom by a heavy angle-iron foot piece. The side members have a series of holes drilled in their face at such locations that panels cut in multiples of 1¾ inches high may he mounted anyplace in the rack without drilling additional holes. Equipment panels are cut in multiples of 1¾ inches and are slotted at both ends for mounting screws.

It is the practice in modern sound installations to employ cabinet-type racks, as shown in Fig. 24-36B. Here the equipment is mounted in a shielded environment and protected; also, it is prevented from inducing fields into surrounding equipment. This is particularly important for magnetic recording and reproducing equipment.

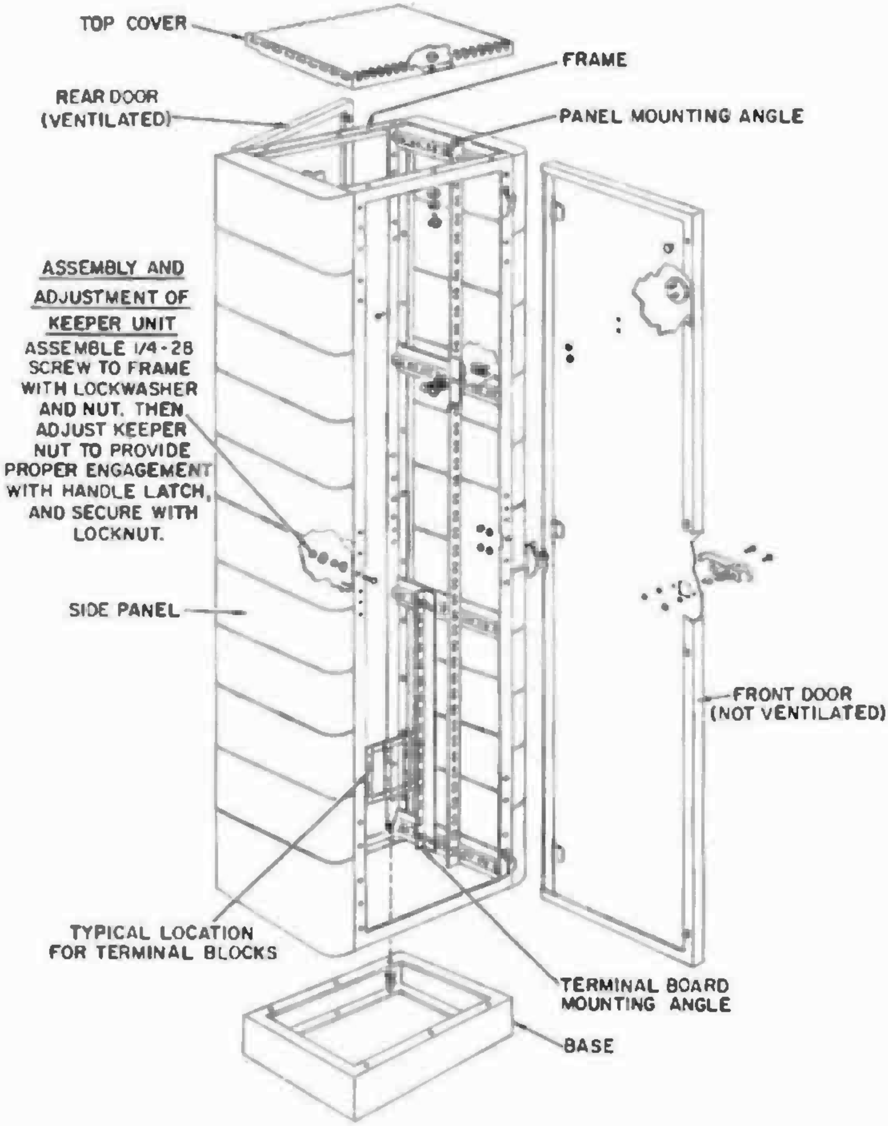


Fig. 24-36B. RCA BR-84 cabinet-type equipment rack. (Courtesy, Radio Corporation of America.)

Cabinet-type racks employ the same multiple drilling dimensions as for the open-channel type of rack, but they provide many convenient mounting areas for terminal blocks and power-block connections. The multicell gutter is placed over the cabinet (see Fig. 18-330A) and pipe nipples are run through the top cover. The wiring is then brought to the terminal blocks mounted at the rear of the rack, near the top and bottom. It will be noted that the angle bracket may be placed to fit any standard piece of equipment. The rounded side members may be replaced with flat plates, and the cabinet used with or without doors.

24.37 ***What are the dimensions for cutting and notching panels for relay-rack mounting?***—Fig. 24-37 gives the cutting and notching dimensions for relay-rack panels. It will be noted the panels are cut in multiples of 1¾ inches high and 19 inches long. The height is cut $\frac{1}{64}$ inch undersize to each side of the centerline of the panel. This undersizing allows a small clearance between panels when they are mounted in a rack. The smallest panel is one multiple high, and the largest is seven multiples

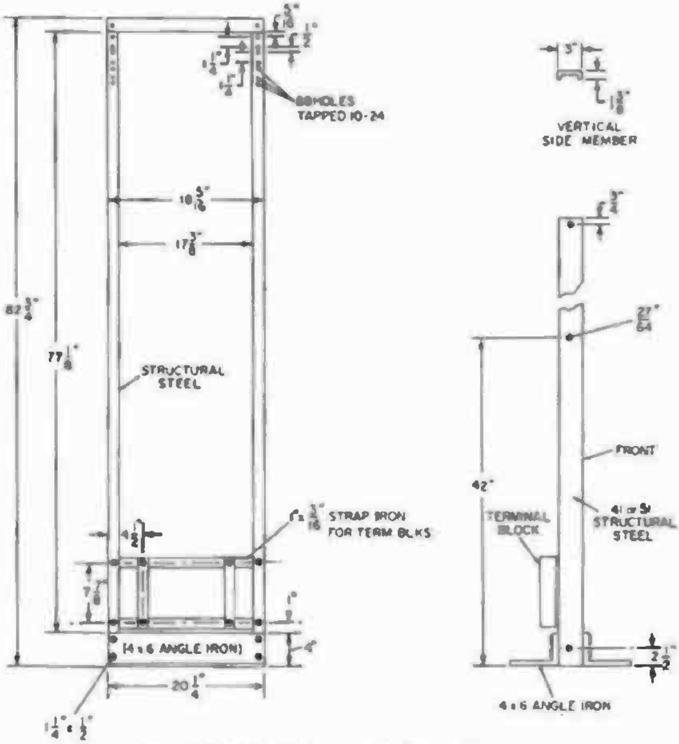


Fig. 24-36C. Relay-rack dimensions.

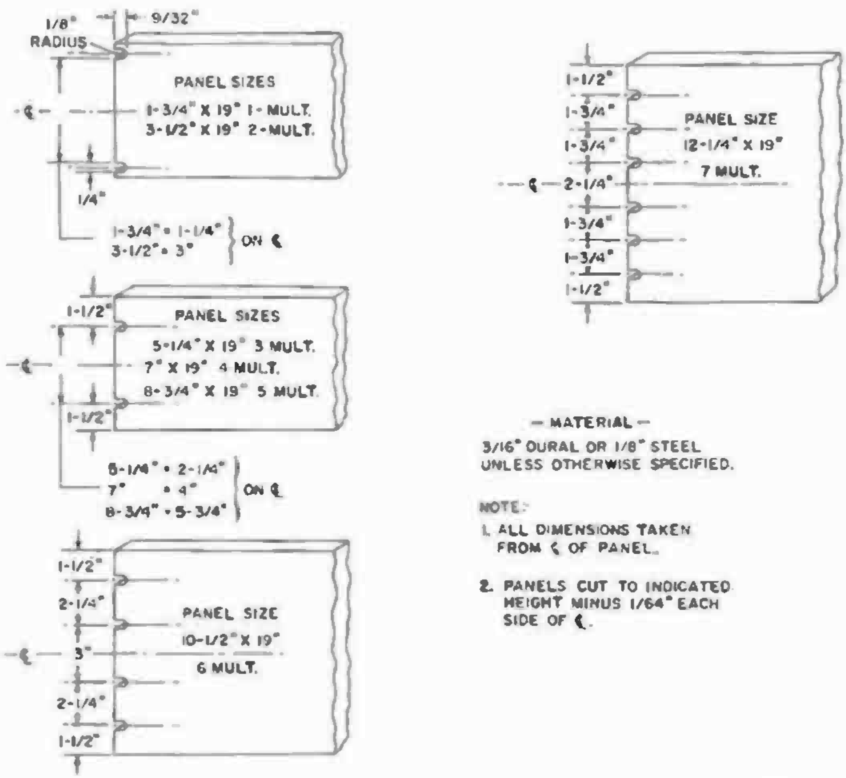


Fig. 24-37. Dimensions for cutting and notching panels for relay mounting. The panels are cut in multiples of 1¾ inches high and 19 inches long.

high. Panels may be constructed of either 3⁄16-inch Duralumin or ⅛-inch steel. The notches are dimensioned for 10-24 roundhead machine screws.

24.38 What is the procedure for connecting to terminal blocks? — The terminal count is started at the top right-hand terminal, as shown in Fig. 24-38, and is terminal number 1. The bottom right-hand terminal is number 4. The terminal to the left of number 1 in the top row is number 5 and the terminal to the left of number 4 in the bottom row is number 8. This method of counting continues to the left as shown. The odd-numbered terminals are always the ground or low-potential side of the circuit. As a rule, two-conductor shielded pairs consist of a red/green and red conductor. The red/green conductor is connected to the odd-numbered terminals. If the pair consists of a red and white pair, the red is connected to the high side of the circuit or the even-numbered terminals. If a pair consists of a red and black pair, the black is connected to the odd-numbered terminals.

24.39 What is a punching or running sheet?—A sheet showing the terminal numbers and connections of a terminal block. A typical running sheet corresponding to the jack assignment sheet of Fig. 24-40 is shown in Fig. 24-39. Incoming and outgoing lines from the rack are shown at the right. The connections for the internal cabling of the rack appear at the left.

24.40 What is a jack assignment sheet?—A jack assignment sheet corresponding to the running sheet of Fig. 24-39 appears in Fig. 24-40. Each square on the sheet represents two jacks of a double-jack circuit.

The upper jack strip has been split into two parts. Circuits 1 to 6 are incoming microphone lines from a studio, and normal to the inputs of six microphone preamplifiers, circuits 1 to 6 in jack row 2. The outputs of the preamplifiers appear on circuits 7 to 12 of jack strip 1. These circuits normal to circuits 7 to 12 of jack strip 2.

Row 3 contains jacks which appear in various parts of a recording circuit. Row 4 contains a strap jack, 600-ohm resistive termination, input to a VU meter, and two bridging-bus extension circuits. Normal connections between circuits are indicated by the short vertical and horizontal lines between jacks.

24.41 What is a barrier strip?—A terminal strip as shown in Fig. 24-41 for terminating incoming and outgoing power circuits at the bottom of a relay rack or inside equipment cases. Each circuit is separated from an adjacent circuit by an insulated barrier to prevent flashovers and short circuits.

24.42 What are pin counts? — In certain phases of the sound industry the larger suppliers of equipment, particularly of recording equipment, have standardized on a given method of connecting audio and power circuits to cable plugs and their receptacles. The pin counts used by RCA and Westrex for broadcast and recording equipment are shown in Figs. 24-42A, B, and C. The conductors are connected to the plug pins by the number, regardless of the type of plug or receptacle used.

24.43 What is the proper method of lacing a cableform?—The pairs are laid in straight lines, in positions that will permit them to be broken-out at the proper intervals for connection to equipment. The form is tightly laced together using No. 6 or No. 9 waxed lacing cord. The stitch used to hold the pairs is called a *lockstitch* and is shown in Fig. 24-43.

The size of cord and the distance between stitches is dependent on the number of pairs in a form and the number and type of bends required.

24.44 What does "serving a pair" mean?—Serving a pair means securing the end, or ends, of a conductor so the shield or covering will not unravel or short circuit to another pair or to ground. Serving may be accomplished in several different ways.

In Fig. 24-44A, the serving on the end of the pair consists of a wrap made with No. 6 waxed lacing cord. In Fig.

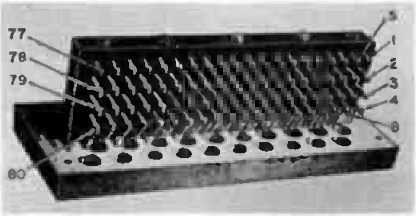


Fig. 24-38. An 80-terminal block, used at the bottom of a relay rack for terminating cableforms and incoming and outgoing lines.

24-44B, the pair is served with vinyl plastic tubing. The tubing is soaked in lacquer thinner (without butyl acetate) for a few minutes, and then spread with a pair of long-nosed pliers. The tubing is then slipped over the end of the wire shield or covering. After the lacquer thinner has evaporated, the plastic tubing will shrink tightly around the wire.

In Fig. 24-44C, the pair has been served with a Thomas and Betts fitting. This metal fitting consists of two parts: an inner part which is tapered on the outside, and an outer piece which fits

Rack		*Gutter*
Microphone Tnk. #1 R1, J1,	1 2	From Microphone Connector #1 in Studio
Microphone Tnk. #2 R1, J2,	3 4	From Microphone Connector #2 in Studio
Microphone Tnk. #3 R1, J3,	5 6	From Microphone Connector #3 in Studio
Microphone Tnk. #4 R1, J4,	7 8	From Microphone Connector #4 in Studio
Microphone Tnk. #5 R1, J5,	9 10	From Microphone Connector #5 in Studio
Microphone Tnk. #6 R1, J6,	11 12	From Microphone Connector #6 in Studio
Mixer Tnk. #1 R2, J7,	13 14	To Mixer Input #1 in Console
Mixer Tnk. #2 R2, J8,	15 16	To Mixer Input #2 in Console
Mixer Tnk. #3 R2, J9,	17 18	To Mixer Input #3 in Console
Mixer Tnk. #4 R2, J10,	19 20	To Mixer Input #4 in Console
Mixer Tnk. #5 R2, J11,	21 22	To Mixer Input #5 in Console
Mixer Tnk. #6 R2, J12,	23 24	To Mixer Input #6 in Console
Mixer Out R3, J1,	25 26	From Mixer Output, Console
Master Gain Cont. In R3, J4,	27 28	To Master Gain Cont. in Console
Master Gain Cont. Out R3, J5,	29 30	From Master Gain Cont. in Console
	31 32	
	33 34	
	35 36	
LL Test Tnk. #1 R, J,	37 38	To Rack #2
LL Test Tnk. #2 R, J,	39 40	To Rack #2

RUNNING SHEET
LL 40 BLOCK
RACK #1
SHEET #1 of 2
DATE 3 Dec. '56
H. M. T.

NOTE: Terminal #1 is the SHORT PIN and is COMMON.

Fig. 24-39. A typical running sheet used with a 40-terminal block similar to that in Fig. 24-38.

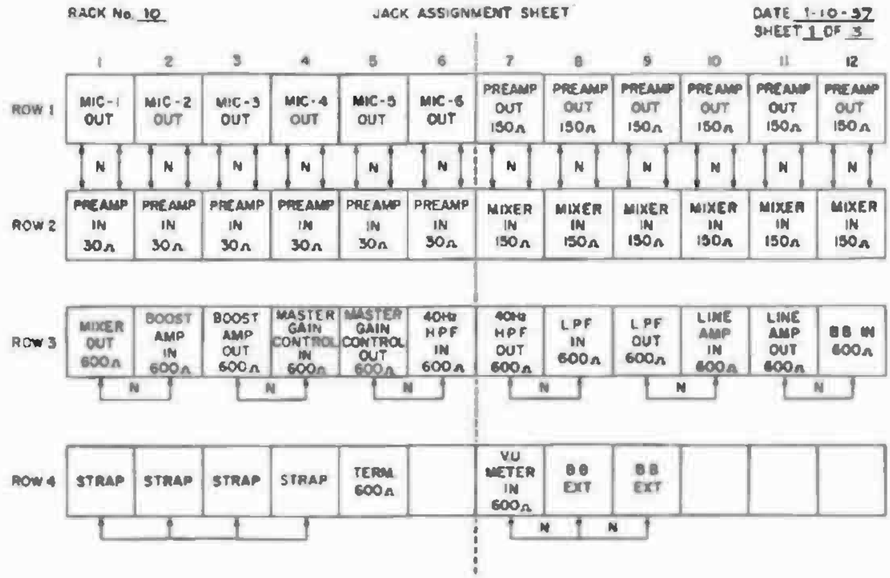


Fig. 24-40. A typical jack-assignment sheet.

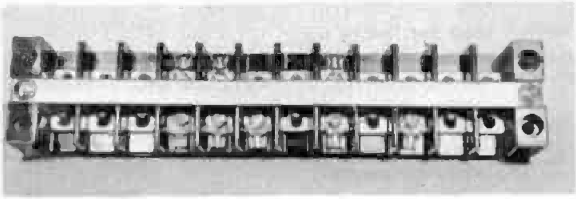


Fig. 24-41. Barrier terminal strip for power circuits.

the large end of the inner component. The metal shield of the wire is slipped over the outside of the inner part and the outer part clamps over the shield by means of a special tool. This results in a clean serve which may be used to ground the shield.

Fig. 24-44D shows a shielded pair unserved and Fig. 24-44E depicts a single shielded conductor unserved. It will be noted in these photographs, except for Fig. 24-44E, a tinned ground wire is shown projecting from the end of the shield. This wire is continuous throughout the entire length of the pair and is used to ground the shield of the pair. By the use of a continuous ground wire, as shown, the ground wire is in contact throughout the entire length of shield, reducing resistance between shield and actual ground. This type of wire is used almost exclusively in large installations.

24.45 What type of actual ground is recommended for a large sound installation?—Two types of grounds are recommended. The first is a water-pipe ground, the connection being made at a water main where it enters a building. The second method consists of burying several large copper plates in moist earth which has been well mixed with

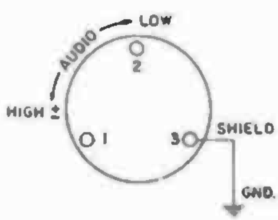


Fig. 24-42A. Westrex pin count.

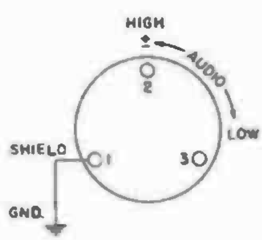


Fig. 24-42B. RCA pin count.

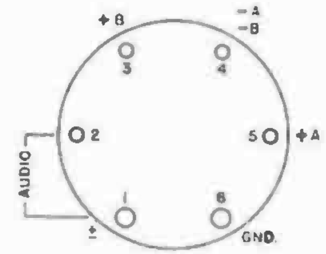


Fig. 24-42C. RCA pin count for 6-pin plugs.

rock salt. If this is not practical, several heavy wires may be buried in a trench prepared in a similar manner. In any event, the ground must be of low resistance, which can be obtained only in moist earth.

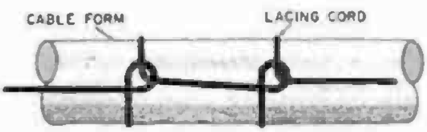


Fig. 24-43. Lockstitch used for lacing a cableform.

24.46 How can ac ground loops be avoided when several pieces of equipment are fed from the same ac source? —This subject is discussed at length in Question 23.217, which outlines the procedure that should be followed when several pieces of equipment (particularly those of high gain) are mounted in one rack.

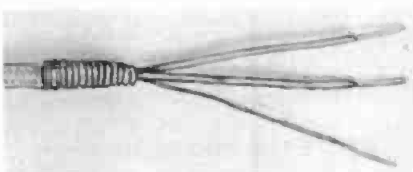


Fig. 24-44A. Two-conductor shielded pair, with ground wire and cloth braid overall served with No. 6 lacing cord.

24.47 What is a grasshopper fuse? —An indicator alarm fuse which will operate a bell or light to show when it has blown. A typical fuse of this type is shown in Fig. 24-47. The fuse wire is connected to a spring that is released when the fuse blows. The spring makes contact with another contact on the mounting fixture, causing a bell to ring.

24.48 What is a main frame?—A large framework supporting a group of

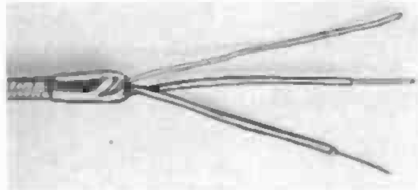


Fig. 24-44B. Two-conductor shielded pair, with ground wire, cloth braid overall, served with single tubing.

terminal blocks to which incoming lines are terminated. Lines are then run from the main frame to different parts of the installation. Main frames are used only in very large plants where several hundred circuits enter a building and are distributed over a large area.

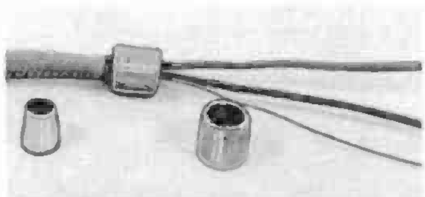


Fig. 24-44C. Two-conductor shielded pair with ground wire, cloth braid overall, served with Thomas and Betts fitting.

24.49 Describe the different types of metal gutter used in sound installations.—It is the general practice in sound installations to provide a metal gutter for running the ac power, control circuits, dc power supply, audio transmission lines to the equipment and between the various racks and power

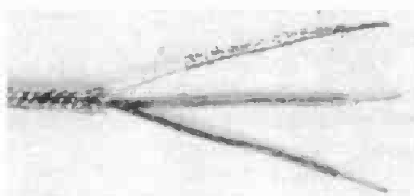


Fig. 24-44D. Two-conductor pair with ground wire and shield unserved.

room. The gutter not only provides protection for the lines, but also shields them and prevents inducing electromagnetic energy into other equipment.

Fig. 24-44E. Single conductor with copper tinned shield unserved.

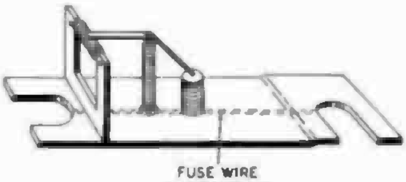


Fig. 24-47. Alarm or grasshopper fuse.

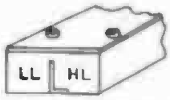


Fig. 24-49A. Metal gutter used for audio-circuit wiring.

A gutter also provides a convenient means of making changes in existing equipment, as well as future expansion.

In the power room the ac lines are run in 4 × 4-inch gutter with a removable cover, mounted on the walls. In small installations the high- and low-level audio lines may be run in a double gutter (Fig. 24-49A). For larger plants a three-cell gutter (Fig. 24-49B) may be used for the power and control circuits.

For large installations which involve the use of many racks, a six-cell gutter is generally employed. This gutter runs the full length of the installation, from the mixer console to projection room, machine room, transfer channel, and the power room. It is mounted over the equipment cabinets and runs on the wall to other parts of the plant. A typical installation is shown in Fig. 18-330A, and in Figs. 24-49C and D. The cables or wires are broken out at various positions in the gutter and mechanically connected to cabinets by 2-inch pipe nipples. The wiring is then brought to terminal blocks which are mounted at the top or bottom at the rear of the racks.

All ac and dc wiring is firmly twisted before being laid in the gutter. The

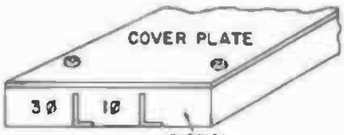


Fig. 24-49B. Metal gutter used for power wiring and control circuits.

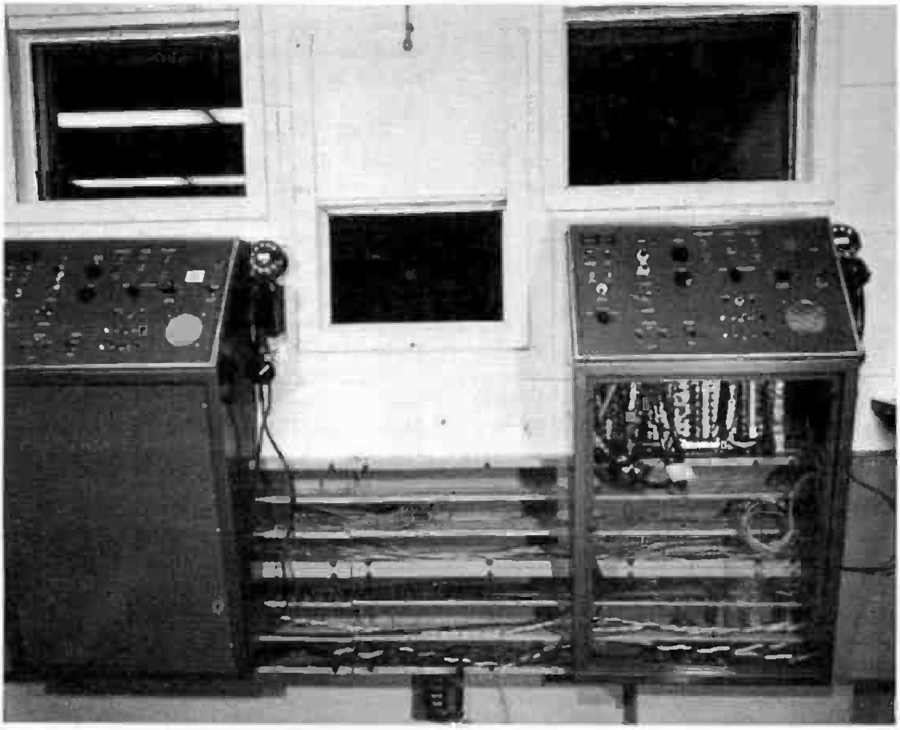


Fig. 24-49C. Interior view of a six-cell metal gutter interconnecting two control positions in a projection room.

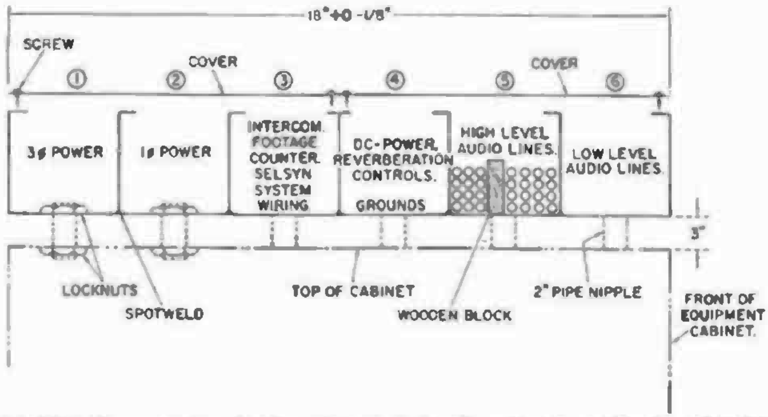


Fig. 24-49D. Cross-sectional view of typical 6-cell gutter run, showing the division of the power, control, and audio wiring for a sound recording installation.

low-level lines are separated into two groups, with the higher- and lower-level lines separated by wooden blocks every few feet. High-level lines are treated similarly. To facilitate its removal at some future time, none of the wiring is laced. Ground wires may be laid in either the lower- or higher-level gutter if necessary; however, they should be consigned to the dc channel whenever possible. Ground wires *must be insulated*. The gutter is firmly grounded to the rear of each cabinet with a flexible 1-inch copper strap. The gutter is grounded to the conduit system. Each cabinet is bonded to the adjoining rack with a flexible copper strap. The gutter may be suspended from the ceiling, where practical, by 1-inch hangers placed approximately every 10 feet.

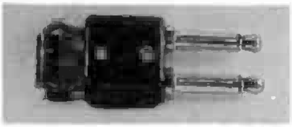


Fig. 24-50. Terminating resistor mounted on a Western Electric 289A double-circuit plug.

24.50 What is a termination plug? —A standard cord plug fitted with a terminating resistor, as shown in Fig. 24-50. An installation where a considerable amount of patching and frequent transmission measurements are made generally requires a number of such plugs with different values of terminating resistors. These plugs are patched into strap jacks for terminating a line, amplifier, or other device. For heavy power-output loading, a wirewound resistor of the proper wattage is used.

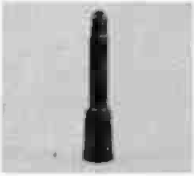


Fig. 24-51. Dummy plug used for lifting a normal circuit in a jack. The plug is made of molded Bakelite.

24.51 What is a dummy plug?—An insulated plug, as shown in Fig. 24-51, used for the purpose of lifting a normal circuit in a jack or patching out a circuit.

24.52 What is a forming board?—A board with a number of pegs or nails driven into it for the purpose of laying out a cableform. The pegs are placed at the points where the conductors break out of the form or run to components. After the wires are laid in, the form is laced or bound together with plastic fasteners.

Cableforms for relay racks are laid out on a board the full size of the rack. Holes are drilled at the points where the pairs break out to the jack rows. The pairs are pushed through the holes to hold them in place as the cable is laced. Pegs are placed at the points where the pairs break out for the various pieces of equipment.

Forming and lacing the cable before mounting it in the rack facilitates the identification of the pairs and the grounding of the shields.

24.53 What are the drilling dimensions for single- and double-jack rows?—Fig. 24-53A shows the dimensions for

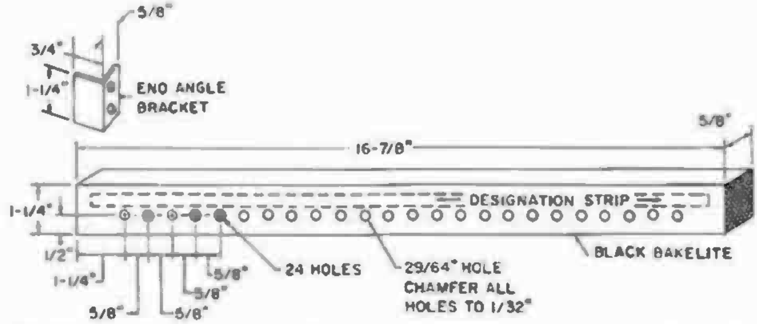


Fig. 24-53A. Drilling dimensions for a single jack row mounting strip and for the mounting bracket.

drilling the mounting strips for a single-jack row. The mounting strip may be double X (XX) black bakelite, or metal. Small angle pieces, slotted for single multiple mountings, are used to firmly support the mounting strip at the ends.

The mounting dimensions given are standard and will accommodate either single- or double-circuit plugs described in Questions 24.3 and 24.4.

Dimensions for double-row jack strips are given in Fig. 24-53B. As a rule, modern jack strips are constructed one multiple high, and slotted for single multiple mounting.

24.54 What are the dimensions for jack spacers?—At times it may be necessary to use an odd number of jack rows in a rack, which causes the mounting pattern to end in a fraction of a multiple. Such spaces are filled with a hardwood spacer as shown in Fig. 24-54. It is of the same length as a jack mounting strip and is supported at the ends by angle brackets, using a Western Electric No. 25 jack fastener.

24.55 What is a circuit laboratory?—A test section in a large sound installation. This name originated in the telephone industry. The equipment usually found in such a laboratory consists of an oscillator, gain set, distortion-measuring equipment, amplifiers, coils, terminations, and other transmission-measuring equipment.

Several high- and low-level trunks are originated in the laboratory to run to several patch bays throughout the installation to facilitate the testing of equipment at a remote location.

A typical circuit laboratory in the Sound Department of Paramount Productions, Inc., Hollywood, California, appears in Fig. 24-55.

24.56 What are the advantages of using a buzzer test set rather than an ohmmeter when testing for direct-current continuity?—It is the general practice among sound and telephone main-

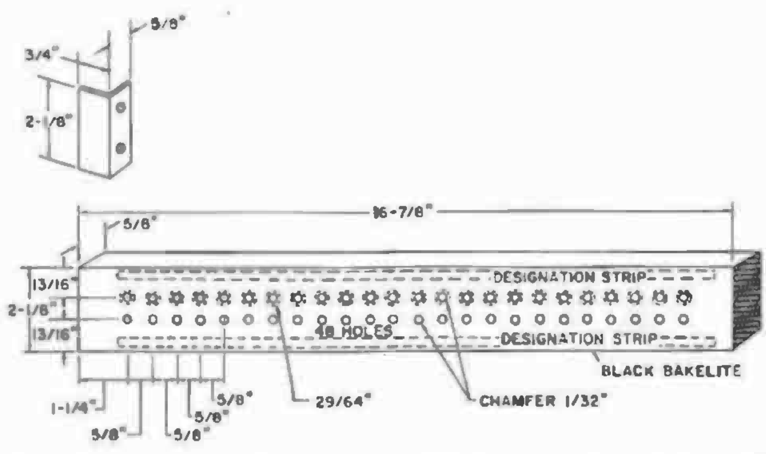


Fig. 24-53B. Drilling dimensions for a double jack row mounting strip and for the mounting bracket.

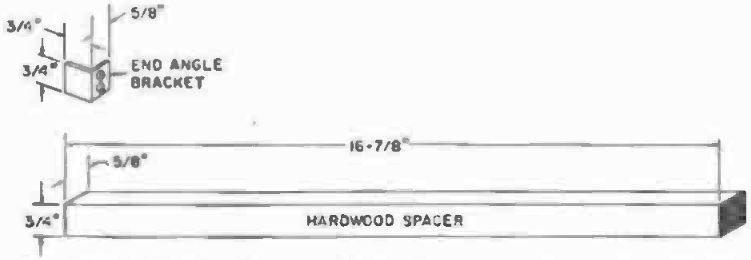


Fig. 24-54. Hardwood spacer for jack strips, and mounting bracket.

tenance men to use a buzzer test set rather than the conventional volt-ohmmeter when testing the continuity of lines, grounds, and other circuits of low resistance. A buzzer has several advantages over an ohmmeter because it indicates continuity by hearing rather than by observing a meter scale. Buzzers require considerably more current to operate than does an ohmmeter; thus,

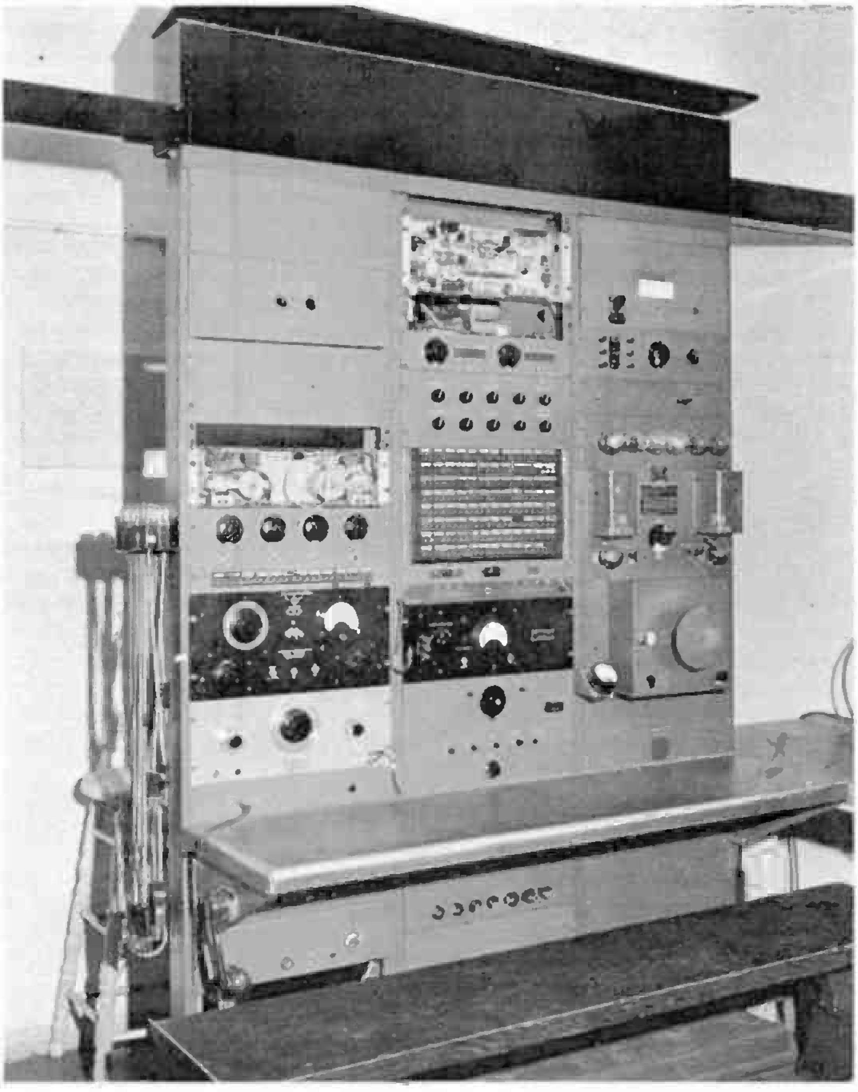


Fig. 24-55. One of two circuit laboratories in the Sound Department of Paramount Productions, Inc., Hollywood, Calif.

if a circuit such as a line or ground wire which should be of low resistance indicates continuity by a weak buzz, it is an indication that the circuit has some resistance. Even if the resistance of the circuit is only 1 ohm, the tone of the buzzer will show a noticeable change in its intensity and tone quality.

Buzzers can be of great help when attempting to locate a line in a cableform. The buzzer is connected to a line and opposite ends of the lines shorted until buzzer is caused to operate. Point-to-point testing is simplified because the eyes do not have to be shifted every time a test is made, to determine whether the circuit is continuous.

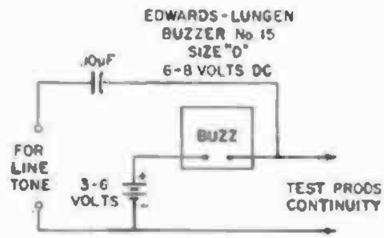


Fig. 24-56. Buzzer test set. For continuity tests the test prods are used. For applying a constant tone to a line, the line is connected to the tone terminals. Short test prods to actuate buzzer.

The only time an ohmmeter is actually required is when the circuit resistance is used to identify a component such as a resistor or coil. The circuit of a typical buzzer test set is shown in Fig. 24-56. Buzzer test sets may also be used as a source of steady tone for testing long lines in which the dc resistance of the line would prevent the bnzzer from operating. The terminals of the buzzer test set are connected to the line at one end and the test probes shorted to cause the buzzer to operate.

A headphone is used at the other end to locate the pair to which the buzzer is connected. The proper line may be identified by a high-pitched, steady tone.

24.57 What are relay-adjusting tools?—To properly adjust relay and jack springs, special tools are required, as shown in Fig. 24-57. These tools are fitted with a small slot of the correct dimensions to permit them to be slipped over the springs for bending purposes. The springs of a relay or jack should never be adjusted with a pair of pliers.

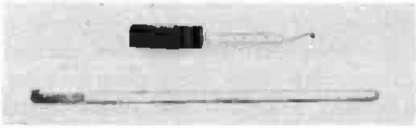


Fig. 24-57. Western Electric 535A and 505A relay-adjusting tools.

24.58 What is a burnishing tool? —A tool used for cleaning or polishing the surface of relay and jack spring contacts (Fig. 24-58). The tool consists of a holder similar to a fountain pen. The burnishing tool consists of a finely sandblasted strip of metal. This strip is placed between the contacts, and the spring is held closed by applying a light pressure with the fingers. The burnishing tool is then rubbed across the contacts and, because of its fine surface, the contacts are cleaned rather than filed. The handle holds spare burnishing blades and round burnishing inserts for jack cleaning. *Emery or crocus cloth must never be used on relay springs.*

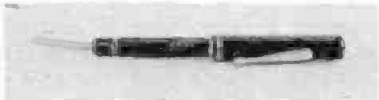


Fig. 24-58. Western Electric 256C burnishing tool.

24.59 Show a circuit suitable for testing patch cords and cables.—Fig. 24-59 shows a patch-cord and cable test set that may be used for checking the continuity and polarity of both single- and double-circuit patch cords, as well as three-pin microphone cables.

The cord or cable to be tested is inserted in the proper receptacle and switches SW1 and SW2 turned to their corresponding positions. If the cord or cable is continuons, a buzz will be heard. If the polarity is reversed, the buzz will not be heard on a corresponding switch position.

The leakage of a cord or cable may be measured by connecting a megger (see Question 25.116) to the center lower terminals and turning switches SW1 and SW2 to opposite positions and switch SW3 to Megger. The buzzer may be tested by turning switches SW3 to the position indicated Buzz, and SW1 and SW2 to position 5.

It is the practice in the motion picture industry to test microphone and camera cables by measuring the dc resistance of each conductor while passing a considerable amount of current through the cable from plug to plug. A record is kept of the resistance of each cable and when repairs are made; this information is used to determine if the cable is normal. This is also a quick and positive test for soldered connections within the plugs.

If the cable is shielded, its continuity to ground is tested by connecting a sharp-pointed probe to the pin jack, and turning switch SW1 to position 1 and SW2 to position 5. The probe is pushed through the outer covering just far enough to make contact with the shield. Continuity will be indicated by the buzzer.

24.60 What is the standard color code used in electronic equipment?—In early electronic equipment a portion of the standard telephone code was employed, with a separate code used for the power wiring. The standard code now used in electronic equipment is given in Fig. 24-60. For electronic equipment a *black* wire is generally used for ground, while in commercial power systems (117 Vac), the Underwriters require the ground wire to be *white* and the hot wire to be *black*.

24.61 What is the color code for small power transformers?—The code is given in Fig. 24-61.

24.62 What is the color code for audio transformers?—The code is given in Fig. 24-62.

24.63 What is the recommended method of testing transmission lines for their operating characteristics?—In large recording plants where many lines terminate at a patch bay, a record is kept of the loop dc resistance of each line and its leakage to ground under both dry and wet conditions of weather. The usual practice is to test the line by shorting the far end and measuring the dc loop resistance with a resistance bridge or, in some cases, a sensitive ohmmeter. The leakage to ground is measured with a megger.

If it becomes necessary to measure the frequency characteristics of a line, this is accomplished as described in Question 23.96.

24.64 Describe a solid-state intercommunication system with a switching matrix.—The basic diagram for a SS-1026 solid-state switching matrix intercommunication system, manufactured by McCurdy Radio Industries (Canada), is shown in Fig. 24-64A. Intercommunication systems such as this one may be used for any purpose requiring selective talking, group talking, or all-talk by the key-switch configuration used. The system to be described involves four main functional elements:

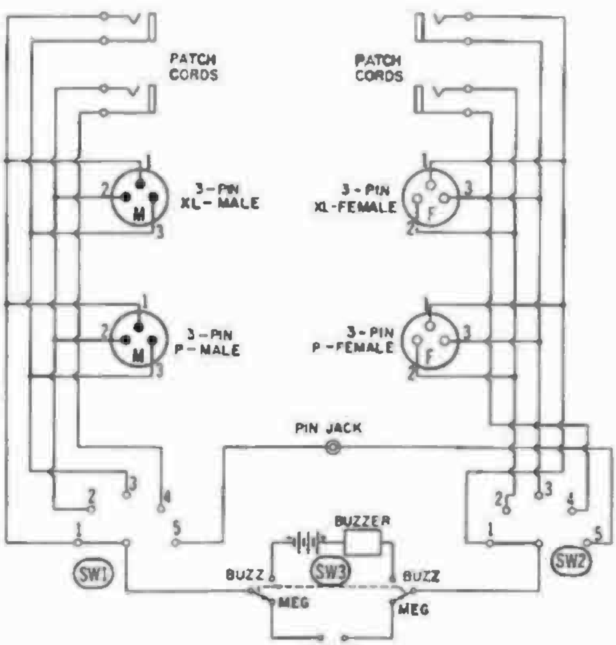


Fig. 24-59. Patch cord and cable test set.

Conductor Number	Base Color	Tracer Color	Tracer Color
1	Black		
2	White		
3	Red		
4	Green		
5	Orange		
6	Blue		
7	White	Black	
8	Red	Black	
9	Green	Black	
10	Orange	Black	
11	Blue	Black	
12	Black	White	
13	Red	White	
14	Green	White	
15	Blue	White	
16	Black	Red	
17	White	Red	
18	Orange	Red	
19	Blue	Red	
20	Red	Green	
21	Orange	Green	
22	Black	White	Red
23	White	Black	Red
24	Red	Black	White
25	Green	Black	White
26	Orange	Black	White
27	Blue	Black	White
28	Black	Red	Green
29	White	Red	Green
30	Red	Black	Green
31	Green	Black	Orange
32	Orange	Black	Green
33	Blue	White	Orange
34	Black	White	Orange
35	White	Red	Orange
36	Orange	White	Blue
37	White	Red	Blue
38	Brown	....	
39	Brown	Black	
40	Brown	White	
41	Brown	Red	
42	Brown	Green	
43	Brown	Orange	
44	Brown	Blue	

Fig. 24-60. Standard color code used in electronic equipment.

power supply, power distribution and control circuits, switching circuits, and input-output amplifiers.

The heart of the system is a solid-state switching matrix (Fig. 24-64A), which determines the routing of the path from any message source in the system to the desired destination and which performs the required switching functions. Essentially the switching matrix is a group of input buses provided with cross-connection points, each of which feeds a pad having a high input impedance and a low output impedance. The output terminating resistance of each pad is part of a second, or output, bus. In some instances, connection between an input bus and an output bus may be required continuously, and the cross-point is simply a resistance chosen to provide the required isolation between the two buses. In a majority of instances, however, a connection from a given input to a given output is required only on an intermittent basis, via a pad which provides the desired amount of attenuation or introduces additional attenuation into a connection which is normally made.

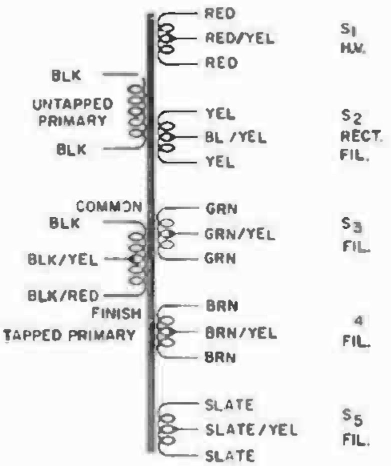
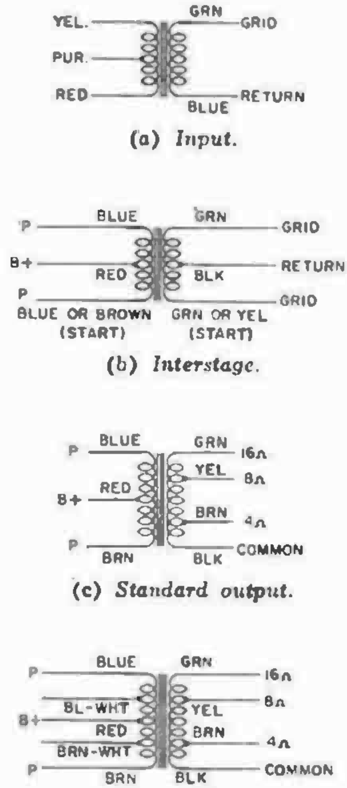


Fig. 24-61. Color code that is used for small power transformers.

In the switching matrix shown, diodes are used to perform the switching functions, and the switching element and isolation pad are combined to form an integrated switching module. Four types of modules are used to accomplish the various types of switching and muting required in television programming. One of these, a "Normally Off/Switch ON" module, is shown in Fig. 24-64B. The others include a "Normally On/Mute 15 dB" module, a "Normally Mute 15 dB/Mute 30 dB" module and a "Normally On/Mute 30 dB" module.



(a) *Input.*

(b) *Interstage.*

(c) *Standard output.*

(d) *Ultralinear.*

Fig. 24-62. Audio transformer color codes.

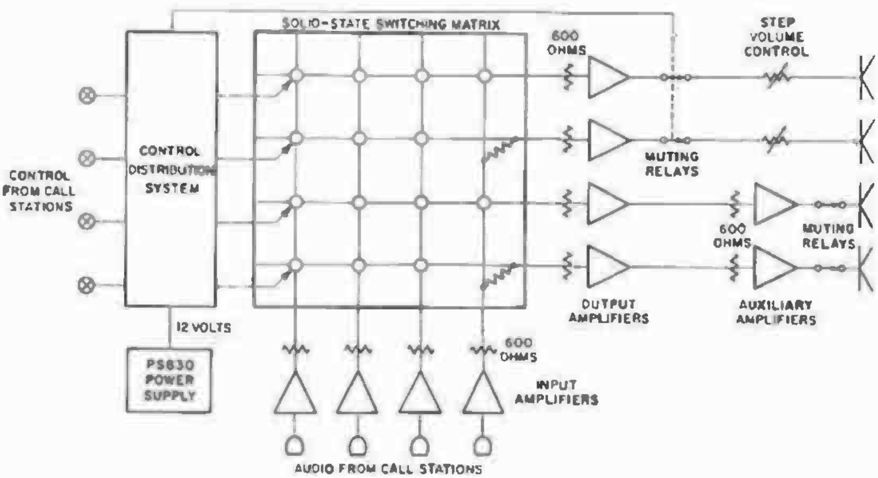


Fig. 24-64A. Basic matrix for McCurdy Radio Industries solid-state intercommunication system.

The minimum attenuation in any cross-point is 30 dB, so that isolation between any two input/output paths is at least 60 dB. For continuous connections the 30 dB of attenuation is provided by a single 15,000-ohm resistor. The various types of switching modules present an impedance of more than 10,000 ohms to the input bus, so that it may feed a large number of output buses without being affected by loading. As shown in the block diagram, the cross-points are designed to feed a 600-ohm resistive load. To eliminate pickup in the matrix from external sources, rf bypass capacitors are used across all terminating resistors.

To permit switching at a common level and to overcome the attenuation required for isolation in the matrix, both input and output amplifiers are used. Two types of input amplifiers are provided: a carbon-microphone/line-input amplifier, and a dynamic microphone amplifier which incorporates compression. Both types of amplifiers are designed to supply 0 dBm into the 600-ohm impedance of the input bus. Two types of output amplifiers are also

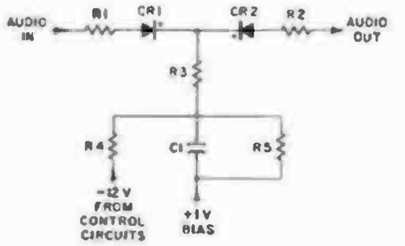


Fig. 24-64B. Matrix switching module.

available: a full-frequency-range unit supplying +18 dBm into 600 ohms, and a limited-range unit which provides the same output power but has a tailored frequency response to improve voice intelligibility in noisy locations.

Relays are provided to mute the local speaker when an intercom selector switch is depressed at locations where acoustical feedback might occur. In applications such as studio address, where higher amplifier powers are required, the output amplifier drives auxiliary units.

Connections between sources and destinations within the system are established by a dc control system

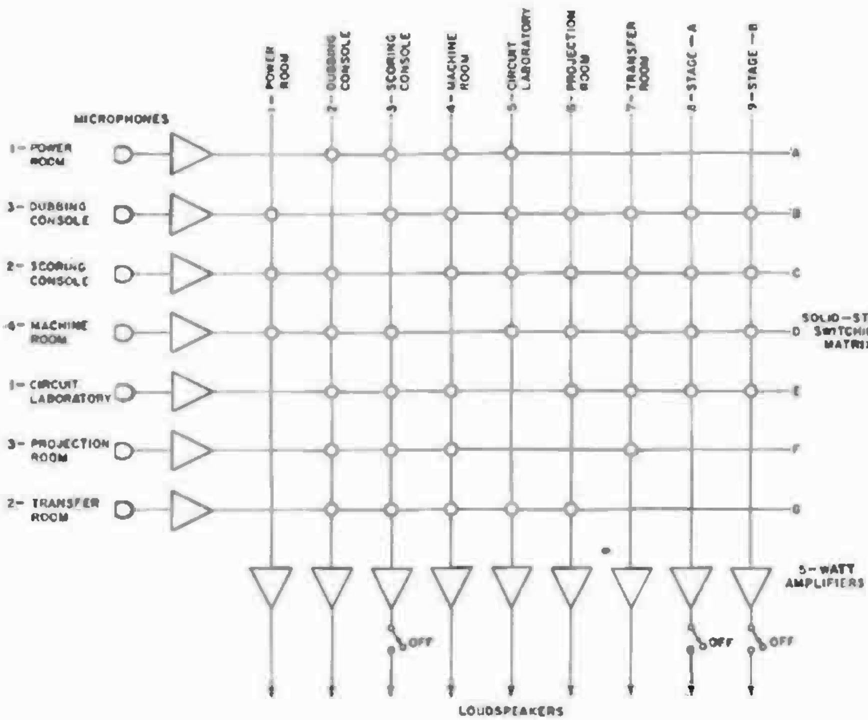


Fig. 24-64C. Typical matrix layout for a sound department intercommunication system. (Courtesy, McCurdy Radio Industries Ltd., Canada)

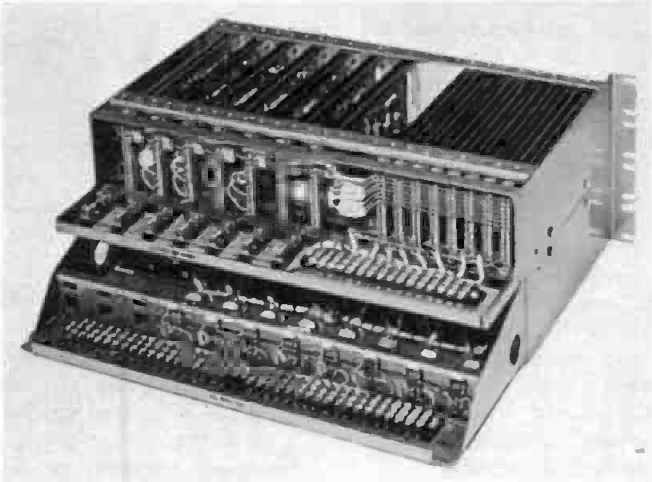


Fig. 24-64D. Matrix and amplifier modules for McCurdy Radio Industries Ltd. (Canada) Model SS-1026 solid-state intercommunication system.

which applies 12 volts to the appropriate module when a selector key is depressed. The modules are normally held off by a 1-volt bias voltage. The basic

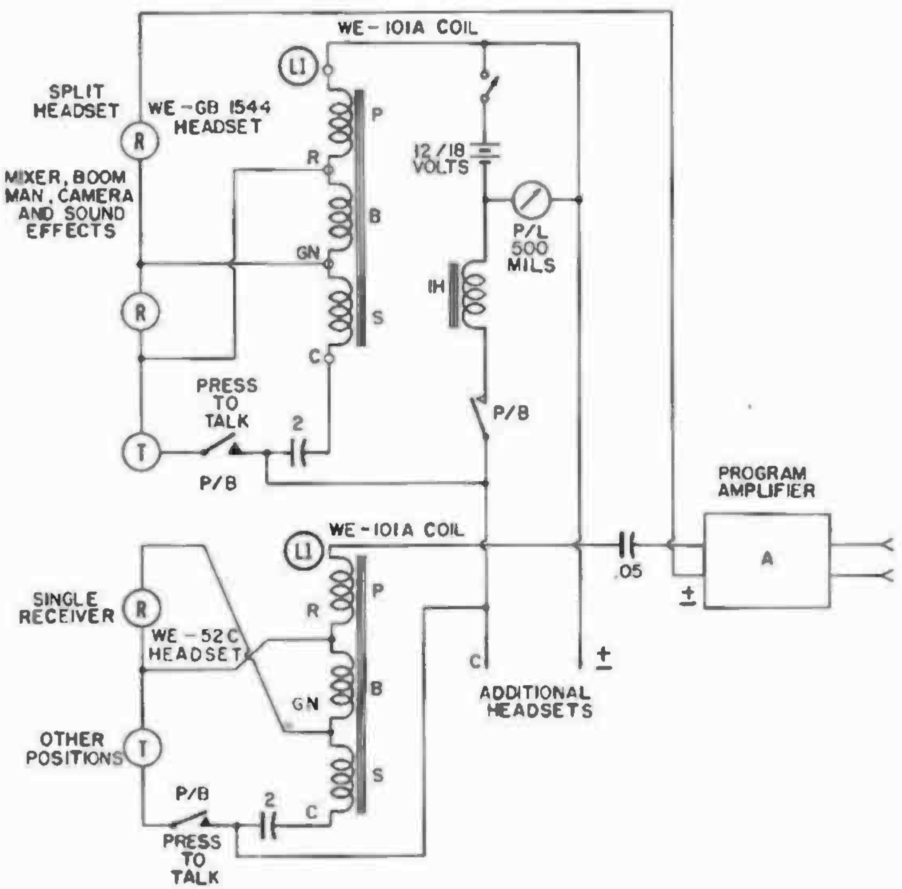


Fig. 24-65. Split-headphone intercommunication system suitable for television or recording.

diagram for a typical switching module is given in Fig. 24-64B. The power supply is solid state and supplies 12 volts at 5 amperes for both the amplifiers and control circuits.

A typical switching matrix for a sound department in a large plant is given in Fig. 24-64C. A rear view of the amplifier-matrix chassis is shown in Fig. 24-64D. The module cards for the amplifiers and switching diodes may be removed from the front. (See Questions 24.65 and 24.66.)

24.65 Show an intercommunication system suitable for use with television productions, using split headphone monitoring.—The upper left-hand portion of Fig. 24-65 shows a split headphone monitor and order-wire circuit. The split circuit is generally confined to use by the mixer, boom men, sound-effects man, and cameramen. For other positions in the chain where monitoring of the program material is not important, one receiver is connected to the order-wire circuit only. It will be noted a program amplifier with a low output impedance feeds the receivers of the split monitoring circuit. The impedance feeding this circuit will depend on the number of headphones used and their internal impedances.

Additional headsets of the split- or double-circuit type may be added, as desired, by connecting them in parallel with the circuit. The 1-henry choke in the battery circuit is necessary to prevent the low impedance of the battery from short circuiting the coil. It must be capable of carrying the current of all transmitters. A single transmitter draws about 30 milliamperes of current.

The press-to-talk buttons are necessary to eliminate talking and noises that would be picked up by the microphones, if directly connected to the circuit at all times.

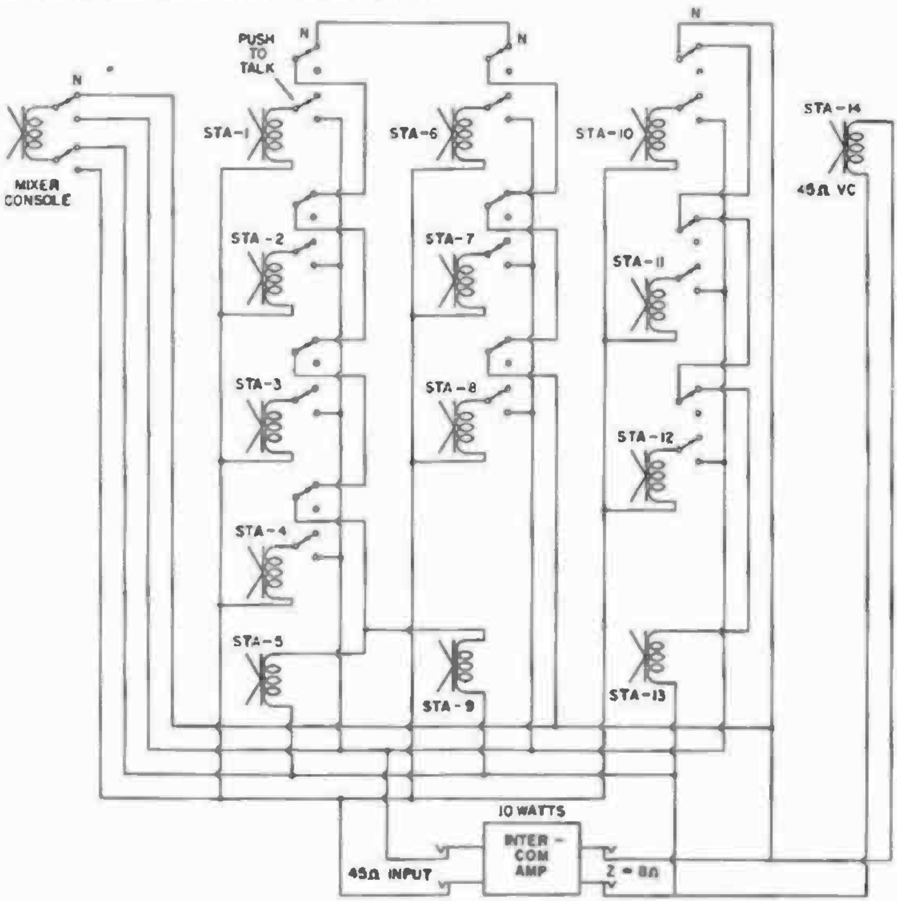


Fig. 24-66. Talk-back system using a 45-ohm loudspeaker for both microphone and speaker.

24.66 ***Describe a talk-back system using speakers for both microphone and speaker.*** — Fig. 24-66 shows a simple talk-back system suitable for use in a large recording department. The system consists of a group of 3-inch speakers with a 45-ohm voice-coil winding connected to a push-to-talk switch to function as both microphone and speaker.

The system is set up in such a manner that the station at the mixer console can always talk to a given group of stations, such as the projection room, machine room, narration booth, and the recording machine operator. The mixer speaker circuit is normalled and is available to all stations by operating the push-to-talk button at any one station. Other stations can originate a call, but cannot be called unless they are within hearing distance of one of the normalled stations.

The diagram shown may not apply to all situations; however, it can be readily adapted to any installation by the proper connections at the press-to-talk button.

24.67 ***What is a depopping or click suppressor?***—A capacitor and resistor connected across the contacts of a switch or relay to prevent a surge being introduced into an adjacent circuit. Click-suppressor circuits take several forms. The most common is a capacitor of 0.10 to 1 μF and a resistor of 1 to 10 ohms connected in series with the capacitor across the coil or contacts of the offending relay or switch, as shown in Fig. 24-67A. It is common practice in large recording installations, where several distributor sets or motors must be operated simultaneously, to connect click suppressors across all remote dc relay coils in starting boxes. No hard and fast rule can be laid down for such devices, as each installation presents its own problem. Many times the resistor will not be required, only the capacitor.

The best manner in which to approach the problem of click suppression in a recording installation is to set the gain of the playback system about 20 dB above normal and operate the control circuits while listening for clicks. If a click is heard, a click suppressor with a capacitor in series with a 10-ohm resistor is connected across the coil of the relay at the coil terminals. The value of the resistor is varied for the best results. Additional suppressor circuits are added to other parts of the control system until the clicks are reduced to inaudibility.

The results are finally checked by making a recording while starting and stopping the equipment. Usually the greatest disturbance will be noted when the equipment is stopped because of the collapsing field around the relay coils. The capacitors must be of 600- to 1000-volt rating since the peak voltage across a 24-volt circuit can rise to 1000 volts.

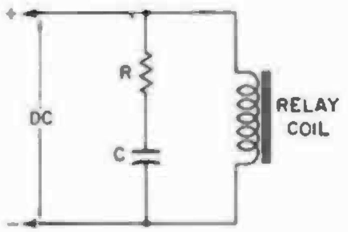


Fig. 24-67A. Click-suppression circuit using a resistor and capacitor. In some instances, only the capacitor will be required. Used in dc circuits only.

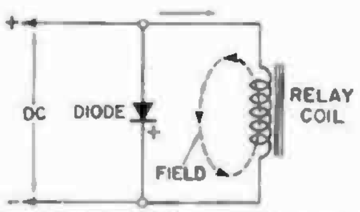


Fig. 24-67B. Click-suppression circuit connected across a dc relay coil.

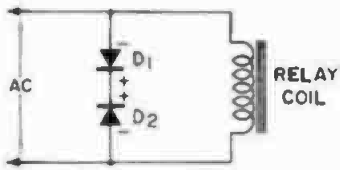


Fig. 24-67C. Click-suppression circuit suitable for either dc or ac control relay circuits. The diodes are connected back to back.

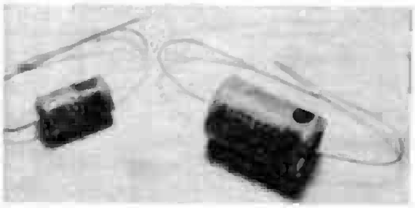


Fig. 24-67D. Nonpolarized click suppressor or depopping device, manufactured by International Rectifier Corp.

The resistor must be of the composition type, not wirewound, to reduce inductive effects. In some instances it may be necessary to completely enclose the relay in a high-permeability metal shield to reduce surges being transmitted electromagnetically to surrounding equipment. Every piece of equipment must be well grounded, particularly generators, motor frames, and starting boxes.

Many times a ground loop in the system will cause clicks to be transmitted to a remote part of the installation. This makes it doubly important that the system be installed in such a manner that ground loops are avoided. This subject is discussed in Question 24.33.

One of the best ways to avoid clicks in a large plant is to twist tightly all ac lines, both single- and three-phase, during the installation of the equipmeut. This also applies to dc lines as well. (See Question 24.49.)

Click suppression may also be achieved in dc circuits by using a small diode rectifier connected across the offending circuit. First, the polarity of the circuit is determined. Then the diode is connected in reverse polarity across the terminals of the relay control coil; that is, the positive terminal of the diode is connected to the negative terminal of the control voltage, as shown in Fig. 24-67B. If the diode is not connected as shown, it will not function, as its purpose is to short circuit the collapsing magnetic field of the relay coil when the control voltage is removed. The polarity of the collapsing field is opposite to the polarity of the control voltage. Also, if the foregoing precautions are not observed, the diode will conduct when the control voltage is applied to the coil, and the diode will be damaged.

Diodes selected for click suppression must be capable of carrying high current because the current rises to a high value when the circuit is broken. For use across relay coils operating at 28 volts or less, the 1N1764 diode serves quite well, since it has an instantaneous forward-current rating of 15 amperes.

The voltage induced in the relay coil when the contacts of the control circuit are opened may be calculated as follows:

$$E_r = \left(1 + \frac{R_s}{RL}\right), \text{ when } t = 0$$

where,

E_r is the voltage built up in the coil,
E is the dc control voltage,
R is the dc resistance of the coil (E/I),
R_s is the forward resistance of the diode,
L is the inductance of the relay coil in millihenries.

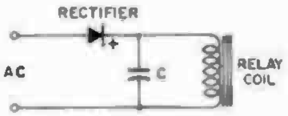
The value of the peak voltage (E_r) across the coil is dependent on the value of the control voltage and the ratio of coil resistance to forward resistance of the diode. The lower the resistance of the diode, the lower the peak voltage across the coil; however, the decay time (to zero) becomes longer.

The decay time can be reduced by the use of the circuit shown in Fig. 24-67C, consisting of two identical diodes connected back to back, which will short circuit voltages equally of either polarity.

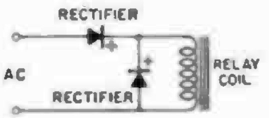
With a high induced voltage and a low diode reverse resistance, considerable current flows through the diode, quickly dissipating the induced energy. The induced voltage across the coil will drop rapidly, lowering the reverse voltage across the diode. Lowering the reverse voltage increases the diode resistance, and decreases the current proportionally. In this manner the nonlinear characteristics of the diode are taken advantage of to dissipate the induced energy in the coil, and at the same time to apply a damping effect across the coil, reducing the decay time. The circuit of Fig. 24-67C may also be used across ac control circuits. Regardless of the type of suppressor circuit employed, it must be connected physically at the terminals of the relay coil.

For click suppression of ac-operated relays, the back-to-back connection is used exclusively. These diodes are especially developed for this service, and are rated by their voltage operating conditions. They may be obtained for both single- and three-phase operation and from 27 to 420 volts rms. The clamping voltages range from 85 to over 1000 volts (clamping voltage is the voltage across the coil at the instant of breaking the circuit), with peak discharge currents of 180 amperes. It is not uncommon in a large sound installation to use several hundred of the devices. A typical ac back-to-back suppressor is pictured in Fig. 24-67D, and is manufactured by International Rectifier Corp. (See Question 11.158.)

24.68 ***How can dc relays be prevented from chattering when operated from a half-wave rectifier?*** — Relays operating from a half-wave rectifier circuit should always use a relay coil with a copper ring to prevent chattering. For coils without a copper ring, a capacitor may be connected in parallel with the coil winding, as shown at (a) of Fig. 24-68, or a half-wave, back-wave rectifier element may be connected across the coil as shown at (b) of Fig. 24-68.



(a) Capacitor across coil.



(b) Back-wave rectifier to smooth current through coil.

Fig. 24-68. Circuits for eliminating chatter in relays operating on rectified ac without the benefit of a copper ring.

24.69 ***How are telephone lines transposed and what is the purpose?***— Transposing telephone lines is a method used for the reduction of cross talk between adjacent pairs which are run parallel for relatively long distances.

Cross talk can be reduced to a great extent by transposing the lines at intervals on the pin positions of the telephone pole crossarms. The technique for transposition becomes quite involved, particularly if a large number of pairs is involved. The length of the transposition is dictated by the frequency of operation and the cross-talk tendency of the pattern employed.

Fig. 24-69 shows the basic patterns standardized by the Bell Telephone System for the transposition of telephone line pairs. (See Questions 23.97 and 23.98.)

24.70 ***Describe a 70.7-volt constant-voltage speaker distribution system?***—It is a method of matching speaker impedances for paging and public address systems. This system was adopted by the EIA and is slowly being adopted by the manufacturers of announcing systems equipment.

The 70.7-volt method of impedance matching is based on a constant-voltage system and, to a great extent, simplifies the computation of transformer imped-

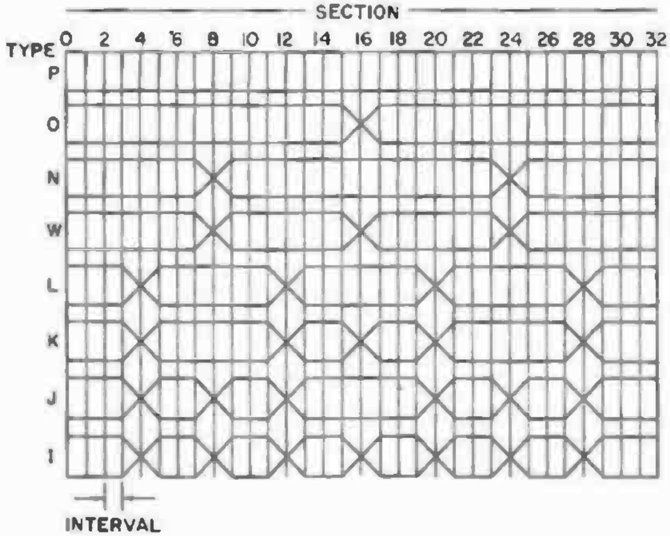


Fig. 24-69. Basic patterns standardized by the Bell Telephone System for the transposition of telephone line pairs.

ance ratios when different sound levels are required at different points in the system. The 70.7-volt system permits the addition of speakers to an existing system without recalculation of the load and source impedances.

As long as the total power consumed by the speakers is less or eqnal to the amplifier power rating, favorable load conditions will prevail. Furthermore, the 70.7-volt system permits the connection of speakers across a transmission line in a manner similar to that of connecting a light bulb across a power line, up to the capacity of the power circuit. The system requires that the amplifier output transformer deliver 70.7 volts at its rated power outpnt for a given percent harmonic distortion. The EIA selected 70.7 volts because of Underwriters' requirements which, in many locations throughout the country, limit speaker voltages to 125 volts, unless the lines are run in conduit.

The term 70.7-volt speaker line does not mean that the voltage on the line is always 70.7 volts. The voltage will vary, depending on the power level at which the line is operated. The 70.7 volts is the maximum voltage of sine-wave character for a given amount of distortion. Standardizing the output voltage means the voltage will be the same for either a low- or a high-power amplifier.

The value of 70.7 volts will only exist at the output of the amplifier when it is terminated in its specified terminating impedance and operated at its rated power output.

If the transformers at the speakers are rated in watts, no calculations are required. For transformers not marked in watts but in load impedance, the following formula may be used:

$$Z = \frac{E^2}{P}$$

where,

Z is the outpnt impedance.
P is the desired level in watts at a particular location.

For an amplifier with a 70.7-volt output, the formula is reduced to:

$$Z_s = \frac{5000}{P}$$

where,

5000 is the approximate square of 70.7,
P is the desired power level in watts,
Z_s is the unknown impedance.

After a plan has been developed showing the power required at each individual speaker in the system, the powers are totaled and an amplifier is selected which is capable of supplying at least the total power and greater, if practicable. If a single amplifier is not practicable, the system may be divided into two or more branches, each amplifier being fed from a common input source impedance.

Two or more power amplifiers may be used by connecting their outputs in parallel, if they are of similar design; however, this is not recommended unless the branch system is impracticable. The connection of power amplifiers in parallel is discussed in Question 12.201. After the 70.7-volt line system has once been established, 70.7-volt matching transformers may be connected across the line without further consideration as to impedance match.

To use the conventional line transformer with a 70.7-volt system, assume three speakers are to be fed, 10 watts, 5 watts, and 1 watt of power, respectively. The primary impedance is calculated as follows.

For the 10-watt speaker:

$$Z_p = \frac{E^2}{P} = \frac{5000}{10} = 500 \text{ ohms}$$

For the 5-watt speaker:

$$Z_p = \frac{E^2}{P} = \frac{5000}{5} = 1000 \text{ ohms}$$

For the 1-watt speaker:

$$Z_p = \frac{E^2}{P} = \frac{5000}{1} = 5000 \text{ ohms}$$

where,

E is the line voltage (70.7 volts),
P is the power in watts at the speaker,
Z_p is the primary impedance to be connected across the transmission line.

Impedance mismatches up to 15 percent are permissible. The total power must not exceed that of the maximum power output of the amplifier. The 70.7-volt transformer must be capable of carrying the full power output of the amplifier. Circuits not in nse must be terminated in a dummy load of the correct impedance and power rating.

24.71 How can high-frequency radio interference be prevented or eliminated from high-gain audio circuits?—Because of the increased use of high-frequency radio transmissions, problems of interference in audio circnits have been

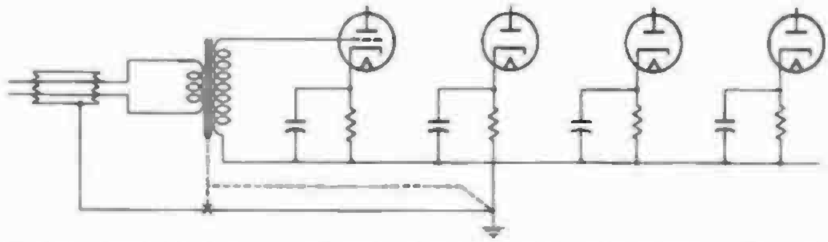


Fig. 24-71A. Method of grounding the shield on an incoming line to reduce radio interference. The transformer core is grounded at the point having the least amount of interference.

intensified. Many radio transmitters, such as police and broadcasting stations, cause interference in high-gain systems, such as motion picture theaters and public address systems.

By observing certain precautions during the installation of such equipment, a satisfactory elimination or suppression of radio interference may be accomplished. In general, interference of this nature can be minimized by improving the effectiveness of the shielding, preventing rectification of the interfering signal, or eliminating it by the use of radio-frequency filters. Where interference from radio high-frequency energy (30 MHz and up) is encountered, the suggestions to follow may be of help.

In view of the fact that rectification generally occurs in an early amplifier stage and a large amount of gain is usually necessary before the interference can be detected at a speaker, special consideration should be given to the low-level points in a system, particularly where photocell amplifiers, magnetic reproducing heads, pickups, and microphones are concerned. The system should be connected to a true ground as closely as the physical location will permit. A cold-water pipe is generally the best source of grounding. Under no circumstances should a gas pipe be used, because an insulated bushing is generally inserted at the meter; also, the compound used for sealing the joints offers a high resistance.

Electrical conduit should not be used for grounding because it is connected to many different pieces of electrical equipment which induce ground noise. As this noise travels along the conduit, it is effectively connected in series with the ground return of the amplifier system. (See Question 24.32.) Whenever possible, the ground system should be continuous and the shields of the system tied to a single point near the low-level stages. Cable shields must be continuous without breaks or high-resistance solder joints. Single ground wires are run from sound heads and other low-level devices directly to this ground point.

It is essential the shielding be connected to the transmission ground at a point that will not permit radio-frequency energy picked up by the shields to pass through the transmission ground system. Fig. 24-71A shows the proper method of connecting a ground point and the core of an input transformer. The success or failure of a grounding system is controlled by such small details.

The elimination of radio-frequency energy from a high-gain sound system does not always follow conventional methods of grounding as described in

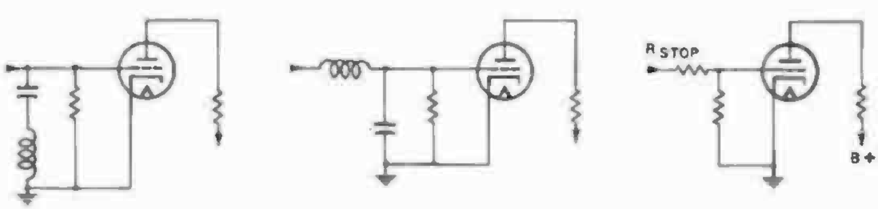


Fig. 24-71B. Grid-stopping resistor and radio frequency filters for eliminating radio frequency interference in a high-gain audio system.

other parts of this section. Many times it requires a combination of several methods to completely eliminate the interference.

When multiple ground connections to shields or equipment cannot be avoided, interference due to resonance of loops formed by the shielding can be minimized by reducing the contact resistance, shifting the position of the contact, changing the length of the leads, or otherwise shifting the resonant period of the loop causing the interference.

Leads running to an amplifier rack must be well shielded. Contact resistance between lead-shielded wire (if used), conduit, and flexible tubing (Greenfield) must be kept low.

Radio interference can also be caused by a poor solder joint causing rectification. Electrolytic capacitors that have been in service several years can often cause the difficulty, because of high internal resistance (power factor) as a result of their age. This latter fault may be checked for by paralleling a new capacitor across the cathode circuit, or any other circuit that depends on an electrolytic capacitor to supply a low-impedance path to ground.

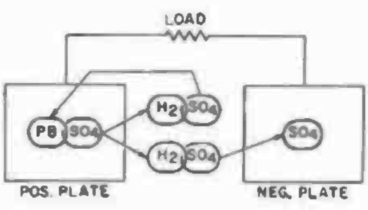
If these corrections do not effect a complete cure, small radio-frequency chokes may be connected in the cathodes of the low-level, high-gain stages. As a rule, for high-frequency interference, a choke of 2 to 4 microhenries of inductance will suffice. However, for broadcast frequency interference, about 5 millihenries will be required. These chokes are connected at the socket at the top end of the cathode circuit and must not be bypassed. (See Question 12.197.)

At times it may be necessary to install radio-frequency filters as shown in (a) and (b) of Fig. 24-71B. The circuit component values should be made as large as possible without affecting the frequency response of the amplifier system. Grid-stopping resistors are connected at the socket pins as shown at (c) of Fig. 24-71B, and range from 1000 to 50,000 ohms in value. If a filter is required, as shown in (a) and (b), about 2 to 6 microhenries of inductance will be required, with a capacitance of about 10 picofarads. As a last resort, filters may be installed in the power lines feeding the equipment.

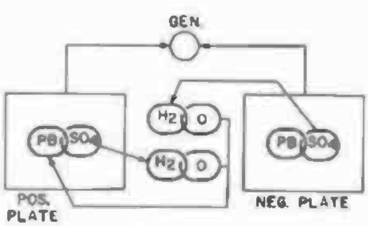
24.72 ***What chemical changes take place when a storage battery is charged or discharged?***—The lead-acid storage cell is an electrochemical device for converting chemical energy into electrical energy. Substances within the battery react chemically to produce an electrical current whenever an electrical device is connected to the battery terminals. When these substances in effect have been used up, the battery is said to be discharged. The battery must then be recharged by a current into the battery from an external source, such as an ac rectifier or generator.

A storage battery consists of one or more cells consisting of internal elements called positive and negative plates filled with a special lead-oxide paste in a rectangular gridlike structure. The plates, after processing to make them solid but porous, become the active materials of the plates after the battery is charged. One type of lead oxide is applied to the positive plates while a different type is applied to the negative plates. Insulated separators prevent the plates from touching and shorting.

The chemical activities within the battery during the charge and discharge cycle are shown diagrammatically in (a) of Fig. 24-72. When a battery is in a fully charged state, the positive plates contain essentially pure lead oxide (PbO_2), which is chocolate brown in color, and the negative plates contain essentially pure sponge lead, which is



(a) Discharge cycle.



(b) Charging cycle.

Fig. 24-72. Chemical action in a lead-acid type storage battery during its discharge and charging cycles.

gray in color. The electrolyte consists of a fairly concentrated solution of sulfuric acid in water. The specific gravity of the solution is approximately 1.290 at 80°F. During the discharge cycle when the battery is connected to an external load, the active materials in the positive and negative plates (PbO_2 and Pb) react chemically with the sulfuric acid as shown in (a) of Fig. 24-72. The sulfuric acid (H_2SO_4) begins to break up. Part of the sulfate (SO_4) enters the positive plate and replaces the oxygen (O_2) while sulfate also combines with sponge lead in the uegative plate. The displaced oxygen leaves the positive plate and combines with the hydrogen (H_2) to form water (H_2O). Thus, in both plates, lead sulfate is formed; the sulfuric acid is replaced with water.

As the discharge continues, more and more of the sulfuric acid combines in this manner with the materials in the positive and negative plates. Finally, most of the active materials will have been converted into lead sulfate, while most of the sulfuric acid will have been used up in the discharging process. The battery is then said to be discharged and cannot deliver any appreciable additional current and must be recharged. During the charging cycle, the chemical action is reversed, as shown in (b) of Fig. 24-72. The sulfate leaves both positive and negative plates to combine with hydrogen in the water. The oxygen displaced by this action re-enters the positive plate, where it combines with the lead once again to form lead peroxide, and the negative plate again becomes a spongy metallic lead.

24.73 What is the proper method of operating a storage battery with a trickle charger?—The use of a trickle charger with a storage battery is detrimental to the battery, as it shortens the life of the battery as the result of overcharging. Trickle chargers should only be used when it is impractical to charge a battery by other means. A practical approach to the problem is to adjust the charging voltage to a value between 2.15 and 2.17 volts.

A better, but more elaborate method, is to check the specific gravity of the cells over a period of several months and to adjust the charging voltage to a value where the specific gravity is maintained at 1.250. Compensation must be made for temperature changes when reading the specific gravity during the cold and warmer months of the year, unless the battery is housed in a room where the temperature is fairly constant. Four gravity points are added to the reading for every 10 degrees the electrolyte is above a temperature of 80 degrees Fahrenheit. A like amount is subtracted for temperatures below 80 degrees Fahrenheit.

The freezing point of a battery electrolyte depends on the specific gravity of the electrolyte. The figures below show how freezing varies with temperature.

Specific Gravity	*Freezing Point*
1.275	−85°F
1.250	−62°F
1.225	−35°F
1.200	−16°F
1.175	+ 4°F
1.150	+ 5°F
1.125	+13°F
1.100	+19°F

24.74 How is the internal resistance of a battery calculated?—To calculate the internal resistance of a single cell or battery, the open-circuit voltage is measured using a meter with an internal resistance of 1000 ohms per volt. Suppose this voltage (E_1) is 10 volts. The battery or cell is then loaded with a 50-ohm resistance, R_1, and the voltage drop measured. Assume this voltage drop E_2 to be 9 volts. The current through the resistor R_1 is:

$$\frac{E_2}{R_1} = \frac{9}{50} = 0.180 \text{ ampere}$$

The internal resistance of the battery may now be calculated:

$$\frac{E_1}{I} - R_1 = \frac{10}{0.180} - 50$$
$$= 55.5 - 50 = 5.5 \text{ ohms}$$

The internal resistance equals 5.5 ohms.

24.75 What is gimp? — A slang name given to the extremely flexible wire used in telephone cords and similar equipment. This wire cannot be directly soldered to, as it is a metallic cloth-type material. The correct way to terminate such wire is to serve the end with fine bright copper wire, then dip the end in molten solder. If cord is to be terminated in a telephone cord tip, the tip is filled with solder, and while hot, the end of cord is inserted into the molten solder and held nntil the solder hardens.

24.76 ***What is the purpose of connecting a storage battery in parallel with a low-voltage power supply?***—To stabilize the output voltage. The battery acts as a voltage regulator and helps to maintain a constant output voltage under varying load conditions.

Due to the fact that a storage battery has a very large internal capacitance, the ripple and noise voltage will be reduced to a minimum. Since the battery supplies only a small portion of the load current, the condition of the battery need not be the best, but it must not be shorted.

24.77 ***Describe a simple intercommunication system using a sound-powered unit.***—Many times on a motion picture set a one-way intercommunication is required that will work into a spare position of a sound mixer panel. Such a device is shown in Fig. 24-77. A sound-powered telephone unit is mounted in a small box which may be held in the hand. A spring-loaded switch in its normal position terminates the input of the spare mixer position in 10 ohms, and disconnects the sound-power unit. The output is connected to a standard microphone cable compatible with the sound mixer in use.

24.78 ***Describe the different methods for lettering panels.***—There are three main methods for lettering panels. They are engraving the letters in the panel, use of wet decalcomania transfers, and the use of dry decalcomania transfers.

The first method requires an engraving machine and a set of master letters, numerals, and symbols. Although it is the most durable method of lettering, it is the most costly of the three methods. It is the practice when engraving a panel to first paint the surface and then engrave. The letters are then filled in with a filler of the desired color. If the panel is aluminum, no filler is required, since the aluminum panel will show up the letters quite nicely, particularly if the panel is a contrasting color.

In the use of wet decalcomania transfers, symbols and letters are mounted on a paper backing. The letters are moistened with water and slid off the paper onto the panel. When dry, the lettering is sprayed with lacquer.

The third and most convenient to use is the dry transfer system. The letters are attached to a thin paper backing. After placing the letter in its proper position on the panel, the letters are rubbed with a ball point pen or stylus and will adhere to the panel. Several coats of lacquer are then sprayed over the surface. Dry lettering is also available for making meter dials.

Still another method that may be useful at times is the use of bakelite laminate. Such material may be obtained with the laminate second layer available in several colors. Engraving through the top layer permits the colored laminate to show up in the engraved letter. This method requires no filler.

24.79 ***Describe a growler and its use.***—A growler is a device used for detecting shorted turns in motor and generator armature coils. The device shown in Fig. 24-79 consists of a laminated core open at one end and a coil around the closed leg. When the suspected armature is placed between the core ends, the armature completes the magnetic circuit. Applying an ac voltage to the coil on the core sets up an alternating magnetic flux through the armature.

As the alternating flux flows through the coils of the armature, it induces a potential in the coil. If the coil has a shorted turn, a circulating current is generated in the coil, resulting in a loud

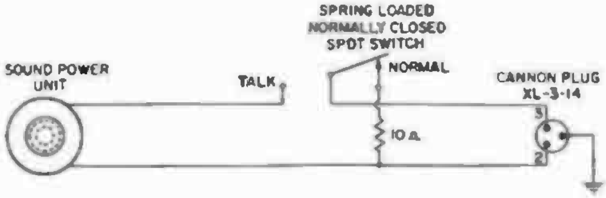


Fig. 24-77. Simple intercommunication unit consisting of a sound power unit for operating into a sound mixer.

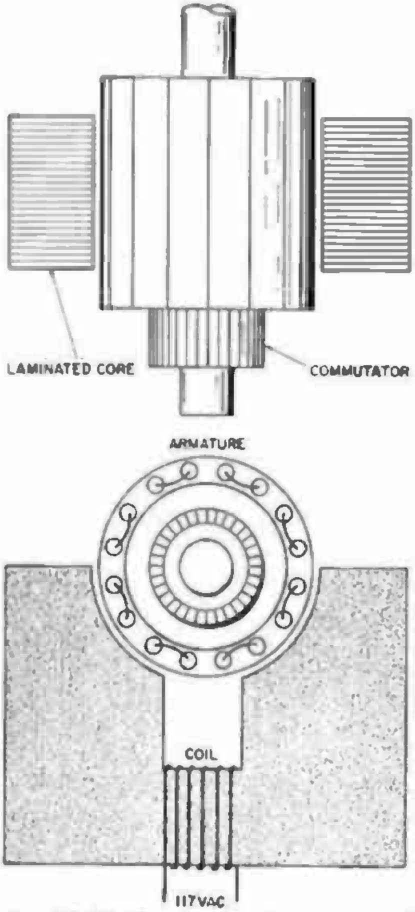


Fig. 24-79. Growler for test motor and generator armature coils.

buzzing sound. Rotating the armature causes the sound to change each time the shorted coil is turned to the center of the pole pieces. Thus the defective coil is indicated. Such devices are used extensively in motor repair shops.

24.80 What are the console dimensions recommended for comfortable operation of electrical and mechanical equipment by humans?—The Bell Telephone Laboratories have made extensive studies of the human body for the most efficient and comfortable operation of both mechanical and electronic equipment to be used as a guide for the construction of mixer consoles or racking up of test equipment. A console should be designed for comfort as many long hours are given to its operation. The height of the console should be low enough that each component of its operation may be easily reached, and the far edge (back) also low enough so that it may be easily seen over by the mixer. The slope of the control panel should be at an angle that causes no strain on the arms of the operator. In addition, the front edge of the console should be padded, as shown in Fig. 9-46A.

Although the dimensions given in Fig. 24-80 are for a table or rack, they are still quite satisfactory for many uses. Dimensions given are for the average-sized man. For a large man the

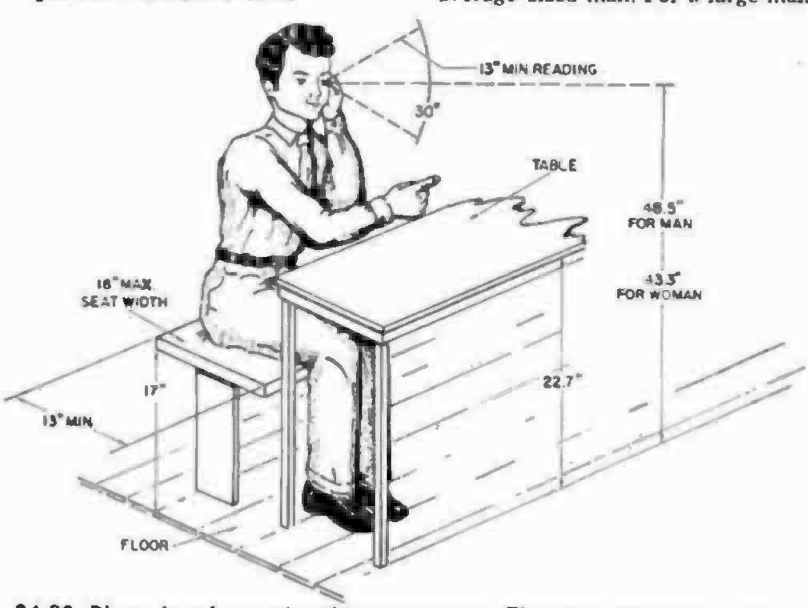


Fig. 24-80. Dimensions for seating the average man. The above diagram may be used as a guide for designing mixer consoles and test equipment racks. (Courtesy, Bell Laboratories Record.)

seat is increased 2 inches and the eye dimensions raised 3.2 inches.

24.81 Describe a zero-suppressed meter. — Zero-suppressed meters are used where only a limited scale of current or voltage is required. By limiting the indications of the meter, an expanded scale can be obtained, with more precise readings within a given range. A typical ammeter of this type, covering a range of 4 to 8 amperes for reading exposure-lamp current in a photographic film recorder, is shown in Fig. 24-81. The most used portion of the scale is that between 5.8 and 7 amperes.

Fig. 24-81. Suppressed zero dc ammeter, used for reading exposure lamp current in a photographic film recorder. The most used portion of the scale is between 5.8 and 7 amperes.

24.82 Describe a flexible mounting for isolating vibration.—One of the most useful shock-mounting devices for eliminating vibrations of motors or other devices, which is manufactured by Lord Manufacturing Co., is shown in Fig. 24-82A. It is used in many different types of electronic equipment where the effects of vibration may be detrimental to its operation.

The mounting consists of a rubber insert bonded to a metal plate (Fig. 24-84B). The center member is tubular for bolting the supported device to the rubber shock mount. The load rating designates the load which the mounting will carry for a pre-determined axial deflection of the floating member. The metal plate is attached to the supporting structure with rivets or screws. Such shock mounts may be obtained for various loads and in many different shapes and sizes.

Fig. 24-82A. Flexible shock mounting made by the Lord Manufacturing Co.

24.83 What is the purpose of connecting a shunt across a meter movement during transportation? — Connecting a wire in shunt with the meter terminals during shipment prevents the movement from being damaged. The meter movement during transportation swings back and forth, and this movement generates a small voltage at the terminals. Short circuiting the terminals loads the meter movement, dissipating the power generated and effectively damping its motion; thus, the movement remains essentially at rest during transportation.

24.84 What is a voltage-dependent resistor (VDR)? — A voltage-dependent resistor does not maintain a constant value of resistance as does the conventional resistor, but its resistance is entirely dependent on the voltage impressed across its terminals. Such devices are used for the protection of

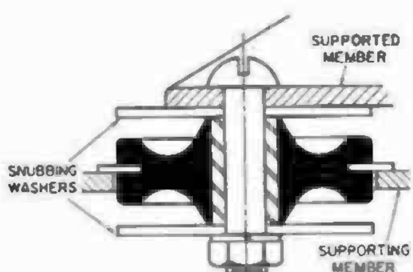


Fig. 24-82B. Method used for mounting flexible shock mounts shown in Fig. 24-82A.

relay contacts, voltage stabilization in circuits where the voltage is continuously changing, and in many other applications.

The material, used for making VDR's consists mainly of silicon carbide (belonging to the semiconductor family). The voltage dependency is caused by the contact resistance between the carbide crystals. The electrical resistance is determined by the crystal contacts which form a complicated network of series and parallel resistors. To put the materials into a practical form, the silicon-carbide grains are pressed together with a ceramic binder, in the shape of a disc or rod, and sintered at high temperature. Terminals are metalized for good contact. When finished, they are comparable to unglazed ceramic.

As the voltage changes across a VDR, its resistance undergoes a change because of the increasing or decreasing current through the VDR. The relationship between the current and voltage for a particular VDR can be approximated:

$$V = C \times I \times \beta$$

where,

- C is the voltage across the VDR terminals,
- I is the current in amperes,
- β is an exponential coefficient approximately constant for a given VDR.

24.85 What is the effect of heat on permanent magnets?—There are three noticeable effects. They are the *reversible* effect, *irreversible* effect, and the *material* effect. The remanent induction decreases as the temperature is increased. For small changes this decrease can be entirely reversible with no loss after returning to room temperature. For a wider temperature range, an irreversible loss of remanence is likely to occur in addition to the reversible change (on return to room temperature). The remanent induction is below the original value, which may be regained by remagnetization. At high temperatures the material itself may change, influencing the hysteresis loop and remanent induction. Changes in the material may result in lowering or raising the remanent induction. The original value cannot be regained by remagnetization. At high temperatures all three changes can occur. Irreversible losses occur only during the first three cycles, and the changes of remanent induction become smaller with each cycle. In this manner a magnet may be stabilized for a given temperature range, through at least three exposures. (See Questions 17.21 and 20.23.)

24.86 How are miniature lamps (pilot lights) coded?—Outline drawings and coding information are given in Fig. 24-86.

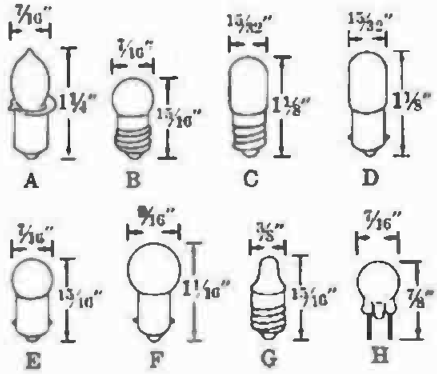
24.87 How can the wire size and voltage drop for regulated power supplies be determined?—The voltage drop can be determined by Ohm's law. However, by the use of the nomograph in Fig. 24-87, the problems are simplified. Since regulated power supplies are designed to control the output at the power-supply output terminals, the conductors used for the supply line must be considered as a part of the power-supply load. At high-current loads the voltage drop across the conductors may appreciably degrade the regulatiou at the point of load. This can be overcome to a great extent by the use of regulated supplies employing remote sensing circuits, which automatically compensate for the voltage drop on the load line conductors.

To determine the maximum current-carrying capacity for a given wire size, place a straightedge from the wire size on scale 2 to point A on scale 3 (the maximum permissible voltage drop in millivolts). The maximum operating current is then read on scale 1.

The voltage drop per foot for a known wire size and load current is determined by placing the straightedge from the current on scale 1 to the wire size on scale 2, and the voltage drop in millivolts per foot is read on scale 3.

The wire size required for a known load current and maximum tolerable voltage drop across the conductors is found by determining the maximum tolerable drop in millivolts per foot of conductor (sum of both conductors). Lay the straightedge from this value read on scale 3 to the known current on scale 1. Read the required wire size on scale 2. The figures thus obtained from the nomograph are based on a minimum of 500 circular mils per ampere. Higher current drains can be used with high-temperature conductor insulations.

24.88 What is the procedure for rewinding a relay coil for a different



Lamp No.	Volts	Amps	Bead Color	Base	Bulb Type	Fig. No.
PR2	2.4	0.50	Blue	Flange	B-3½	A
PR3	3.6	0.50	Green	Flange	B-3½	A
PR4	2.3	0.27	Yellow	Flange	B-3½	A
PR6	2.5	0.30	Brown	Flange	B-3½	A
PR12	5.95	0.50	White	Flange	B-3½	A
12	6.3	0.15	— — —	2-Pin	G-3½	H
13	3.8	0.30	Green	Screw	G-3½	B
14	2.5	0.30	Blue	Screw	G-3½	B
40	6.3	0.15	Brown	Screw	T-3¼	C
41	2.5	0.50	White	Screw	T-3¼	C
42	3.2	0.35*	Green	Screw	T-3¼	C
43	2.5	0.50	White	Bayonet	T-3¼	D
44	6.3	0.25	Blue	Bayonet	T-3¼	D
45	3.2	0.35†	Green†	Bayonet	T-3¼	D
46	6.3	0.25	Blue	Screw	T-3¼‡	C
47	6.3	0.15	Brown	Bayonet	T-3¼	D
48	2.0	0.06	Pink	Screw	T-3¼	C
49	2.0	0.06	Pink	Bayonet	T-3¼	D
50	6.3	0.20	White	Screw	G-3½	B
51	6.3	0.20	White	Bayonet	G-3½	E
55	6.3	0.40	White	Bayonet	G-4½	F
57	14.0	0.24	White	Bayonet	G-4½	F
112	1.1	0.22	Pink	Screw	TL-3	G
222	2.2	0.25	White	Screw	TL-3	G
233	2.3	0.27	Purple	Screw	G-3½	B
291	2.9	0.17	White	Screw	T-3¼	C
292	2.9	0.17	White	Screw	T-3¼	C
1490	3.2	0.16	White	Bayonet	T-3¼	D
1891	14.0	0.23	Pink	Bayonet	T-3¼	D
1892	14.0	0.12	White	Screw	T-3¼	C

* Some brands are .60 amp.
† Some brands are .60 amp and white bead.
‡ Frosted.

Fig. 24-86. Identification data for miniature lamps (pilot lights).

voltage?—As this is a rather involved operation, the data may be arrived at more easily by the use of the nomograph in Fig. 24-88. The new wire size is determined by laying a straightedge across the points where the present voltage and the desired voltage appear on the first two columns, and noting the point where the straightedge intersects the ratio column. The straightedge is then laid across this point on the ratio column and the present wire gauge column. Read the wire gauge needed for the desired voltage from the fifth column. Directly opposite this point the wire diameter in inches is also given.

For example: What gauge wire is required to rewind for 24-volt operation a relay coil wound with number 23 wire designed for 12-volt operation? Lay a straightedge from 12 volts in the first column to 24 volts in the second column. Note the point where the straightedge crosses the third column (ratio), which in this instance is 0.5. Connect this ratio with the present wire size (23) in the fourth column, and read the new wire size in the fifth column; this is 26 gauge (AWG), or 0.016 inch in diameter.

24.89 Describe a turret socket.—This socket is a one-hole mounting socket with a tubular piece running downward from the center of the socket, with terminals for attaching circuit components. Such sockets are available in many different designs and are used for compact electronic vacuum-tube construction.

24.90 Describe a sealed lead-acid battery.—The lead-acid storage battery was invented by Gaston Planté in 1860 and is one of the most widely used forms of battery power. The principal drawback to this type of battery has been the liquid electrolyte and the fumes given off when charging and discharging. However, due to new developments in sealing and venting, the

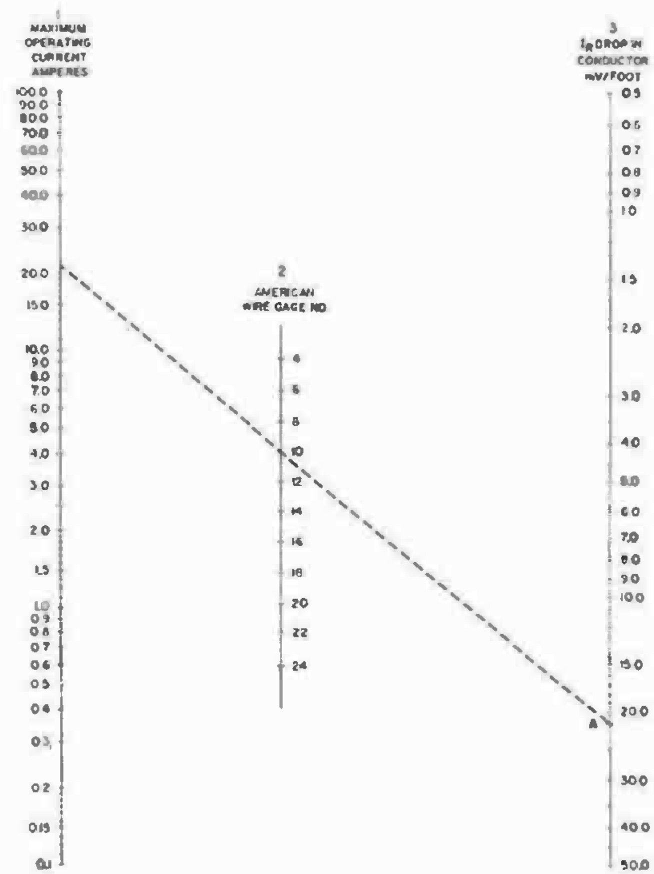


Fig. 24-87. Nomograph for determining the wire size, current-carrying capacity, and voltage drop for regulated power supply load conductors. (Courtesy, Kepco Inc.)

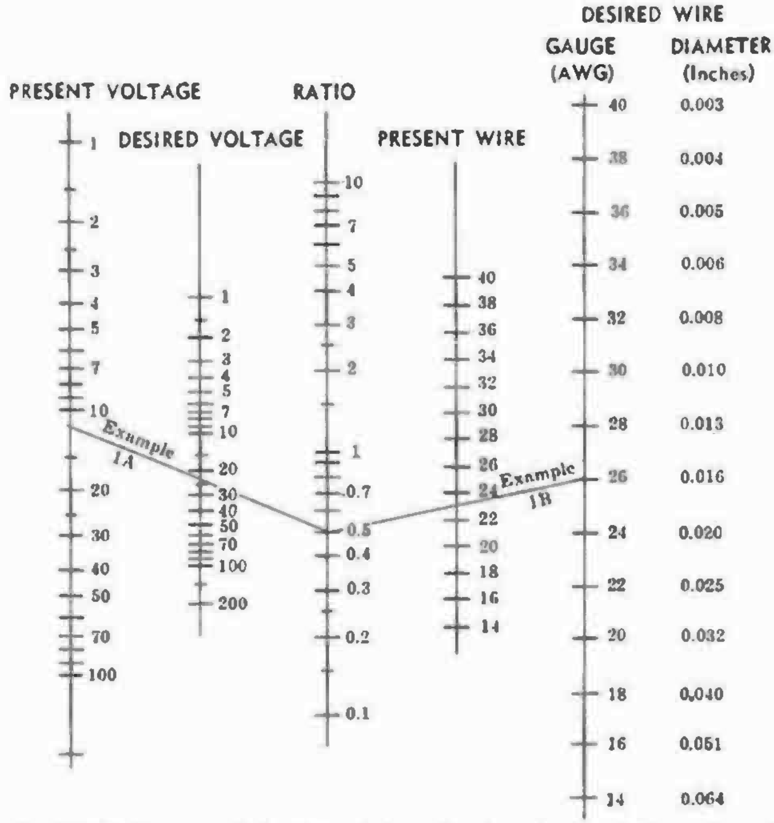


Fig. 24-88. Nomograph for determining wire size when rewinding relay coils.

sealed lead-acid battery may now take its place with other rechargeable batteries such as the nickel-cadmium battery. Lead-acid batteries are now available with voltages ranging from 2 to 8 volts and in 1- to 8-amperes capacities. Since small amounts of gas may be generated in any battery during the charge or discharge cycle, lead-acid batteries are vented so that the gas escapes, but not the electrolyte.

To prevent electrolyte movement in the battery, the electrolyte may be immobilized by the use of a gelling agent. A second method stores the electrolyte in highly porous separators, while a third method makes use of calcium-lead grids. By the use of such construction, the loss of water is minimized. Since water cannot be supplied to sealed lead-acid cells, they are designed to minimize the loss of water.

Lead-acid batteries have several advantages over other types of rechargeable batteries, one advantage being the discharge rate, as shown in Fig. 24-90.

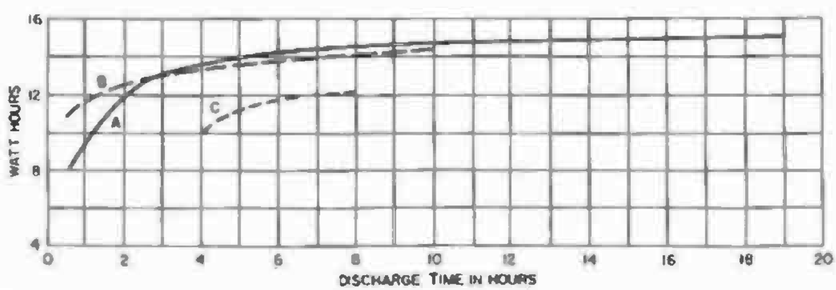


Fig. 24-90. Time variations taken to totally discharge comparable lead-acid, nickel-cadmium and alkaline Mn-Zn batteries. A lead-acid, B nickel-cadmium, C alkaline manganese-zinc.

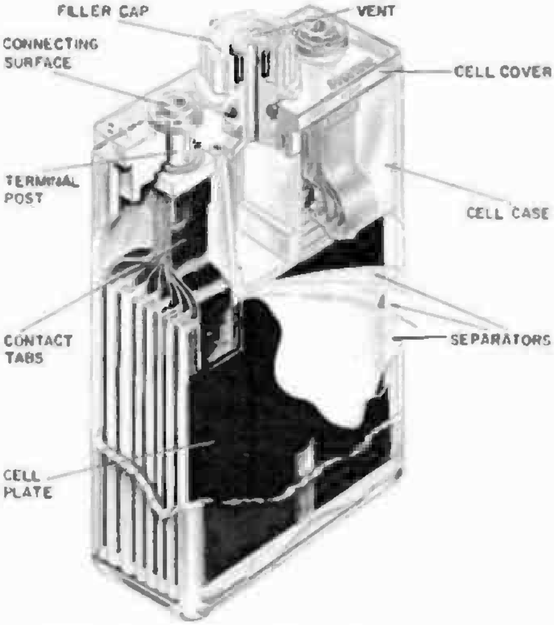


Fig. 24-91A. Interior view of Sonotone rechargeable nickel-cadmium battery.

24.91 Describe a nickel-cadmium battery.—Nickel-cadmium batteries are used extensively in portable electronic equipment and particularly in portable recording equipment. Fig. 24-91A is a cutaway view of a sinter-plate nickel-cadmium battery, manufactured by the Sonotone Corporation. A cell consists of sintered positive and negative plates, separators, safety vent cap, electrolyte, and plastic container.

The plates are made by firing a microfine nickel powder at high temperature until the particles weld or sinter together to form a porous structure. The porous structure is sintered on a fine-mesh nickel screen. These plates are then put through electrochemical processes designed to deposit active nickel and cadmium oxides in the fine pores of the plate. The cells are then assembled and connected to the terminals mounted in a polystyrene or nylon case.

The sealed construction eliminates the need to add water or electrolyte and under certain conditions, the cell will operate on overcharge for an indefinite period. A typical discharge curve for a Model 20L420 cell, rated at 25 ampere-hours and weighing approximately 2 pounds, is given in Fig. 24-91B.

Batteries of this type may be charged using either a constant-current or constant-voltage charger. The interior of a

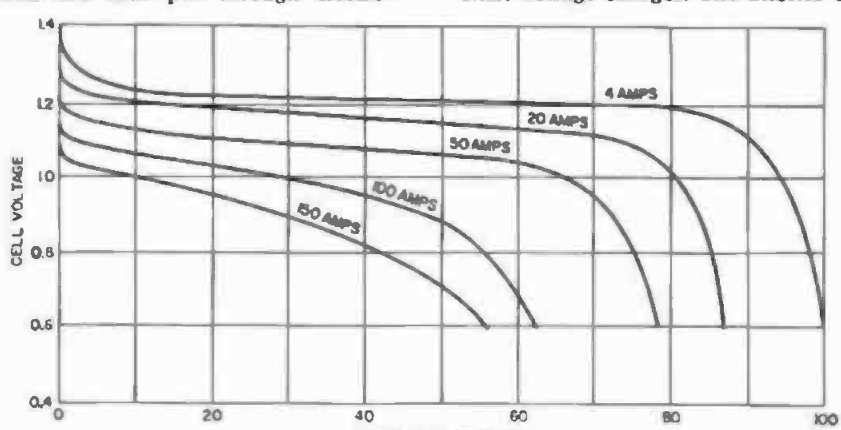


Fig. 24-91B. Discharge characteristic for Sonotone type 20L420 nickel-cadmium battery, rated 25 ampere-hours.

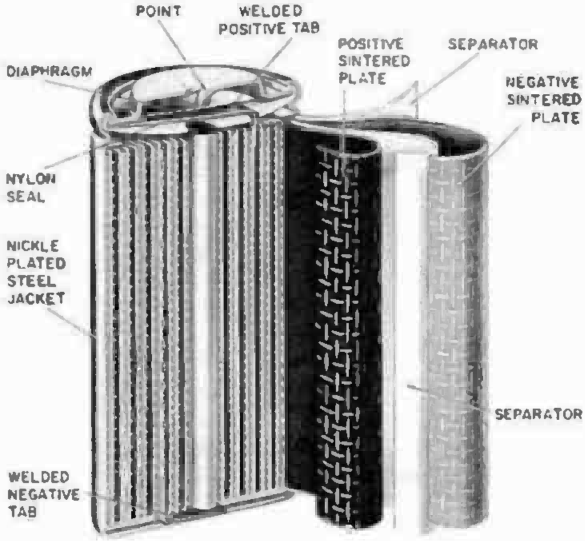


Fig. 24-91C. Interior of a Sonotone tubular nickel-cadmium rechargeable battery.

tubular-type nickel-cadmium cell is shown in Fig. 24-91C.

24.92 What is a bias cell?—It is a very small battery designed for connection in the grid return of a vacuum tube for supplying bias voltage. As a rule such cells are only used in the first stage of a class-A amplifier. This method of biasing is now obsolete.

24.93 Describe an Edison battery.—It is an alkaline storage battery using potassium hydroxide as an electrolyte, powdered iron and mercury for the negative plates, and flaked iron and mercury for the positive plates. This battery has exceptionally long life, and will stand heavy overloading and standing idle without requiring charging.

24.94 Describe the construction of a zinc-carbon (LeClanche) cell.—A cross-sectional view of a typical zinc-carbon (LeClanche) cell is given in Fig. 24-94, with its principal components called out. Basically the cell consists of a zinc anode, manganese-dioxide cathode, and electrolyte solution of ammonium chloride, zinc chloride, and mercury chloride in water. The nominal voltage is 1.5 volts. This type of cell is quite inefficient at heavy loads, and its capacity depends considerably on the duty cycle. Less power is available when it is used without a rest period. Maximum power is produced when it is given frequent rest periods, as the voltage drops continuously under load.

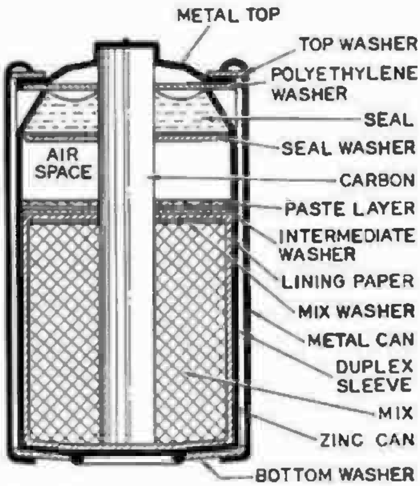


Fig. 24-94. Interior construction of a zinc-carbon (LeClanche) cell.

Shelf life is limited by the drying out of the electrolyte. A typical discharge curve is given in Fig. 24-95.

Zinc-carbon cells may be recharged for a limited number of cycles. The following information is extracted from *National Bureau of Standards Circular 965*:

> The cell voltage for recharge must not be less than 1 volt and should be recharged within a short time after removing from service. The ampere-hours of charge should be within 120 to 180 percent of the

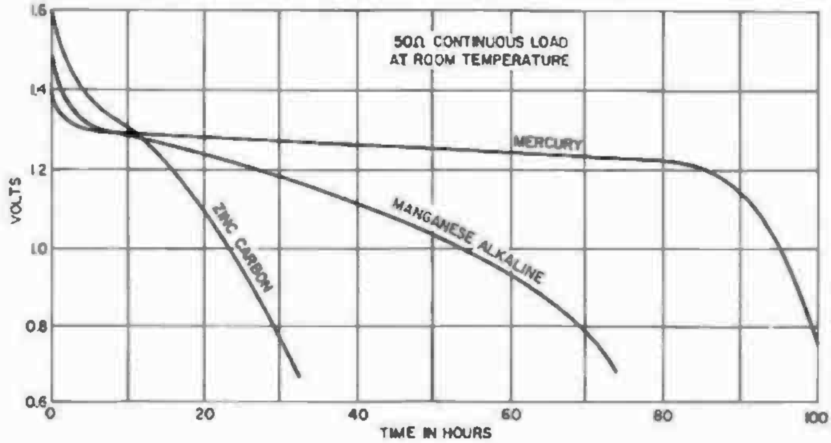


Fig. 24-95. Typical discharge curves for three different type penlight cells discharged continuously into a 50-ohm load.

discharge rate. The charging rate is to be low enough to distribute the recharge over 12 to 16 hours. Cells must be put into service soon after recharging as the shelf life is poor.

24.95 ***Give the discharge rate of a zinc-carbon cell as compared to a mercury battery and a manganese-alkaline battery.***—Typical discharge curves for the three batteries are given in Fig. 24-95.

24.96 ***What is a mercury dry cell?*** —The mercury cell using a zinc-mercury oxide alkaline system was invented by Samuel Ruben during World War II. There are two kinds of mercury cells: one with a voltage of 1.35 volts,

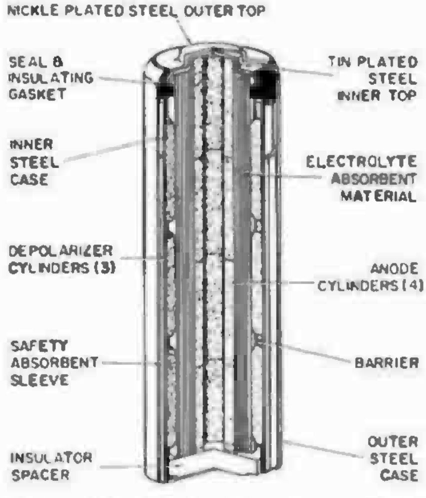


Fig. 24-96A. Interior view of a vertical-type mercury battery.

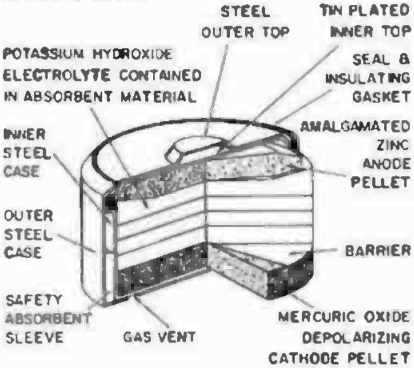


Fig. 24-96B. Interior view of a pellet-type mercury battery.

and one with 1.4 volts. Both have pure zinc anodes amalgamated with mercury, and an electrolyte of potassium hydroxide solution with some zinc oxide. The difference between the two cells lies in the cathode material.

The 1.35-volt cell has a pure mercuric-oxide cathode. On discharge its

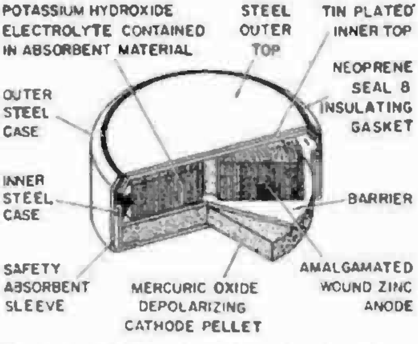


Fig. 24-96C. Interior view of a low-temperature type mercury battery.

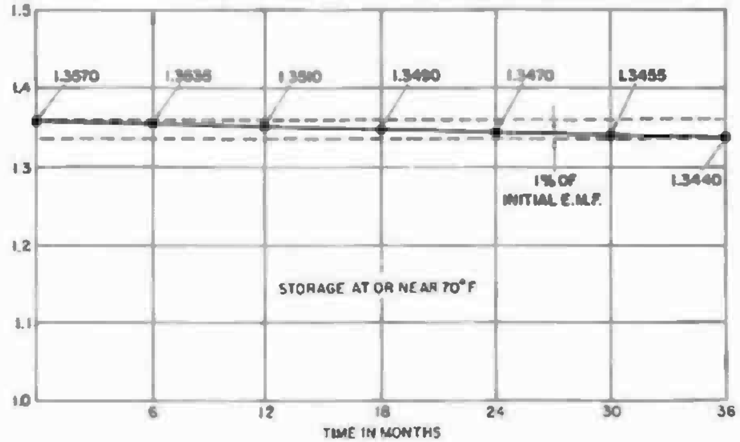


Fig. 24-96D. Stability curve for a single-cell mercury battery. The drop over a period of three years is 13 millivolts.

voltage drops only slightly until close to the end of the cell life, when it then drops rapidly. The 1.4-volt cell has a cathode of mercuric oxide and manganese dioxide. On discharge, its voltage is not quite as well regulated as the 1.35-volt cell, but it is considerably better than the manganese-alkaline or zinc-carbon cell.

Mercury cells have excellent storage stability. A typical cell will indicate a voltage of 1.3569 volts, with a cell-to-cell variation of only 150 microvolts. Temperature variation is 42 μV per degree ranging from minus 70 to plus 70 degrees Fahrenheit, with a slight increase of voltage with temperature. The internal resistance is approximately 0.75 ohms. Voltage loss during storage is about 360 μV per month. Therefore, a single cell can be used as a stable source of voltage, which is considered to be 1.3544 volts, plus or minus 0.17 percent. The voltage is defined under a load condition of 5 percent of the maximum current capacity of the cell. Normal shelf life is on the order of 3 years. (See Question 24.95.)

Recharging of mercury cells is *not recommended because of the danger of explosion*. Interior views of the battery construction are given in Figs. 24-96A, B, and C. A typical discharge curve appears in Fig. 24-95, and the stability characteristics for a single cell over a period of 36 months is shown in Fig. 24-96D. The drop in voltage over this period is 13 millivolts.

Fig. 24-96E shows a voltage-reference battery, manufactured by Mallory Battery Co., and designed for laboratory use as a reference voltage. The stability is plus or minus 0.5 percent from the rated voltage. Voltage stability is maintained for 3 or more years. For short periods it will provide accurately one part in a million as determined by a potentiometer calibration compared with a primary standard. The voltage increments are in steps of 1.35 volts up to 10.8 volts.

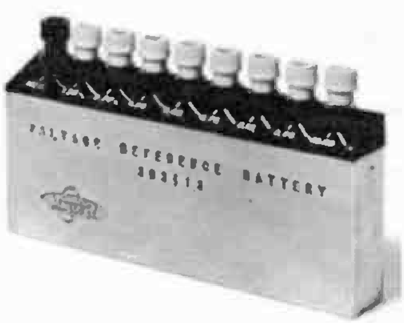


Fig. 24-96E. Mallory voltage reference battery. Voltage will remain within 1 percent over a period of 3 years.

24.97 Describe an alkaline-manganese battery.—The alkaline manganese battery is gaining considerable importance in the electronic field, as it is a primary battery and is rechargeable. A cross-sectional view of a typical alkaline-manganese battery is given in Fig. 24-97A, with the principal components called out. This cell uses a cylindrical depolarizer in contact with a cell container of nickel-plated steel. Because of the passivity of steel in

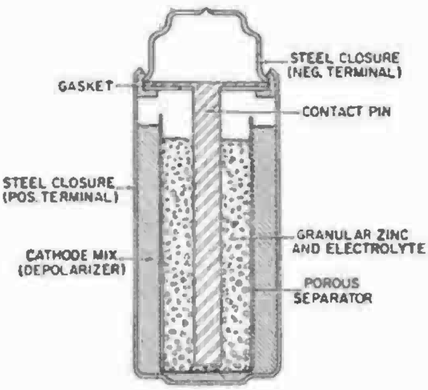


Fig. 24-97A. Cross-sectional view of an alkaline-manganese battery cell. Through external jacketing and terminal connections, cells can be made to appear as a conventionally polarized battery. The center button is positive.

alkaline electrolytes, there is no chemical reaction between the depolarizer and the steel, thus permitting the latter to be both a current collector and container. The depolarizer surrounds a cylindrical, granular, zinc anode, with the two electrochemical components being separated by porous material.

It will be observed that the polarity of this cell is reversed from the conventional zinc-carbon cell, in which the can is negative. However, because of packaging, the outward appearance is similar to the zinc-carbon cell, with the same terminal arrangement. Although this cell has an open-circuit voltage of approximately 1.5 volts, it discharges at a lower voltage than the zinc-carbon cell; also, the discharge voltage decreases steadily but more slowly. Alkaline-manganese batteries have 50 to 100 percent more capacity than their zinc-carbon counterparts. Zinc-carbon cells yield most of their energy above 1.25 volts and are virtually exhausted at 1 volt, while the alkaline cell yields most of its energy below 1.25 volts with a considerable portion released at less than 1 volt.

If the discharge rate is limited to 40 percent of the nominal capacity of the cell, and recharge is carried out over a period of 10 to 20 hours, alkaline-manganese cells can be cycled 50 to 150 times. A typical discharge curve is shown in Fig. 24-97B. (See Question 24.95.)

24.98 ***When setting up a laboratory bench, what facilities must be considered?***—Benches designed for electronic laboratories should be built not only to provide a convenient place to construct equipment, but also to provide storage and space for test equipment. An end view of a typical bench is shown in Fig. 24-98. At the back of the table is a shelf running the full length of the bench to support test equipment. Convenience outlets in metal-moulding type of conduit every two feet or so provide power for the test equipment. If possible, the power for the test equipment and equipment under test should be supplied from a sine-wave-type voltage regulator such as is discussed in Question 8.100.

A shelf approximately 12 inches wide runs the full length of the bench under the table at the rear. The top of the bench is covered with ¼-inch tempered masonite. Several benches may be constructed and then set in a semicircle or placed in a "U." This latter construction is used quite frequently in laboratories where a number of individual benches are required. The sides are

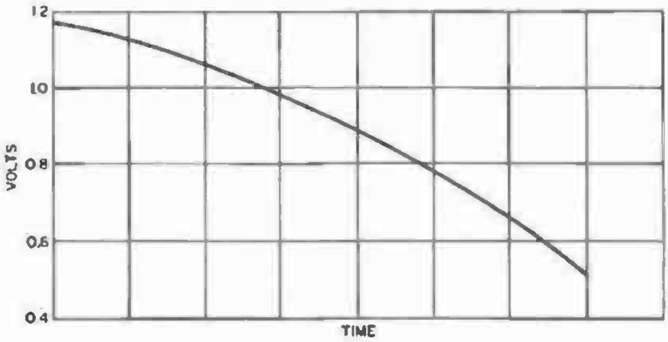


Fig. 24-97B. Discharge characteristics of an alkaline-manganese cell, on an arbitrary time scale. The cell discharges at lower voltages as compared to a zinc-carbon cell.

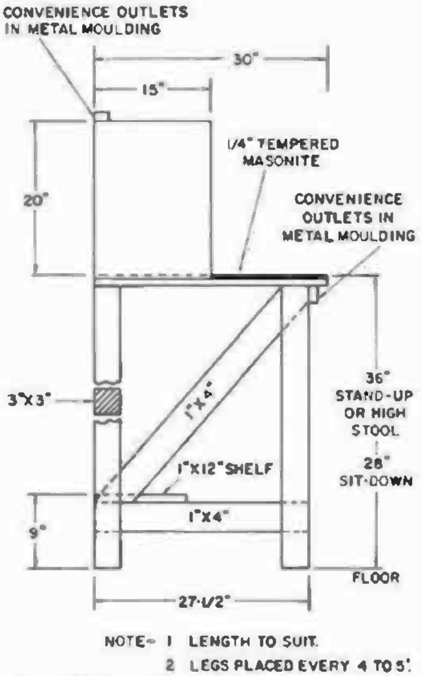


Fig. 24-98. Dimensions for a typical electronic laboratory bench.

backed up with a 6-foot plywood wall to afford a certain amount of privacy.

A good ground and overhead lighting complete the installation. If fluorescent lights are used, the fixtures must be grounded, and if possible the starters and reactors removed to a remote position, to reduce the effects of stray magnetic fields.

24.99 ***Describe the procedure for determining the performance of equipment cabinet cooling fans.***—First the designer must know the amount of dissipated heat to be removed and the maximum temperature rise desired for the coolant (air). The equation relating to the heat removed equals the weight of the coolant times the specific heat of the coolant times the temperature rise of the coolant. If this equation is applied to air at room conditions on a rate basis, it may be computed:

$$\text{cfm} = \frac{\text{Btu/hr}}{1.08\ (\Delta T_F)} = \frac{3170\ \text{kW}}{(\Delta T_F)} = \frac{\text{Btu/hr}}{1.94\ (\Delta T_c)} = \frac{1760\ \text{kW}}{(\Delta T_c)}$$

where,

Btu is the heat to be removed,
T is the temperature in Fahrenheit or centigrade,
cfm is the cubic feet of air required.

In order to ensure adequate cooling, about 25 percent of excess air should be planned. Common-design temperature rise is 10 degrees centigrade. The performance limits for cabinet cooling fans are given in Fig. 24-99.

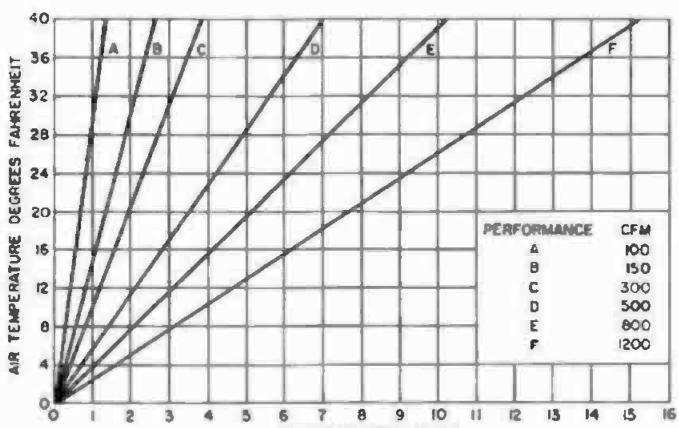


Fig. 24-99. Graph for determining the performance limits of cabinet cooling fans.

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Section 25

General Information, Charts, and Tables

Information of general interest and significance to the audio engineer or technician which did not rightly belong to any of the foregoing sections has been delegated to this section, along with useful charts, tables, and formulas to serve as a ready reference. Data on batteries, wire, lamps, and relays are given attention. A list of technical societies, international commissions, standards committees, and organizations allied to the industry and their official titles and abbreviations are also listed.

25.1 ***What are typical insulation measurements for two-conductor shielded pairs used in sound installations?***—For a No. 20 gauge, the leakage from the conductors to the surrounding shield will measure between 40 and 60 megohms.

25.2 ***What is pulse modulation?***—A system of modulation used in radar and special systems of communication.

25.3 ***What is phase modulation?***—A system of modulation in which the audio signal is used to shift the phase of the carrier frequency in proportion to the instantaneous amplitude of the audio-frequency signal. The carrier frequency changes, but the power remains constant. It is used in both amplitude- and frequency-modulation systems.

The modulated carrier frequency for a phase-modulated radio transmitter has an appearance similar to that of the frequency-modulated carrier shown in Figs. 25-5A to D. Signals modulated by a phase-modulated transmitter will be received on an fm receiver the same as a frequency-modulated signal.

25.4 ***What is amplitude modulation and how is it accomplished?***—Amplitude modulation is a system for the transmission of intelligence via radio. It employs a high-frequency wave of constant amplitude and phase called a *carrier frequency* and a lower frequency called a *modulation frequency*. The modulating frequency is superimposed on the carrier frequency and modulates the carrier so that it corresponds with the changes of amplitude of the modulating frequency.

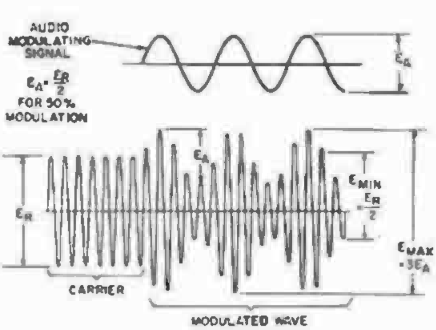


Fig. 25-4A. A radio-frequency carrier, amplitude modulated 50%.

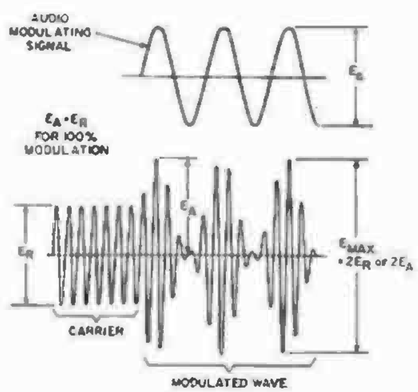


Fig. 25-4B. A radio-frequency carrier, amplitude modulated 100%.

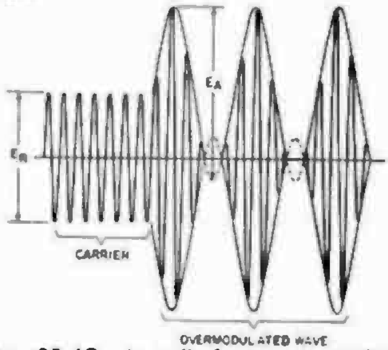


Fig. 25-4C. A radio-frequency carrier, amplitude modulated over 100%. Note loss of negative peak.

For radio broadcasting, the carrier frequency lies between 535 and 1605 kHz, with modulating frequencies between 40 and 5000 Hz. Carrier frequencies modulated 50 and 100 percent are shown in Figs. 25-4A and B, and an overmodulated carrier is shown in Fig. 25-4C.

The percent of modulation may be expressed:

$$\% \text{ Modulation} = \frac{A - B}{A + B} \times 100$$

where,

A is the maximum and B the minimum peak-to-peak amplitude of the modulated carrier frequency.

If the percent of modulation is increased above 100 percent by making the modulating voltage greater than that of the carrier frequency (Fig. 25-4C), the carrier will be overmodulated, causing a distorted waveform to be transmitted, because the carrier frequency is cut off for negative swings of the modulating voltage. Therefore, amplitude-modulation systems must not be modnlated greater than 100 percent.

Although the preceding discussion deals entirely with sine waves and a radio-frequency carrier, the same priuciples apply to lower frequencies and complex waveforms.

25.5 What is frequency modulation and how is it accomplished?—Frequency modulation is a system for the transmissiou of intelligence, via radio, employing a radio-frequency carrier of constant amplitude which is varied in frequency as the amplitude of the modulating frequency changes. For radio broadcasting, the carrier frequency lies between 88 and 108 MHz, with the modulating frequencies between 50 and 15,000 Hz.

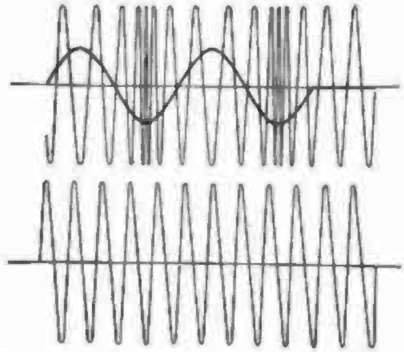


Fig. 25-5A. A radio-frequency carrier, frequency modulated with a low-frequency modulating voltage.

The percent of modulation of a frequency-modulated carrier cannot be determined as it can for an amplitude-modulated carrier because the amplitude of the carrier remains constant and the frequency is varied directly with the frequeucy of the modulating voltage. The limits of the frequency shift of the carrier, caused by the modulating voltage, are called the *frequency-deviation limits.*

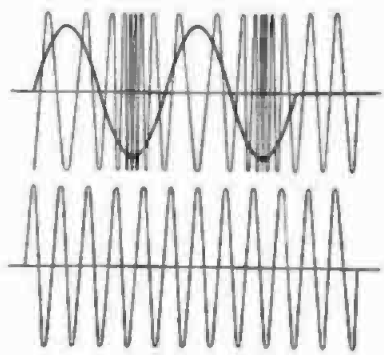


Fig. 25-5B. A radio-frequency carrier, frequency modulated with the same frequency as in Fig. 25-5A but with greater amplitude. Note the greater deviation of the carrier frequency.

Figs. 25-5A and B show a carrier frequency modulated by the same modulating voltage but for two different modulation factors (modulation index). Fig. 25-5C shows the same carrier modulated by voltages of different frequencies.

The higher the frequency of the modulating voltage, the greater number of times the carrier frequency will be varied within the deviation limits. The ratio of the maximum frequency deviation to the maximum frequency of the modulating voltage is called the *modulation factor*, or *modulation index*, and may be expressed:

$$\text{Modulation index} = \frac{\text{Max. frequency deviation of carrier}}{\text{Max. frequency of modulating voltage}}$$

Thus, an fm transmitter with a maximum deviation of 75 kHz is modulated 100 percent when the transmitter swings a full 75 kHz. At 37.5 kHz the transmitter is modulated 50 percent.

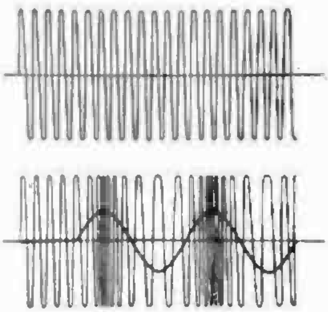


Fig. 25-5C. The same carrier frequency shown in Figs. 25-5A and B but using different modulating frequencies. Note the difference in the carrier deviation.

Although the foregoing discussion deals with high-freqnency voltages, the same principles apply to lower frequencies and waveforms of complex nature.

25.6 ***Describe a chopper.***—It is an electromechanical or electronic device used to interrupt a dc or low-frequency ac signal at regular intervals to permit amplification of the signal by an ac amplifier. Such devices are used in highly sensitive dc voltmeters. (See Question 22.97.)

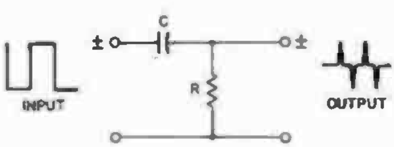


Fig. 25-7A. RC differentiating circuit.

25.7 ***What is a differentiation circuit?***—A circuit consisting of a resistor and a capacitor connected as shown in Fig. 25-7A. If a square wave is applied to the input, it is converted to positive and negative pulses. The pulse length is governed by the values of circuit elements C and R.

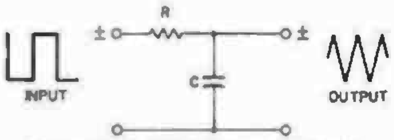


Fig. 25-7B. RC integrator circuit.

If the components are reversed as shown in Fig. 25-7B the circuit becomes an integrator. Now, if a square wave is applied to the input, the output signal is a sharply peaked sine wave. Circuits snch as these are often used when conducting square-wave tests on various types of equipment.

25.8 ***What are ferrite beads and how are they used?***—Ferrite beads consist wholly of powdered iron without a binder. They may be strung on leads carrying radio-frequency current for decoupling, without any dc or low-freqnency power loss. They may also be used for filament isolation or decoupling, as well as for suppressing parasitics in grid circuits, without altering the dc resistance. They are often nsed in oscilloscopes to reduce the effects of coupling at the high frequencies and to prevent oscillation. A cross-sectional drawing of a typical bead is shown in Fig. 25-8.

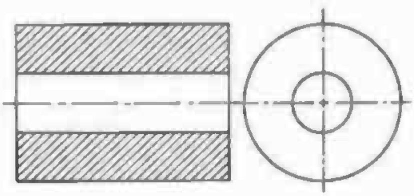


Fig. 25-8. Cross-sectional view of a typical ferrite bead.

25.9 ***What is the effective power of a circuit?***—The true power of the circuit.

25.10 ***What is an abscissa?*** — The horizontal plane of a graph.

25.11 ***What is an ordinate?***—The vertical plane of a graph.

25.12 ***What is an absorption circuit?***—A tuned circuit that dissipates energy taken from another circuit by electromagnetic coupling.

25.13 ***What is an alternating-current resistance?***—The impedance of a circuit at a given frequency.

25.14 ***What is an ampere-hour?***—The quantity of electricity flowing from a power source in 1 hour. A current of 2 amperes for a period of 1 hour is equivalent to 2 ampere-hours.

25.15 ***Define the angle of lag or lead.***—It is the number of electrical degrees by which an alternating current reaches its peak and zero values after, or before, the applied voltage reaches its peak or zero values.

25.16 ***What is an anticapacitance switch?***—A switch in which the current-carrying parts are separated to reduce the electrostatic capacitance existing between them. Some designs include shields between the various sections of the switch.

25.17 ***What is an aperiodic circuit?***—An untuned circuit not resonant at any one frequency.

25.18 ***What does the term "arithmetic mean" denote?***—One-half the sum of two values, or the average.

25.19 ***What is a circular mil?***—The unit of measurement of the cross-sectional area of a conductor. The area of a circle having a diameter of 0.001 inch.

25.20 ***What is a collector ring?***—A metal ring used in conjunction with a brush contact to obtain a connection between a rotating element and an external circuit, such as between the armature of a generator and the load.

25.21 ***What is a passive element?***—A resistor, capacitor, inductance, etc.

25.22 ***What is a cosine?***—In trigonometry, the sine of the complement of an angle.

25.23 ***What is a crest voltmeter?***—A peak-reading voltmeter.

25.24 ***What is a dead-beat instrument?***—A voltmeter, ammeter, or similar device with a highly damped movement. Damping is used to bring the movement to rest quickly.

25.25 ***Describe a readout tube.***—It is a special neon cold-cathode tube with ten independent cathodes, each formed in the shape of a number. Application

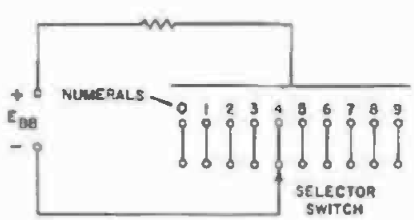


Fig. 25-25A. Basic circuit for a cold-cathode readout tube.

of a negative voltage to a selected cathode causes the gas around the cathode to ionize and glow. The visual effect is a bright glow around the

Fig. 25-25B. Interior view of cold-cathode readout tube Model NL-803 manufactured by National Electronics Inc.

cathode forming the number. Readout tubes are used in many different pieces of electronic equipment, such as digital voltmeters, counters, or where a visual readout is required.

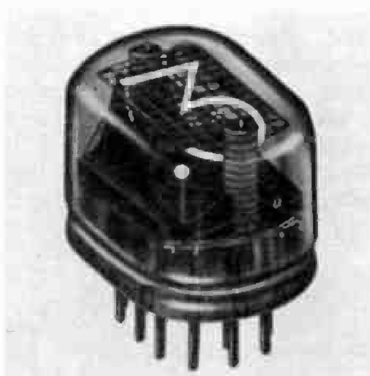


Fig. 25-25C. Interior view of flat cold-cathode readout tube Model NL-809.

For best viewing results, the tubes are mounted in an enclosure with a black interior, behind a red, amber, or polaroid filter plate. The filter improves the viewing in areas having a high ambient light level.

The basic circuitry for operating this type of tube is shown in Fig. 25-25A.

Two interior-construction views for cold-cathode readout tubes, manufactured by National Electronics, Inc., are pictured in Figs. 25-25B and C.

25.26 What is an acoustic blister?—A large dome-shaped piece of acoustic material sometimes placed on walls in broadcast studios to diffuse the sound. It varies in diameter from a few inches to about three feet. Such devices are used in conjunction with broken-up walls.

25.27 What are the advantages of a ceramic magnet?—Ceramic magnets are used in speakers and other devices because of the ease of molding them into a given shape. The ring type used in speaker construction is not always the ideal shape, particularly in the case of steel magnets, as it lowers their efficiency. The ceramic magnet is not affected in so marked a degree as is a material such as Alnico V. (See Question 20.23.)

25.28 What is an electrical degree?—It is one-360th part of an electrical cycle.

25.29 What is an electrolytic capacitor?—A capacitor of high capacitance in which an electrolyte and an electrode serve as plates. The dielectric is a film of gas formed by electrolysis. Such capacitors may be made using either a dry or a wet electrolyte.

25.30 If a new electrolytic capacitor measures less than its rated capacity, is it defective?—No. Electrolytic capacitors which have been on the shelf or out of service for some time may measure only 40 to 60 percent of their rated capacitance. After they have been in use at their normal operating voltage for a short period, the capacitance will return to its normal value or greater. Electrolytic capacitors should be reformed by operating them at their rated voltage before measuring the capacitance. (See Question 22.34.)

In general, electrolytic capacitors measuring 75 percent of their rated value should be replaced. In the use of bypass capacitors there is from the application standpoint usually no upper limit on the value. This is reasonably true for all filter capacitors, except for input capacitors used in power supplies. Here the limit is set by the permissible current through the rectifier element. New dry-electrolytic capacitors generally fall within the limits below.

Rated Voltage	*Percent Tolerance*
3 to 50 volts	—10 to +60
51 to 350 volts	—10 to +40
351 to 600 volts	—10 to +50

(See Question 21.59.)

25.31 What is the average leakage current per microfarad for an electrolytic capacitor?—About 0.2 to 2.5 mA per microfarad. Specification JAN-C-62 for electrolytic capacitors specifies that the leakage current shall not exceed 0.04 times the capacitance in microfarads plus 0.30.

25.32 Are electrolytic capacitors affected by frequency? — Electrolytic capacitors are quite satisfactory for frequencies below 500 kHz. However, at frequencies above 500 kHz, they lose their effectiveness quite rapidly. As an example, a 10-μF electrolytic capacitor at frequencies above 500 kHz will have an effective capacitance of only 0.5 μF.

It is customary, when electrolytic capacitors are used in circuits where the frequency may run from zero to several hundred thousand hertz, such as in a video amplifier, to connect additional paper and mica capacitors in parallel with the electrolytic capacitor. The electrolytic capacitor operates quite effectively at audio frequencies, the paper capacitor operates through the middle range of radio frequencies, and the mica capacitors operate at the extremely high frequencies.

25.33 What is the average power factor of an electrolytic capacitor?—About 5 to 10 percent. Although a high-quality capacitor may indicate a power factor higher than this after lying idle for a while, the power factor will drop after a short operating period of the capacitor at its rated operating voltage.

Electrolytic capacitors should be replaced if the power factor is 15 percent or greater. Wet-type electrolytic capacitors will show a power factor of 25 percent or greater.

25.34 What is fish paper?—A fiber-like insulating paper used in the construction of transformers, coils, and other electrical devices.

25.35 What is the geometric mean?—The square root of the product of two quantities. (See Question 7.34.)

25.36 What is the relationship of current to voltage in a capacitor?—The current leads the voltage by 90 electrical degrees.

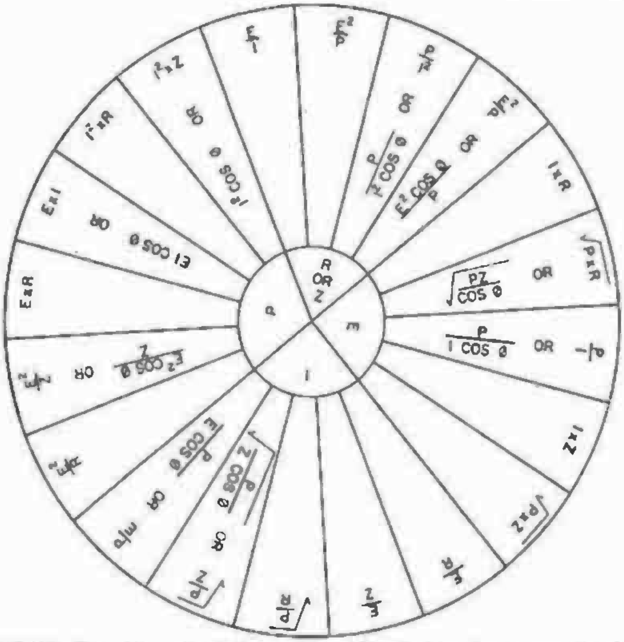


Fig. 25-38. Equation chart for solving Ohm's law for ac or dc circuits.

25.37 What is the relationship of current to voltage in an inductance?—The current lags the voltage by 90 degrees.

25.38 What are the equations used in the solution of ac and dc problems involving the use of Ohm's law?—A ready reference chart which may be used to select the proper equation for the solving of both ac and dc problems involving the use of Ohm's law is given in Fig. 25-38. The desired answer is selected from the inner circle and the proper equation is selected from one of the segments of the outer circle. In each case, two factors of the problem must be known.

25.39 What are the equations used in the solution of problems involving the decibel?—Fig. 25-39 provides a ready reference chart which may be used to select the proper equation for solving

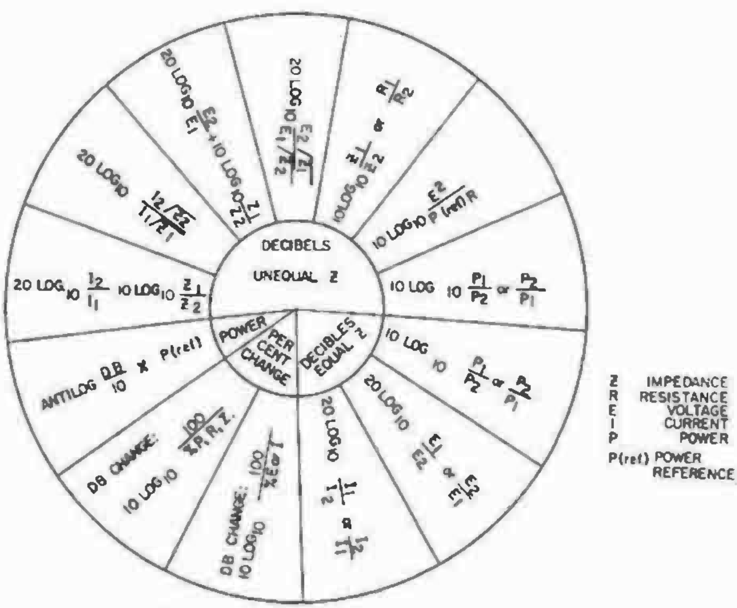


Fig. 25-39. Equations for problems involving decibels.

problems involving the use of decibels. The desired answer is selected from the inner circle and the equation from one of the segments of the outer circle.

25.40 ***Is it permissible to connect power supplies in series?***—Yes, power supplies may be connected in series like batteries to obtain a higher output voltage, and they may be a combination of regulated and nonregulated. Precaution must be taken to remove any grounds to prevent the shorting out of one supply.

The ripple voltage at the output of the combined voltage will be that of the power supply having the highest ripple voltage and in some instances it may be higher. The maximum current will be that of the supply with the lowest current rating. The voltage output will be the sum of the two voltages.

25.41 ***What is a phantom circuit?***—A telephone circuit consisting of two circuits from which a third circuit is taken by the use of repeat coils.

25.42 ***What type of wire is used for precision wirewound resistors?***—Manganin or Evanohm.

25.43 ***Is it permissible to charge storage batteries with unfiltered rectified alternating current?***—Yes.

25.44 ***How much power is required to modulate a radio transmitter of a given power to 100-percent modulation?***—The power output of the modulating amplifier must be at least 50 percent of the power of the stage being modulated.

25.45 ***What is the sensitivity of meter movements used in service voltmeters?***—The sensitivity varies with the ohms-per-volt requirements. The sensitivity of commonly used meters is given below.

Movement Sensitivity	*Ohms per Volt*	*Internal Res. of Meter*
50 μA	20,000	900/2K
100 μA	10,000	900/2K
200 μA	5000	300/1K
500 μA	2000	70/300
1.00 mA	1000	25/100
2.00 mA	500	27/30
5.00 mA	200	27/30
10.00 mA	100	1.5/7.0

The three latter movements are used on heavy-power test equipment and switchboard instruments.

25.46 ***What is litz wire?*** — A stranded wire consisting of several strands of very small wires transposed at intervals to reduce the loss at high frequencies. As an example, 32/28 litz wire is made up of 32 separate strands of No. 28 enameled wire. This wire is used for frequencies within and below broadcast band, as well as for loop antennas.

One of the principal objections to its use is the difficulty in soldering. If two or more of the strands are left unconnected, the radio-frequency resistance rises quite rapidly and the benefits of such wire are lost. The wire was originally developed in Germany many years ago by Litzendrabt and used extensively in wireless receivers for the reception of long-wave transmissions.

25.47 ***What is a negative-resistance circuit?***—A condition existing in a circuit when au increase of voltage causes a decrease of current.

25.48 ***What is parasitic oscillation?***—Oscillation set up in a circuit due to distributed capacitance and inductance. An unwanted oscillation.

25.49 ***What is a pulsating current?***—A current which changes its amplitude but not its direction.

25.50 ***What are the advantages of printed-wiring boards and how are they made?***—Chief advantages of printed-wiring boards are (1) lower production costs, and (2) laboratory circuits may be obtained in production.

Several factors are responsible for the lower production costs. Some are:

1. Greater simplicity in wiring since less wiring is required.
2. Wiring errors are eliminated because each board is identical.
3. Faster production because hand wiring is eliminated. Components may be installed by machines and the entire assembly soldered in one operation.
4. Adaptability of printed boards to the use of printed components.

Printed-wiring boards are produced in several different ways, but the end result is essentially the same—a phenolic circuit board with a wiring pattern on one or both sides.

One method of producing printed-wiring boards consists of bonding a thin layer of copper foil 0.0001 to 0.0004 inch thick to the phenolic material.

A chemical is then applied to the copper foil by a photographic process and the board is given an acid bath. The acid bath removes the copper foil except where the chemical was applied, leaving this area unharmed. The acid is then neutralized and the chemical removed from the copper foil, leaving the phenolic board with the desired printed wiring.

Another method of making printed-wiring boards consists of electrolytically depositing or plating the electrical circuit pattern on the phenolic board. This is accomplished by applying a conductive material to the base material. A chemical is then applied to the background area but not to the pattern area. The board is then copper-plated; however, because of the chemical applied over the conductive material in background area, only the diagram is plated. By means of a solvent, the conductive material and the chemical are then removed from the background area.

A third method of producing printed-wiring boards is embossing. A thin sheet of copper with an adhesive on one side is applied to the base material. The circuit pattern is embossed on the board, using photoetched metal dies. By heat and pressure, the dies force the copper foil into the base material. At the same time, the adhesive bonds the copper foil to the base material. This leaves the entire surface of the base material covered with copper foil, but only the circuit pattern is recessed into the base material. Excess copper foil is then stripped off mechanically, leaving the desired pattern embedded in base material. Circuit components such as resistors, capacitors, and coils are soldered directly to the printed circuitry.

Printed circuits are currently found in nearly all types of electronic equipment. (See Question 25.140.)

25.51 ***Define an electret.*** — This term is applied to a permanently polarized dielectric material and is the analog of a permanent magnet. Among the several types of electrets is that of a plastic surface or wax with a permanent positive charge and a permanent negative charge on its reverse surface, both surfaces having a stable existence. (See Question 4.126.)

25.52 ***What is a tank circuit?***—A tuned oscillatory circuit.

25.53 ***What is a vector?***—A straight line representing the magnitude and phase relationship of a quantity. A quantity which has both direction and magnitude.

25.54 ***What is a zero voltage point?***—The voltage of the earth.

25.55 ***Define the term "common-mode rejection."***—This term is also referred to as *in-phase rejection* and is a measure of how well a differential amplifier ignores a signal which appears simultaneously and in phase at both inputs. Usually and preferably stated as a voltage ratio but more often stated in decibel equivalent of the ratio at a specified frequency, such as 120 dB at 60 Hz, with 1000 ohms of impedance.

25.56 ***What is the equation for calculating the capacitance of a simple capacitor?***—Answer:

$$C = \frac{AK}{4.45D} \qquad \text{(in pF)}$$

where,

A is the area of the plate in square inches,
D is the spacing in inches,
K is the constant of the dielectric (insulation).

The dielectric constants for various materials are given in Question 25.57.

25.57 ***Define the term "dielectric constant."***—It is the property of a given material that determines the amount of electrostatic energy which may be stored in that material per unit volume for a given voltage. The value of the dielectric constant expresses the ratio of a capacitor in a vacuum to one using a given dielectric. The dielectric of air is 1 and is the reference unit employed for expressing dielectric constant.

The dielectric constants and breakdown voltages for commonly used materials are tabulated in the following table.

	K	*Dielectric Strength (volts per 0.001 volt inch)*
Air, at atmospheric pressure	1.0	80
Bakelite	5.0	500
Glass, common	4.2	200
Mica	6.0	2,000
Paraffin paper	3.5	1,200
Porcelain	5.5	750

If the dielectric constant of a capacitor is increased or decreased, the capacitor will increase or decrease in capacitance respectively.

25.58 ***What are Nagaoka's formulas?***—They are formulas used for the design of solenoid coils and may be obtained from Circular No. 74, published by the National Bureau of Standards.

25.59 ***What is the capacitive reactance of two capacitors connected in parallel?***—They are treated as resistors connected in parallel.

25.60 ***What is the capacitive reactance of two capacitors connected in series?***—They are treated as resistors connected in series.

25.61 ***If two ammeters are connected in series, what portion of the total current will be read by each meter?***—Each meter will read the total load current.

25.62 ***Can two ammeters of the same current range and internal resistance be connected in parallel to read double the current of one meter?***—Yes, if the internal resistance is the same for each meter. One meter acts as a shunt for the other; therefore, each meter reads one-half the current. Damage to the meters will result if one meter is removed from the circuit while there is a load current.

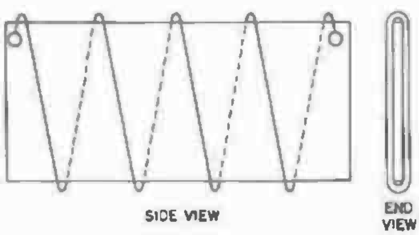


Fig. 25-63A. Construction of a fixed-card resistor.

25.63 ***What is a resistance card?***—A resistor wound on a piece of insulating material in the form of a thin card, as shown in Fig. 25-63A. After winding, the card is formed into a circle and fitted around a form such as is used in rheostats and potentiometers. The card may be tapered to permit various rates of change of resistance for variable controls. (See Fig. 25-63B.)

25.64 ***What is a nonlinear resistor?***—A resistor which does not obey the simple relationship of Ohm's law. Typical nonlinear resistors are: a vacuum tube operated under certain conditions,

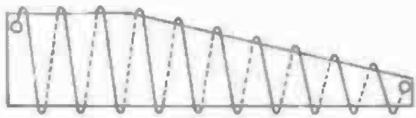


Fig. 25-63B. Construction of a tapered-card resistor.

iron-core inductors operated near saturation, thyrite resistors, thermistors, tungsten filament lamps, diode rectifiers, and special ceramic-dielectric capacitors using a bias voltage.

Nonlinear resistors are used in wave-shaping networks, oscillators, voltage dividers, frequency multipliers, amplitude limiters, and constant-output potentiometers. A typical plot of the characteristics of nonlinear resistors is shown in Fig. 25-64.

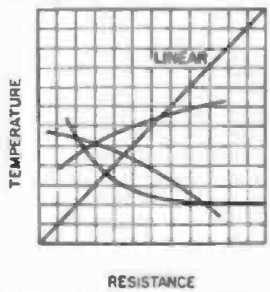


Fig. 25-64. Typical characteristics for nonlinear resistors.

25.65 ***What is a varistor?***—A trade name for a germanium or silicon diode whose resistance changes with changes in the applied voltage.

25.66 ***What is thyrite?***—A trade name for a material whose resistance decreases with an increase of applied voltage.

25.67 ***How are diodes used for the protection of meter movements?***—Two diodes connected in reverse polarity are shunted across the meter terminals, as shown in Fig. 25-67, to afford protection to the meter for either polarity. The meter movement is protected by the forward characteristic of one diode, while the other is inactive. Diodes used for meter protection must have very low forward current in the 0 to 300 millivolt range and very low reverse-current leakage. The dc resistance should be greater than 600,000 ohms below a voltage of 300 millivolts. The connection of a diode across the meter movement results in less than 0.5 percent of error, depending on the internal

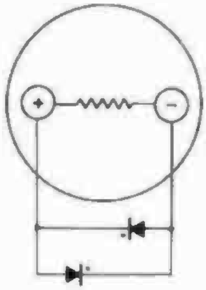


Fig. 25-67. Diodes used for protecting a meter movement.

resistance of the meter. Special diodes are manufactured for this purpose for meters with a given internal resistance.

25.68 ***What is a module unit?*** — A method of mechanically mounting electronic components in a stack of wafers to facilitate the automatic assembly of circuit components. It is a method developed by the National Bureau of Standards in conjunction with the Department of the Navy. A typical unit is shown in Fig. 25-68.

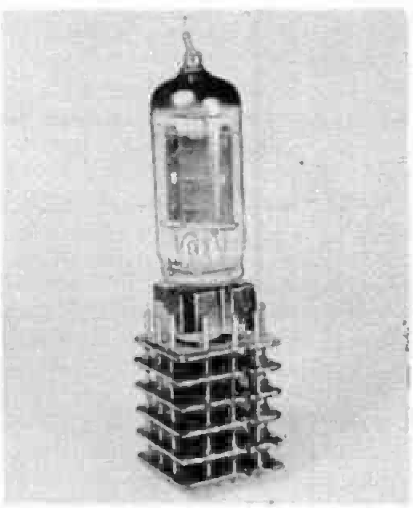


Fig. 25-68. A module unit with a miniature tube.

25.69 ***What is reflection loss?***—The transmission loss in a transmission line when a portion of the transmitted signal is reflected back toward the sending end because of improper terminating impedance.

25.70 ***What is propagation-time delay?***—The time required for a wave to travel from one point to another on a transmission line. Delay time varies with the type of line or cable.

25.71 ***How will the needle of a compass be deflected when placed in a magnetic field?***—The needle will align itself parallel with the lines of force. The north-seeking pole will always point to the north pole because it is, in reality, a south pole.

25.72 ***What is a side tone?*** — The noise reaching the earphone of a telephone subscriber's handset by means of a local path within the telephone circuit.

25.73 ***What is a side circuit?***—One of two circuits employed in a phantom circuit.

25.74 ***What is a padder capacitor?***—A variable capacitor connected in series with a fixed capacitor to obtain an exact value of capacitance. It is also used in tuned circuits to bring them to exact resonance.

25.75 ***What is Celsius temperature?***—The original centigrade temperature scale. It was named after Swedish astronomer Anders Celsius, who first described it.

25.76 ***What is fusetron?***—A trade name for a special type of plug fuse with an overload feature that will permit it to be overloaded, on starting, up to more than 500 percent, with blow times ranging from 0.5 to 25 seconds. Such fuses are used where a heavy starting current is present with a low running current.

25.77 ***What is an electrothermal recorder?***—A recorder in which the image is produced by thermal action on the recording medium in response to the received signals.

25.78 ***What is a facsimile transmission system?***—A system of radio or wire transmission by which illustrations, printed pages, maps, etc. are transmitted and received. The subject to be transmitted is scanned by a light beam along closely spaced parallel lines. The black-and-white copy is transmitted in the form of an electrical signal by the use of a photocell under the copy to be transmitted. At the receiving end, the impulses are translated into small marks on the receiving copy corresponding to the original. This is accomplished by tracing the receiving copy in closely spaced parallel lines similar to the transmitter. The two machines as synchronized at the two ends by an accurate local oscillator.

25.79 ***What is a tuning-fork oscillator?***—An oscillator circuit which em-

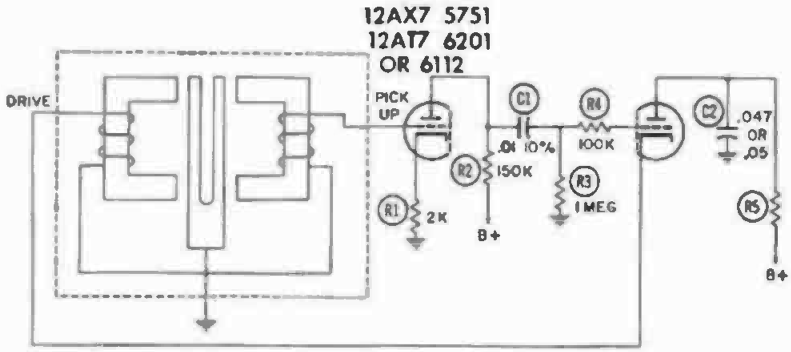


Fig. 25-79. Schematic diagram of a tuning-fork oscillator circuit.

ploys a tuning fork as the frequency-determining element. The elementary circuit for such an oscillator is shown in Fig. 25-79. As a rule, unless the output signal is filtered, the output contains a high percentage of harmonics.

25.80 What is a magnetostriction oscillator?—An oscillator circuit which employs an element in the oscillatory circuit that expands and contracts under the influence of a varying magnetic field. Certain metals, notably nickel and some of its alloys, undergo dimensional changes when subjected to the influence of a magnetic field. Since these changes are minute, in the order of one part per million, the phenomenon is usually observed or employed by placing a rod or tube in a solenoid carrying either an alternating or direct current.

25.81 What is a master oscillator?—An oscillator of stable characteristics used to establish a base frequency for several different parts of a system. By the use of a master oscillator, frequency discrepancies are eliminated.

25.82 What is scrambled speech?—Speech which has been rendered unintelligible for use in secret transmission by inverting the frequencies. These frequencies are converted back to the original form at the receiving end by an unscrambling circuit.

25.83 What is telemetering? — The remote indication of values, readings of meters, etc. by radio, using coded signals which are generally recorded on magnetic tape or film.

25.84 What is a two-channel multiplexed fm transmission?—A system of transmitting two-channel stereophonic program material using one rf carrier frequency. By multiplexing the signals, both the left- and right-hand sides of the audio pickup are transmitted independently and simultaneously. A single fm receiver having a multiplexing circuit separates two sides of the program material. The separated signals are then fed to left and right amplifier systems and reproduced using two speakers, similar to reproducing a stereophonic record or magnetic tape.

25.85 What does the term "admittance" mean?—The ease with which an alternating current flows in a circuit. Admittance is the reciprocal of impedance and is expressed in mhos.

25.86 What is an automatic gain control?—A self-acting compensating device which will maintain the output level of a transmission constant to within narrow limits for a wide variation of input signal level.

25.87 How are frequencies above the audio bands classified?

Very Low Frequency (vlf)	10 to 30 kHz
Low Frequency (lf)	30 to 300 kHz
Medium Frequency (mf)	300 to 3000 kHz
High Frequency (hf)	3 to 30 MHz
Very High Frequency (vhf)	30 to 300 MHz
Ultra High Frequency (uhf)	300 to 3000 MHz
Super High Frequency (shf)	3000 to 30,000 MHz
Extremely High Frequency (ehf)	30 to 300 GHz

25.88 What is a carrier frequency?—A wave which may be marked or modulated either by changing its amplitude, frequency, or phase, so that it may carry intelligence.

25.89 What is a coaxial cable?—A cable in which one conductor is accurately centered inside another. It is used for the transmission of television signals and other high frequencies.

25.90 *What is an electrolytic recorder?*—An electrochemical recorder in which a chemical change is caused by ionization in the recording medium, generally a chemically treated paper.

25.91 *What is an electrosensitive recorder?*—A recorder in which the image is produced by the passage of an electric current into the recording medium.

25.92 *What is an ink-mist recorder?*—A mechanical recorder used in facsimile transmissions in which particles of an ink mist are deposited directly onto the recording medium.

25.93 *What is a voder?*—An electronic device developed by H. Dudley of the Bell Telephone Laboratories for generating artificial speech. The tones are produced by a combination of mechanical keys similar to those of a piano keyboard.

25.94 *What is a vocoder?*—A device for generating artificial speech. The human voice is used to actuate the system rather than mechanical keys as used in the voder described in Question 25.93.

25.95 *What is an annulling network?*—A network of elements connected in parallel with a filter network and designed to cancel inductive or capacitive impedance at the extreme ends of the passband of the filter.

25.96 *What is an iterative impedance?*—The value of a terminating impedance, required at one pair of a four-terminal network, that will terminate the second pair of that network in a value equal to the first pair. The iterative impedance of a transmission line is the same as the characteristic impedance.

25.97 *What is meant by the term "unity coupling"?*—Unity coupling is achieved when the lines of force of one coil cut all of the windings of a second coil. The formula for calculating the coefficient of coupling is:

$$K^2 = \frac{M^2}{L_1 L_2}$$

where,

M is the mutual inductance between the coils,
L is the value of the self-inductance,
K is a constant.

As a rule, the value of K does not exceed unity coupling, or 1.

25.98 *What is the purpose of reversing the polarity of filament-type tubes operated from a direct-current source?*—This procedure is only used with high-power transmitting and rectifier tubes because the emission is not uniform over the entire surface of the filament. This causes certain areas of the filament to be weakened, in time. Reversing the polarity of the heater supply permits a more uniform operation of the filaments and prolongs their life. Also, the maximum life will be obtained when the filament voltage is maintained as near constant to the specified voltage as possible.

25.99 *Describe the characteristics of neon and argon glow lamps.*—Basically, a neon glow lamp is a member of the electric-discharge lamp family. It consists of two electrodes inserted in a glass envelope containing a rare gas. When a sufficiently high voltage is applied to the electrodes, current will flow between the electrodes by means of electrons and positively charged particles in the gas. This causes light to be produced at the negative terminal.

Glow lamps differ from conventional-type lamps in that they do not depend on a filament, and hence do not produce light by incandescence. Glow lamps are rugged in design with long life, and produce little heat in their operation. Such lamps are used for indicators and for voltage regulation.

The useful life of a glow lamp is a function of the lamp current and varies approximately as the inverse cube of the lamp current. Therefore, doubling the current through the lamp reduces its life to approximately one-eighth its normal life.

For some special applications the current may be increased as much as ten times its normal value without immediate injury; however, the characteristics will be changed, resulting in a rapid rise of starting and maintaining voltages, accompanied by a short life.

Glow lamps have a critical starting voltage. Below this critical voltage the lamp may be considered to be an open circuit. When the starting voltage is reached, the glow discharge begins and light is emitted. Once started, the lamp will continue to glow if the maintaining voltage is continued. On alternating current, the maintaining voltage is approximately the same as the starting

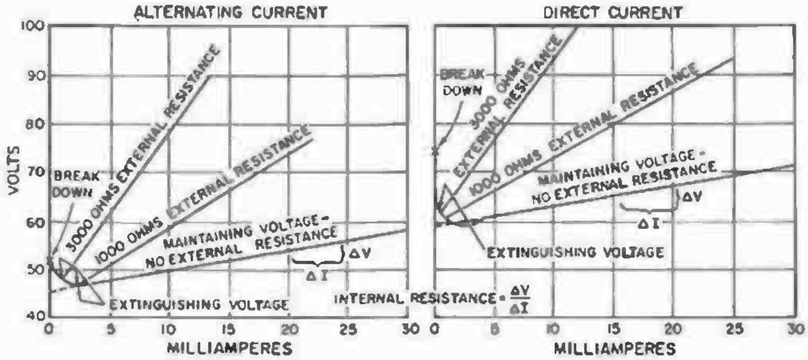


Fig. 25-99A. Runaway characteristics of a glow lamp.

voltage. On direct current, the discharge will be maintained at close to 15 volts below the breakdown voltage. Aging a glow lamp for about 100 hours by running it at its normal maintaining voltage will help to stabilize its characteristics. All glow lamps have a characteristic called the *runaway characteristic* (Fig. 25-99A); therefore, they must be operated in series with a current-limiting resistor. The screw-base lamp contains such a resistor; however, an external resistor must be used with bayonet-base lamps. Conventional carbon-type resistors are quite satisfactory for current-limiting use. By connecting a current-limiting resistor of the proper resistance in series with the lamp, these lamps may be operated from circuit voltages beyond their normal voltage ratings. Resistor values for different voltage ranges are given in Fig. 25-99B.

The glow lamp, when in a state of conduction, may be treated in an electrical circuit as the equivalent of a constant-arc drop in series with an internal resistor. Under the usual operating conditions, this arc drop may be considered equal to the minimum maintaining voltage. Current and resistance values may be calculated:

$$\text{Lamp I} = \frac{V_L - V_{MM}}{R_I + R_E}$$

where,

V_L is the line voltage,
V_{MM} is the minimum maintaining voltage,
R_I is the internal resistance,
R_E is the external resistance.

The efficiency of a glow tube is rather low, about 0.30 lumens per watt, and varies directly with the current. A quarter-watt lamp can be expected to produce six times the light of a $\frac{1}{25}$-watt type, and a 1-watt lamp four times that of a quarter-watt size. The light output may be increased by increasing the current through the lamp, but with a resulting decrease of life.

Ionization in complete darkness and deionization under any conditions may require a fraction of a second. The light output will follow the current in a linear manner up to frequencies of 15,000 Hz. Increasing the ambient light level within a limited range will lower the breakdown voltage because of the photoelectric emission of the electrodes.

Glow lamps are not noticeably affected by the ambient temperature below 300°F; however, above this temperature, gases are evolved from the bulb walls affecting the lamp operating characteristics.

Neon lamps may be used for voltage regulation in the same manner as a VR150 or similar type of tube, using a current-limiting resistor as shown in Fig. 25-99C. The regulated voltage is taken from across the lamp as shown. Neon lamps may also be used as leakage indicators for currents as low as 1 microampere and also as oscillators, as described in Question 22.55.

The spectral energy of such lamps is almost entirely confined to the yellow and red regions of the spectrum. The general appearance of a glow discharge tube is orange-red and may be used as an indicator without the necessity of using a red glass cover. Because of the spectral quality of the glow discharge, only two distinctive colors, red and amber, may be obtained by the use of filters.

The spectral emission characteristics for both the neon- and the argon-type glow lamps are shown in Fig. 25-99D.

The breakdown, maintaining and extinguishing voltage and operating cur-

Type	110-125V	220-300V	300-375V	375-450V	450-600V
AR-1	Included in Base	10,000	18,000	24,000	30,000
AR-3	Included in Base	68,000	91,000	150,000	160,000
AR-4	15,000	82,000	100,000	160,000	180,000
NE-2	200,000	750,000	1,000,000	1,200,000	1,600,000
NE-2A	200,000	750,000	1,000,000	1,200,000	1,600,000
NE-2D	100,000	----	----	----	----
NE-2E	100,000	----	----	----	----
NE-2H	30,000	----	----	----	----
NE-2J	30,000	----	----	----	----
NE-7	30,000	----	----	----	----
NE-17	30,000	110,000	150,000	180,000	240,000
NE-21	30,000	----	----	----	----
NE-30	Included in Base	10,000	20,000	24,000	36,000
NE-32	7,500	18,000	27,000	33,000	43,000
NE-34	Included in Base	9,100	13,000	16,000	22,000
NE-40	Included in base	6,200	8,200	11,000	16,000
NE-45	Included in Base	82,000	120,000	150,000	200,000
NE-48	30,000	110,000	150,000	180,000	240,000
NE-51	200,000	750,000	1,000,000	1,200,000	1,600,000
NE-51H	47,000	----	----	----	----
NE-54	30,000	----	----	----	----
NE-56	Included in Base	----	----	----	----
NE-57	Included in base	82,000	120,000	150,000	200,000
NE-58	Included in Base	----	----	----	----
NE-66	3,600	----	----	----	----
NE-79	7,500	----	----	----	----

Fig. 25-99B. Current-limiting resistor values used with glow-discharge tubes.

rent for both neon and argon lamps, may be measured using the circuit of Fig. 25-99E. When breakdown voltage is measured, the lamp should be operated for about 1 minute at its normal rated current within 1 hour of making the measurement. If it is to be measured in total darkness, the lamp should be inoperative for at least 24 hours before measurement.

The extinguishing voltage is dependent on the impedance in series with the lamp and is the voltage across the lamp at which an abrupt decrease in current between the operating electrodes occurs; it is accompanied by the disappearance of the negative glow.

Using the circuit shown, a 60-Hz voltage with a total harmonic distortion of not more than 3 percent is applied to the potentiometer and then to the lamp. The ripple voltage from the dc power supply must not exceed 0.10 percent.

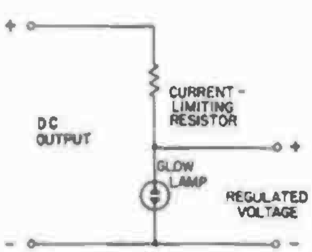


Fig. 25-99C. A voltage-regulation circuit using a glow lamp and a current-limiting resistor.

When ac measurements are made, a *thermocouple* meter should be used for measuring the current, with the series resistor (R.) selected for a value suitable for the lamp under test. The voltmeter across the lamp must have an input resistance of at least 10 megohms or more.

25.100 ***What is an argon glow lamp?***—Argon glow lamps are similar in construction and operation to the neon glow lamps. Argon radiates principally in the blue-violet and ultraviolet region of the spectrum. The visible purple color offers the possibility of their use as a second color to the neon glow lamp. The predominant radiation is in the near ultraviolet or black-light region. The initial black-light output is approximately 3.5 fluorens per watt. Glow lamps using argon gas may be used in timing devices for high-speed photography, as their violet and near-violet radiation occurs in a region where photographic materials have their highest degree of sensitivity. The values of current-limiting resistance required for an argon-type glow lamp are the same as

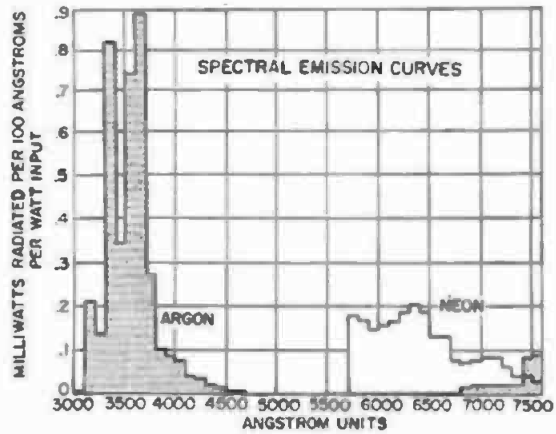


Fig. 25-99D. Spectral emission curves for glow-discharge tubes.

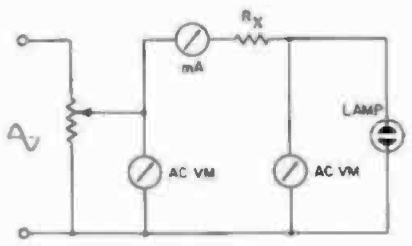


Fig. 25-99E. Circuit for testing gas-filled glow lamps.

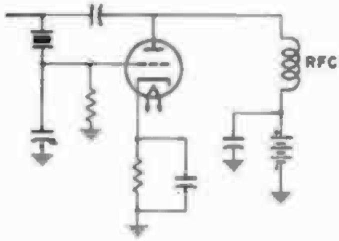


Fig. 25-102. A typical quartz crystal oscillator circuit.

for the neon-type lamp, as shown in Fig. 25-99B.

25.101 How are the specific gravity readings of a storage battery interpreted?

1.275 to 1.300	A high state of charge.
1.250	A medium state of charge.
1.200	A low state of charge.

If a storage battery is left in a discharged condition for any length of time, the plates may be damaged due to sulfation. (See Question 24.73.)

25.102 What causes a quartz crystal to oscillate?—Quartz crystals have two modes of vibration—they can bulge in and out in a direction perpendicular to their long parallel surfaces, or they can expand and contract so that the short parallel surfaces bulge in and out. In the so-called thickness mode of vibration, pressure waves travel from one long surface to the opposite surface and are reflected back again. For a given thickness of the crystal slab, the reflected waves are in phase with the direct waves and reinforce each other. This causes standing waves on the surface of the crystal in the long direction. Under these conditions, it is said the crystal is at resonance.

The natural frequency of resonance will occur, for a particular thickness of the crystal, where at least one complete wavelength can exist between the two long surfaces. For the same thickness, it is also possible for two or more shorter wavelengths to exist between the two surfaces. Crystals also vibrate at harmonics of the fundamental frequency.

The second mode of vibration is determined by the width of the crystal slab measured along the long parallel surfaces. This is called *width vibration*.

In addition to these two modes of vibration and their harmonics, additional modes of vibration will exist because of the twisting and bending of the crystal. The thinner the crystal, the more numerous become the resonances and also the higher the frequency of vibration.

Quartz crystals are manufactured to operate at frequencies of from 50 kHz to 50 MHz. Special crystals for certain types of test work may be obtained that will vibrate at a low frequency of 1000 Hz. A typical circuit for a quartz crystal oscillator is shown in Fig. 25-102.

25.103 Are all quartz crystals cut on the same axis?—No. Quartz crystals designed for oscillatory use are cut from the mother crystal at many different angles to the crystal axis. The axis of the cut will depend on the type of service the crystal is to render. The axis of the cut determines the stability, temperature coefficient, and frequency of oscillation. As no single type of cut will cover all the different frequency ranges required, different cuts are used for different frequency ranges and types of service. For frequencies between 300 kHz and 5 MHz, the A-T cut is generally employed; above 5 MHz, the B-T cut is used. Type X and Y cuts have poor temperature coefficients and are not used where a high degree of stability is required. As a rule, the crystal is cut from selected Brazilian quartz, free from impurities such as optical twinning, electrical twinning, bubbles, fractures, scratches, and mineral inclusions. A typical mother quartz crystal is shown in Fig. 25-103, together with the different cuts used.

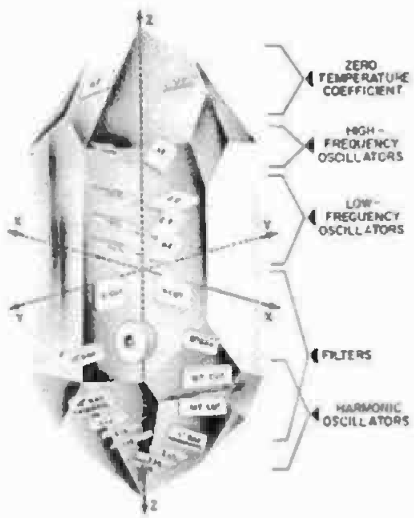


Fig. 25-103. Axes of cuts used for quartz crystals.

6 dB per Octave

f in Hz	*dB*
10	−40.00
20	−33.98
30	−30.46
40	−27.96
50	−26.02
60	−24.44
70	−23.10
80	−21.94
90	−20.92
100	−20.00
200	−13.98
300	−10.46
400	− 7.96
500	− 6.02
600	− 4.44
700	− 3.10
800	− 1.94
900	− 0.92
1,000	0
2,000	+ 6.02
3,000	+ 9.54
4,000	+12.04
5,000	+13.98
6,000	+15.56
7,000	+16.90
8,000	+18.06
9,000	+19.08
10,000	+20.00
11,000	+20.83
12,000	+21.58
13,000	+22.28
14,000	+22.92
15,000	+23.52
16,000	+24.08
17,000	+24.61
18,000	+25.11
19,000	+25.58
20,000	+26.02

Fig. 25-104. Actual decibel change for a 6-dB-per-octave curve.

25.104 Show the relationship between frequency and decibels for a 6-dB-per-octave curve, using a reference frequency of 1000 Hz.—This relationship is shown in Fig. 25-104. It will be noted for frequencies of equal increment that the change in decibels is not exactly 6 dB. Although the differences are small in some instances, they may have to be taken into consideration.

25.105 Show graphically the relationship of decibels to power level in watts.—This relationship is shown graphically in Fig. 25-105A, for a 1-milliwatt reference level. It will be observed, each time the level is increased 10 dB, the power level increases 10 dB. The chart is quite handy in determining power gains and power losses. For example, if a speaker crossover network has a 1-dB insertion loss and is driven by a 40-watt power amplifier, how much power is lost? Following the diagonal line up to the power level of 40

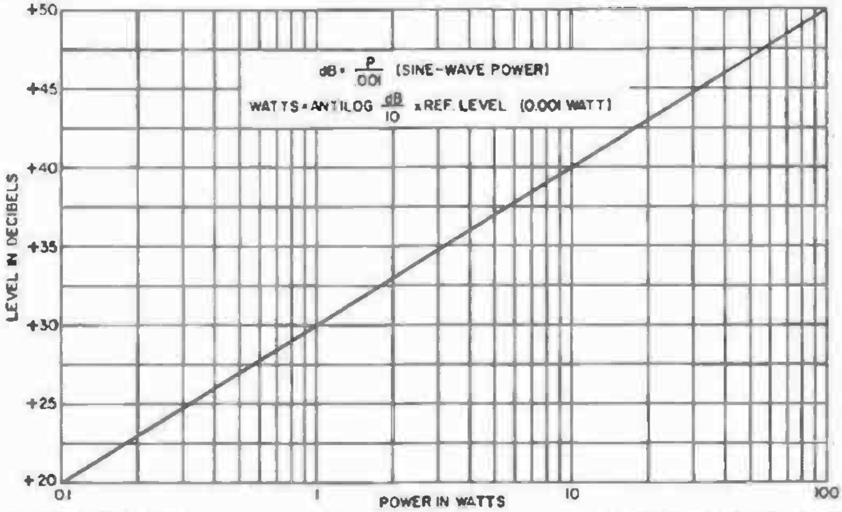


Fig. 25-105A. DB versus power in watts. Reference power 1 milliwatt (600 ohms).

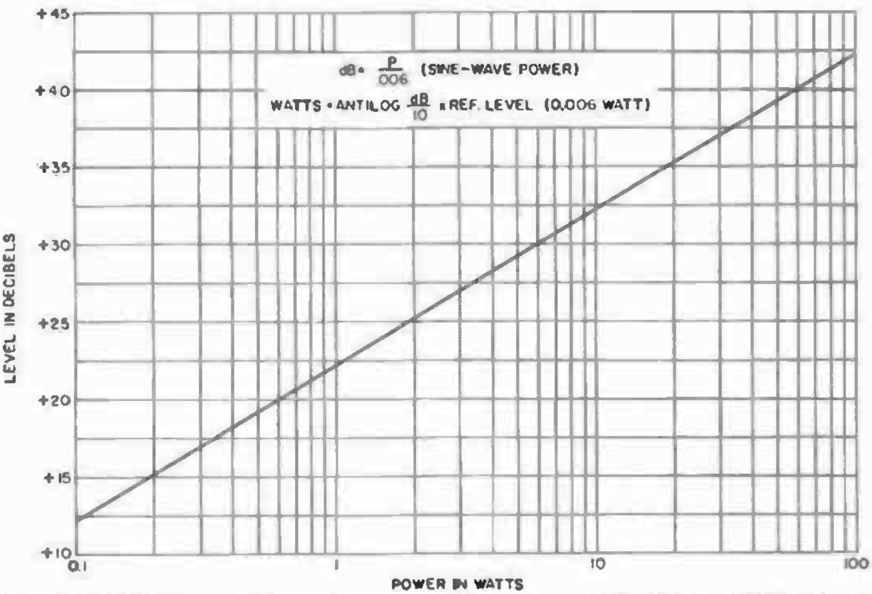


Fig. 25-105B. DB versus power in watts. Reference power 6 milliwatts (500 ohms).

watts, the level in decibels is plus 46 dB, as read from the right-hand margin. The network has a 1-dB insertion loss. Moving downward 1 dB on the diagonal line indicates a power loss of 8 watts. Thus, to restore the original level to 40 watts, the amplifier must produce 48 watts, whereby for an insertion loss of 0.25 dB the power loss is approximately 3 watts.

A similar graph for a 6-milliwatt reference level is given in Fig. 25-105B. It is used in the same manner as the graph in Fig. 25-105A.

25.106 What are the equations for calculating the time constants of a circuit containing resistance and a reactive element?

For resistance and capacitance:

$$t = RC$$

where,

t is the time constant in seconds,
R is the resistance in ohms,
C is the capacitance in farads.

For resistance and inductance:

$$t = \frac{L}{R}$$

where,

t is the time in seconds,
L is the inductance in henries,
R is the resistance in ohms.

To calculate the time constant using other units of measure, the following terms are employed:

Seconds equal megohms, microfarads.
Microseconds equal ohms, microfarads.
Microseconds equal megohms, picofarads.

For calculations involving inductance:

Seconds equal ohms and henries.
Microseconds equal ohms and microhenries.
Microseconds equal megohms and henries.

In a circuit consisting of only resistance and capacitance the time constant T is defined as the time it takes to charge the capacitor to 63.2 percent of the maximum voltage. In a circuit containing inductance and resistance, the time constant is defined as the time it takes for the current to reach 63.2 percent of its maximum voltage. Rise time (T_R) is the time it takes for the charge to rise from 10 percent to 90 percent of its maximum value. Values T and T_R can be found simultaneously by the use of the nomograph in Fig. 25-106A. A straightedge is placed from R_L to L at their respective values for the RL circuit. Values of T and (T_R) are then multiplied by 10^{-n}. Where R_L is multiplied by 10^{n}, then T and T_R are multiplied by 10^{-n}.

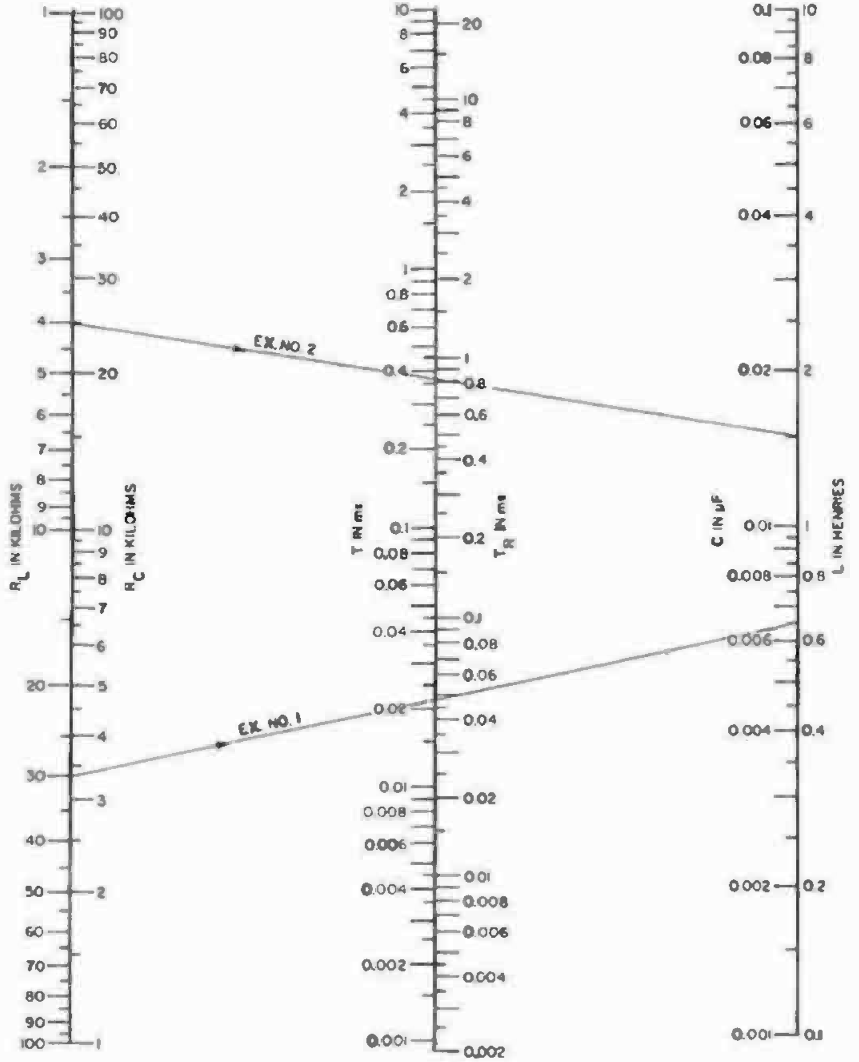


Fig. 25-106A. Time constant and rise-time nomograph (after Applebaum). (Courtesy, Electronics and Communications, Canada, and Electronic Industries, USA)

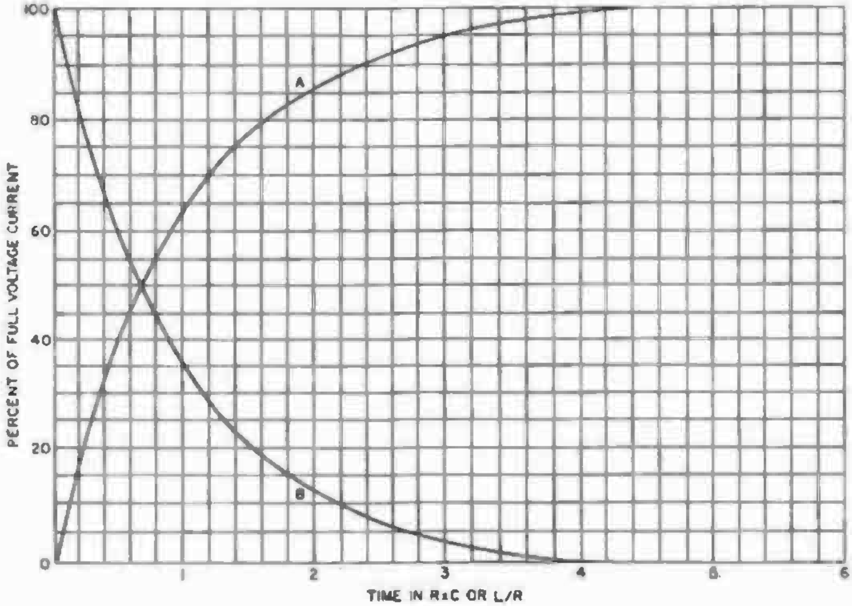


Fig. 25-106B. Universal time graph.

Example: Find T and T_R for an RL circuit in which R_L is 300,000 ohms, and L is 0.65 henry. With a straightedge from 30 on the R_L scale to 0.65 on the L scale, it intersects the T scale at 0.022 and the T_R scale at 0.05. Since R_L is multiplied by 10^4, then T and T_R are multiplied by 10^{-4}.

Example: Find T and T_R for an RC circuit where R_C is 250,000 ohms and C is 0.015 μF. The straightedge from 25 on the R_C scale to 0.015 on the C scale intersects T at 0.37 and T_R at 0.82. Since R_C was multiplied by 10 and C by 10, then T and T_R must be multiplied by 10^2. The answers are: T equals 37 microseconds, and T_R 82 microseconds. (See Question 22.74.)

When a dc voltage is applied to an RC or RL circuit, a certain amount of time is required for the capacitor to charge or the current to build up to a portion of the full value. This is termed the *time constant* of the circuit. However the time constant is not the time required for the voltage or current to reach the full value of the applied voltage. Rather, it is the time required to reach 63.2 percent of the full vallue. During the next time constant, the capacitor is charged or the current builds up to 63.2 percent of the remaining difference of full value, or to 86.5 percent of the full value. Theoretically, the charge on a capacitor or the current through a coil can never actually reach 100 percent, but is considered to be 100 percent after five time constants have passed. When the voltage is removed, the capacitor discharges and the current decays to 63.2 percent of the full value. These two factors are shown graphically in Fig. 25-106B. Curve A shows the voltage across a capacitor when charging, or the current through an inductance on build-up. Curve B shows the capacitor voltage when discharging or an inductance current decay. It is also the voltage across the resistor on charge or discharge.

25.107 Can a low-voltage fuse be used in a high-voltage circuit if the amperage rating is correct?—Low-voltage fuses should not be used in high-voltage circuits because the break distance is less and arc-over may result. The break distance is defined as the separation of the parts supporting the fusible materials. Fuses are constructed of a fast-melting alloy and are so designed that when a given current is exceeded, the alloy will melt. A fuse has no voltage limit, except that imposed by the construction and insulation. Heat developed across the fusible element by the current and the resistance of the alloy (I^2R) will blow the fuse when a given current is exceeded. The only factor important to a fuse is the amount of current through it.

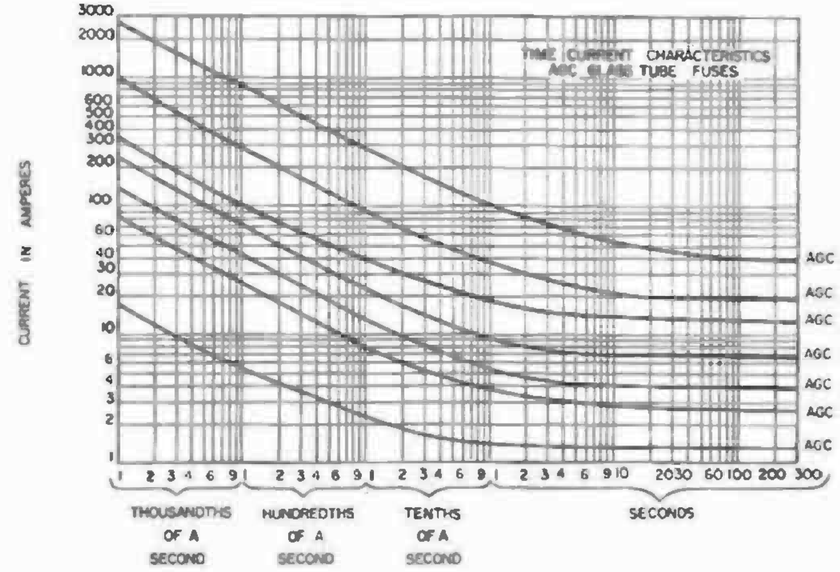


Fig. 25-108. Fusing time chart for small glass-type fuses. (Courtesy, Bussmann Mfg. Co.)

25.108 Can a high-voltage fuse be used in a low-voltage circuit if the amperage rating is correct?—Yes. A higher-voltage-rated fuse can always be used in a low-voltage circuit provided the current-carrying capacity is correct. Fuses rated at 250 volts may be used in 6-, 12-, 24-, 32-, or 115-volt circuits and 32-volt fuses may be used in 6-, 12-, or 24-volt circuits.

The time/current characteristics for small glass-type fuses are given in Fig. 25-108.

25.109 What is a parallel-T network?—A resistive and capacitive network consisting of three resistors and three capacitors, as shown in Fig. 25-109. Such networks are used in bridge circuits and distortion-meter circuits, as well as other devices where an unusually sharp null is required. The three resistors may be ganged mechanically to provide a variable frequency range and permit single-dial operation. The relation of the capacitors and resistors is given:

$$C1 = C2 \qquad C3 = 2C1$$

$$R1 = R2 \qquad R3 = \frac{R1}{R2}$$

The null point of a parallel-T network may be considerably sharpened by using a negative feedback loop around the stages encompassing the network.

To illustrate the relationship of capacitance to resistance in such networks, typical values for a 400-Hz network are given below:

C1 and C2	*C3*	*R1 and R2*	*R3*
0.001	0.002	397,885	198,942
0.01	0.02	39,788	19,894
0.05	0.01	7957	3978

The dip filter described in Question 7.72 is a typical example of a parallel-T type of attenuator.

25.110 Describe a compactron tube.—Compactron is a trade name of General Electric given to a vacuum tube that houses the elements of four individual tubes in one glass envelope. A typical tube of this type might house a detector diode, audio stage, output pentode, and a rectifier. Similar tubes house a pentagrid section and an rf amplifier stage. A T-9 bulb and a 12-pin base are used.

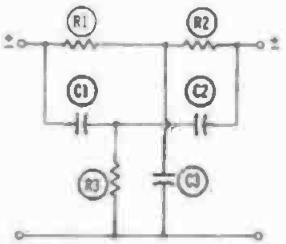


Fig. 25-109. A parallel-T network.

25.111 How are permanent magnets charged?—By passing a heavy direct current through a coil wound around the material to be magnetized. The amount of magnetic energy that can be stored in a permanent magnet is dependent on the volume of the magnetic material and the maximum flux density in the material.

The schematic diagram for a magnet-charging device is shown in Fig. 25-111. The components consist of a step-up transformer, T1, a rectifier, R, a large capacitor bank, C, two ignitron tubes, V1 and V2, and a coupling transformer, T2, for coupling to the device to be magnetized. The circuit shown is commonly known as a stored-energy magnetizer in which the energy is stored as a charge on a capacitor for supplying a high peak current.

The ac voltage at the primary of T1 is stepped up by the secondary and rectified. The rectified dc voltage charges the capacitor bank, C. This actuates the magnetizer switch which causes ignitron tube V1 to conduct, thus permitting the capacitor bank to discharge into the primary of transformer T2. The secondary of T2 is connected directly to the magnetizing loop wound around or slipped through the device to be magnetized. Ignitron tube V2 prevents current reversals in the magnetizing loop. The magnitude of the current in the secondary of T2 is controlled by the value of the dc voltage applied to the capacitor bank and the turns ratio of transformer T2.

The dc voltage applied to the capacitor bank ranges between 1000 and 3000 volts. The turns ratio of transformer T2 may be varied from 100:1 to 600:1, permitting currents of 40,000 to 200,000 amperes to be sent through the magnetizing loop. The capacitor bank for this particular circuit approximates 2000 μF.

A more simple method of magnetizing small magnets such as used in meters and similar devices may be accomplished by wrapping about 3 feet of insulated ⅜-inch copper braid around the magnet to be charged. Polarity of the magnet is determined by means of a compass, remembering the north indicating end of the compass needle is in reality a south pole and hence it will be attracted to the north pole of the magnet.

Next the direction of the current through the coil must be determined. This is accomplished by the use of the right-hand rule. Grasp the coil in the right hand with the extended thumb pointing toward the north pole of the magnet. The four fingers now point in the direction in which the current must flow through the coil. Note which end the current must enter the coil and connect this end to the positive terminals of a storage battery of considerable ampere-hour rating (100 or greater). Touching the other end to the negative terminal once or twice should be sufficient for most purposes.

25.112 What is a discriminator?—A circuit used in fm radio receivers for converting the fm signals to audio frequencies. Discriminators are used in place of the conventional detector in an a-m radio receiver. Very often discriminators are used in audio-frequency test equipment such as a flutter meter, fm pickups, and disc-recorder cutting-head calibrating equipment. A typical circuit appears in Fig. 25-112A.

The fm signal frequency is converted to an audio-frequency signal by rectification of the voltage response of two tuned circuits to the frequency and impressing the difference on the grid of an audio amplifier tube. The response characteristics of the tuned circuits rise from a low value, below the resonant frequency, to a peak at the resonant

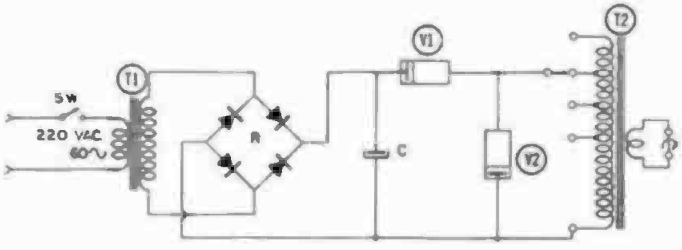


Fig. 25-111. A high-power magnet-charging circuit.

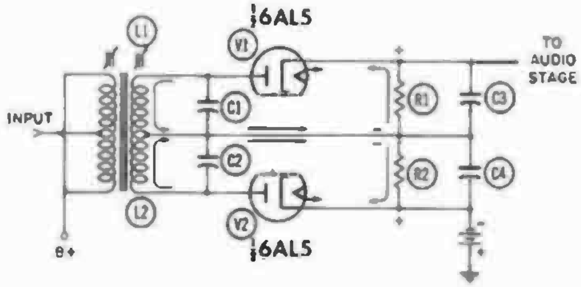


Fig. 25-112A. A balanced-discriminator circuit.

frequency, and then fall to zero above the resonant frequency. Thus, a maximum voltage appears across the terminals of the tuned circuit at the resonant frequency. This voltage decreases as the signal frequency changes above or below the resonant frequency. (See Fig. 25-112B.)

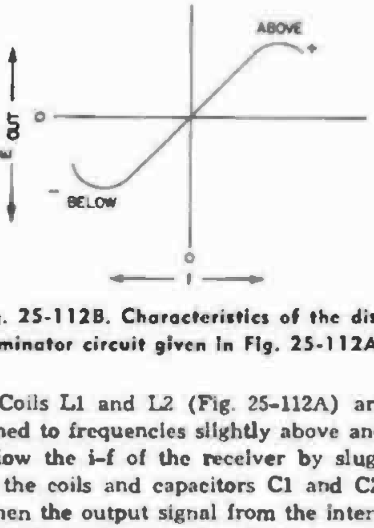


Fig. 25-112B. Characteristics of the discriminator circuit given in Fig. 25-112A.

Coils L1 and L2 (Fig. 25-112A) are tuned to frequencies slightly above and below the i-f of the receiver by slugs in the coils and capacitors C1 and C2. When the output signal from the intermediate amplifier is of an instantaneous positive polarity, the ends of the discriminator secondary coil connected to the plates of the diodes are positive. Therefore, the diodes conduct. When the signal is of negative polarity, the diodes do not conduct.

The magnitude of the voltages developed across load resistors R1 and R2 depends on the frequency of the input signal to the discriminator. The current, shown by the arrows, produces bucking voltages across the load resistors and the difference voltage is taken from across R1 and impressed on the control grid of the audio amplifier.

At the center frequency, the response of both circuits is the same and, consequently, the opposed voltages developed across R1 and R2 are the same, resulting in a zero output voltage from the discriminator. At frequencies above the center frequency, a larger current flows in the circuit consisting of L1, C1, R1, and C3. As a result, a voltage of greater magnitude is developed across R1, and the output of the discriminator is positive.

At frequencies below the center frequency, a larger current flows in the circuit consisting of L2, C2, R2, and C4. The larger voltage is then developed across resistor R2 and the output signal becomes negative. In this manner, an audio-frequency signal is obtained from the high-frequency fm signal. The bottom end of R2 is returned to a negative bias voltage point and is used as a reference voltage for the audio signal. During negative half-cycles of the input, capacitors C3 and C4, because of their low discharge rate relative to the intermediate frequency, maintain the voltage developed across resistors R1 and R2, ensuring a distortionless audio output signal.

It is customary to precede the discriminator by one or more limiter amplifier stages which remove amplitude variations from the fm signal. This stage may be a tube operating as a class-B amplifier with the plate and screen voltage so low that saturation is obtained with a very small signal voltage at the control grid.

25.113 What is a thyratron switching circuit?—Two type-885 thyratron tubes connected in a double-ended circuit, as shown in Fig. 25-113. A modified square wave is applied to the control grid of tube V1, amplified, and then applied through an interstage transformer T1 to the control grids of two thyratrons, V2 and V3. The signal voltages at the two control grids are of equal amplitude and 180 degrees out of phase.

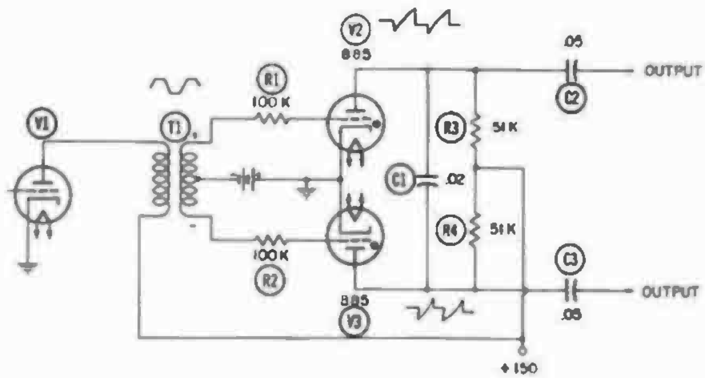


Fig. 25-113. A thyratron switching circuit.

Assume the modified square wave applied to the control grid of V2 has a positive instantaneous polarity. Tube V2 will fire, charging capacitor C1, whereas the control grid of V3 becomes negative. During the second half-cycle, the voltages at the control grids of V2 and V3 are reversed and V3 fires, charging capacitor C1 in the opposite direction.

This action will continue as long as the signal is maintained at the control grid of V1. The waveform at the plate of V2 and V3 is independent of the waveform applied to the control grid of V1. The output voltage is taken from plate load resistors R3 and R4 through capacitors C2 and C3. Resistors R1 and R2 are current-limiting resistors to prevent damage to the thyratrons. Such circuits are used to control servomechanisms and motor circuits.

25.114 How may a glow tube be used as a fuse indicator?—By connecting the lamp in series with a current-limiting resistor which is in parallel with the fuse, as shown in Fig. 25-114. If the fuse is good, it acts as a short circuit across the glow lamp and the lamp remains dark. Where the fuse blows, the circuit potential is applied to the lamp, causing it to glow, indicating the fuse has blown.

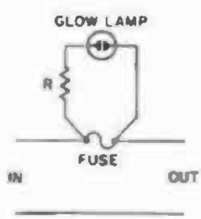


Fig. 25-114. A fuse-indicator circuit using a glow lamp.

25.115 What is the relation between decibels, voltage, and power?—This relationship, in tabulated form, is given in Fig. 25-115. The power level in decibels appears in the first column, the rms voltage in the second column, and the power in watts in the third column. The foregoing values are in reference to 1 milliwatt of power in a 600-ohm circuit.

25.116 What is a megger?—A small constant-voltage, hand-cranked generator manufactured by the James Biddle Co. and used for measuring the resistance, leakage of audio- and radio-frequency transmission lines, power cables, generator and motor insulation, transformers, and other electrical and electronic equipment. A typical instrument is shown in Fig. 25-116A.

Basically, the device consists of a hand-cranked generator, which when cranked at approximately 160 rpm will produce 500 volts dc, and a meter calibrated to read directly in ohms similar to an ohmmeter. An overriding clutch limits the speed of the armature and, with the circuit as shown in Fig. 25-116B, the voltage is held constant.

Referring to the schematic diagram, coils A and B are mounted in a fixed relationship to each other and on a moving system, which is pivoted in spring-supported jewel bearings. The moving system carries a pointer and a balancing weight, and is free to rotate through about 70 degrees in a permanent magnetic field. A C-shaped iron core, C, is mounted in a fixed position coaxial with the moving system and forms an important part of the ohmmeter magnetic circuit. There are no control springs such as are used in the conventional voltmeter or ammeter. Current is fed to the coils by lead-in spirals which offer only slight resistance.

Coil A is connected across the current supply in series with the resistance under test and fixed ballast resistor R1. It is called the current coil. Coil B is also connected across the current supply in series with fixed resistor

REFERENCE LEVEL .001 w. in 600Ω Decibels	*Volts rms*	*Watts*	*REFERENCE LEVEL .001 w. in 600Ω Decibels*	*Volts rms*	*Watts*
−20	**.07746**	**.000010**	16	4.8873	0.03981
−19	.08691	.0000126	17	5.4838	0.05012
−18	.09752	.0000158	18	6.1531	0.06310
−17	.10949	.0000199	19	6.9036	0.07943
−16	.12275	.0000251	20	7.7460	0.10000
−15	.13773	.0000316	**+21**	**8.6913**	**0.12589**
−14	.15454	.0000398	22	9.7519	0.1585
−13	.17323	.0000501	23	10.949	0.1995
−12	.19458	.0000631	24	12.275	0.2512
−11	.21830	.0000794	25	13.773	0.3162
−10	**.24495**	**.0001000**	26	15.454	0.3981
− 9	.27485	.0001259	27	17.323	0.5012
− 8	.30838	.0001585	28	19.458	0.6310
− 7	.34742	.0001995	29	21.830	0.7943
− 6	.38823	.0002512	30	24.495	1.0000
− 5	.43557	.0003162	**+31**	**27.485**	**1.2589**
− 4	.48873	.0003981	32	30.838	1.5849
− 3	.54838	.0005012	33	34.742	1.9953
− 2	.61531	.0006310	34	38.823	2.5119
− 1	.69036	.0007943	35	43.557	3.1623
0	.77460	.0010000	36	48.873	3.9811
+ 1	**.86913**	**.001259**	37	54.838	5.0119
2	.97519	.001585	38	61.531	6.3096
3	1.0949	.001995	39	69.036	7.9433
4	1.2275	.002512	40	77.460	10.000
5	1.3773	.003162	**+41**	**86.91**	**12.58**
6	1.5454	.003981	42	97.51	15.84
7	1.7323	.005012	43	109.49	19.95
8	1.9458	.006310	44	122.75	25.11
9	2.1830	.007943	45	137.73	31.62
10	2.4495	0.01000	46	154.54	39.81
+11	**2.7485**	**0.01259**	47	173.23	50.11
12	3.0838	0.01585	48	194.58	63.09
13	3.4742	0.01995	49	218.30	79.43
14	3.8823	0.02512	50	244.95	100.00
15	4.3557	0.03162			

Fig. 25-115. Decibels, rms voltages, and watts.

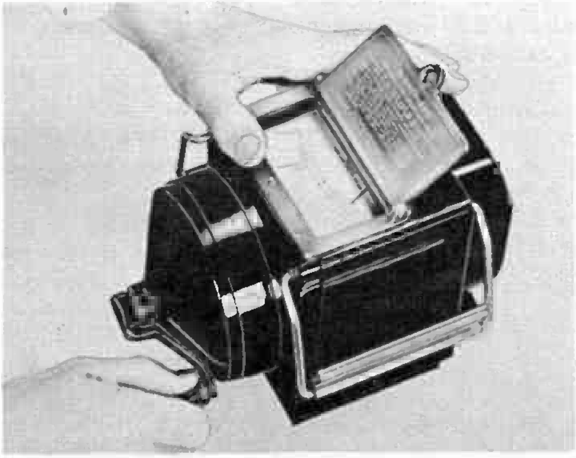


Fig. 25-116A. A megger manufactured by the James G. Biddle Co.

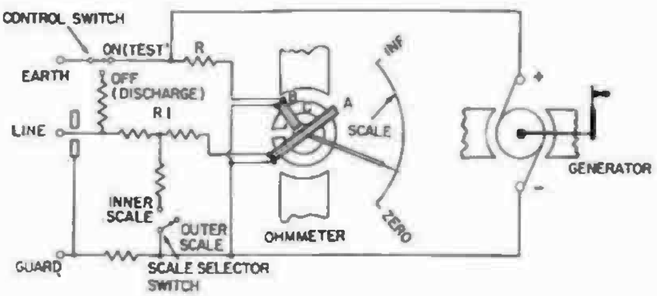


Fig. 25-116B. Schematic diagram of the megger pictured in Fig. 25-116A. The resistance to be measured is connected between the line and earth terminals.

R and is called the potential coil. Coils A and B are so connected that when current is supplied they develop opposing torques and tend to turn the moving system in opposite directions. The pointer assumes a position where the two torques balance.

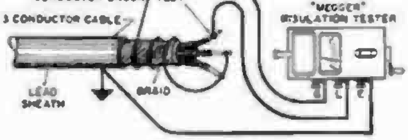


Fig. 25-116C. Use of the guard connection on the megger to eliminate the effect of leakage to ground.

When the megger is operated with either a perfect insulator or is open between the line and ground terminal, no current will flow in the deflecting coil A. However, coil B receives current from the generator and will position itself opposite the gap in the C-shaped

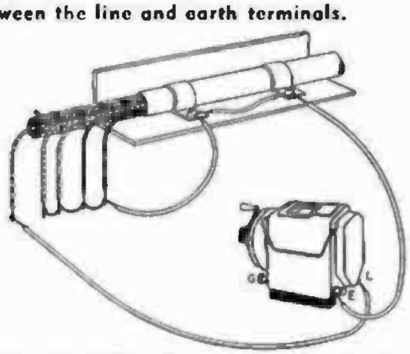


Fig. 25-116D. Connections for testing the insulation resistance of one wire in a multiconductor cable against all other wires and the sheath.

iron core, C. This is the infinity calibration point on the meter scale.

When a resistance is connected across the terminals, current will flow in coil A and the corresponding torque will draw the control coil B away from the infinity calibration position into a field of gradually increasing magnetic strength until a balance is obtained between the forces acting on the two coils. Thus, control coil B acts as a retaining spring. The scale calibrations are independent of the generator voltage because the two coils receive current from the same source (the generator) and any change in the generator voltage will affect both coils in the same proportion. Such instruments may be obtained with internal voltages up to 2500 volts.

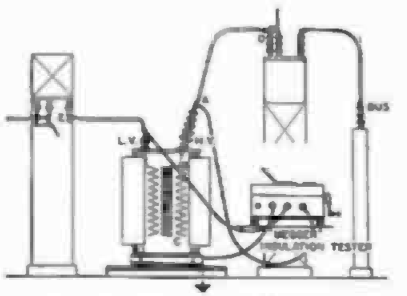


Fig. 25-116E. Connections for measuring the insulation resistance between a high- and low-voltage transformer winding without being affected by leakage to ground. Note the use of the guard connection.

The methods of connecting the megger to a power cable, transmission line, and transformer for measuring leakage resistance to ground are shown in Figs. 25-116C to E. The instrument pictured has two resistance ranges: 200 megohms and 20 megohms. It will permit the measurement of insulation resistance to as low as 1000 ohms. The guard connection prevents errors due to leakage internally or externally between the positive and ground sides of the megger. Such leakage is caused by moisture or dirt between the terminals. The guard circuit offers a low-resistance path for leakage current directly to ground without passing through the deflecting coil of the ohmmeter. The resistance coils and other elements of the instrument are mounted on the guard-circuit supports inside the case.

25.117 ***What causes an arc lamp operated from rectified ac to whistle acoustically, and how can this be eliminated?***—Carbon-arc lamps used on motion picture sets are generally supplied from a source of rectified ac or from a dc generator. A singing noise is sometimes radiated acoustically and picked up by the microphone. This noise is caused by the ripple frequency

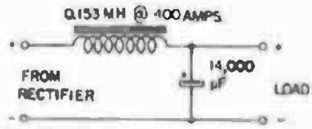


Fig. 25-117A. A 400-ampere, low-pass filter for removing arc-lamp whistle.

of the rectifier or by the generator commutator. If the rectified voltage is from a three-phase source, the acoustically radiated frequency is the second harmonic of 360 Hz, or 720 Hz. If the rectified voltage is 60 Hz, the ripple frequency is 120 Hz.

This annoying acoustical interference can be eliminated in two ways. In the instance of the rectifier a low-pass, brute-force filter, shown in Fig. 25-117A, is connected in the supply line at the rectifier. It will be noted the inductance required is quite small, but must be capable of carrying at least 10 percent more current than the full-load current. The filter shown was designed for use with a 400-ampere rectifier at 120 volts dc.

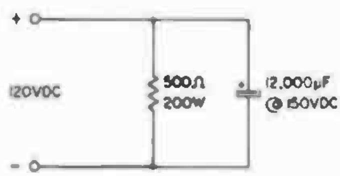


Fig. 25-117B. Capacitor bank for removing arc-lamp whistle.

To remove generator commutator ripple will require at least 10,000 µF and preferably 20,000 µF, connected directly across the output terminals of the generator. In addition, at the end of the supply line an additional filter of 10,000 µF is connected at each arc lamp. These units are portable and connected as required. It will be noted that because of the danger of serious injury from these capacitor banks when charged, a 500-ohm 200-watt resistor is connected *permanently across the bank* to discharge it when it is disconnected.

The capacitor banks consist of bakelite-encased 150-Vdc electrolytic capacitors mounted in a wooden box. The capacitor values given are the minimum and may be increased if desired. (See Question 3.46.)

25.118 ***What is the time duration per cycle for frequencies between 10 Hz***

and 20 MHz?—The time in microseconds, milliseconds, and seconds is given in the table below.

Freq.	*Micro-seconds*	*Milli-seconds*	*Seconds*
1.0 Hz	1,000,000	1000	1.0
10 Hz	100,000	100	0.10
50 Hz	20,000	20	0.02
100 Hz	10,000	10	0.01
200 Hz	5000	5	0.005
500 Hz	2000	2	0.002
1 kHz	1000	1.0	0.001
2 kHz	500	0.5	0.0005
5 kHz	200	0.2	0.0002
10 kHz	100	0.1	0.0001
20 kHz	50	0.05	0.00005
50 kHz	20	0.02	0.00002
100 kHz	10	0.01	0.00001
200 kHz	5	0.005	0.000005
500 kHz	2	0.002	0.000002
1 MHz	1.0	0.001	0.000001
2 MHz	0.5	0.0005	0.0000005
5 MHz	0.2	0.0002	0.0000002
10 MHz	0.10	0.0001	0.0000001
20 MHz	0.05	0.00005	0.00000005

The time period equals 1/f. Thus, the time period for 1 cycle of 400 Hz is 1/400, or 0.0025 second. The relationship of milliseconds to microseconds and vice versa is shown graphically in Figs. 25-118A and B for ready reference, with milliseconds to seconds shown in Fig. 25-118C.

25.119 ***How may the balance of a d'Arsonval meter movement be checked?*** —Lay the meter flat on its back and note whether the pointer coincides with the zero calibration mark. If not, adjust the set screw on the front of the meter to bring the pointer to zero. If the meter hand is bent, it must be straightened before this adjustment is made.

Next, place the meter in an upright positiou and turn it until the pointer is in a vertical direction or position. If the pointer does not zero, the horizontal balance weight needs adjustment. Now, continue to hold the meter in an upright position and rotate it, either clockwise or counterclockwise, until the pointer is horizontal. If the pointer does not zero in the horizontal position, the tail weight on the pointer needs adjustment.

25.120 ***How is the maximum safe capacitance calculated for connection***

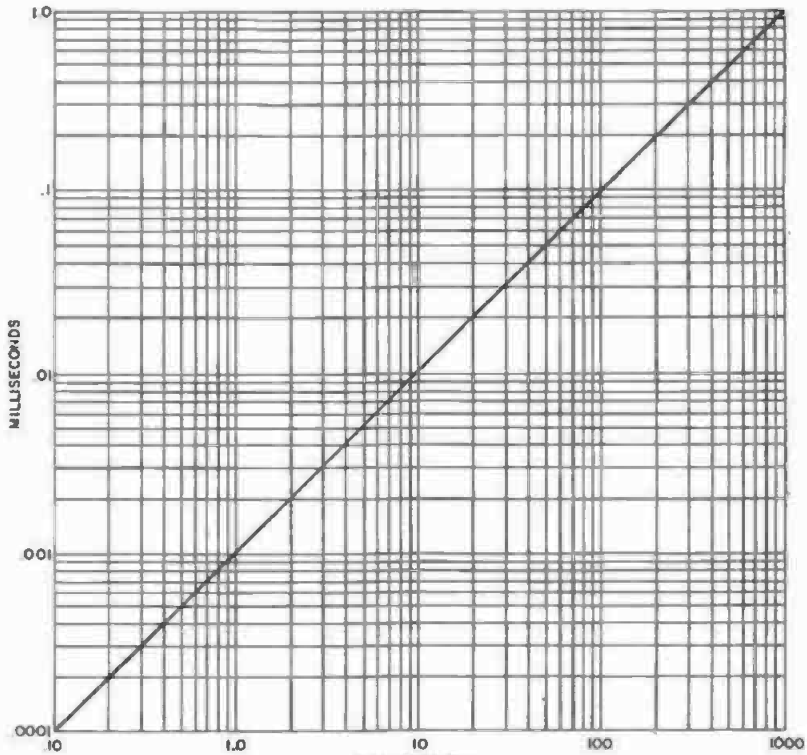


Fig. 25-118A. Conversion chart, milliseconds to microseconds.

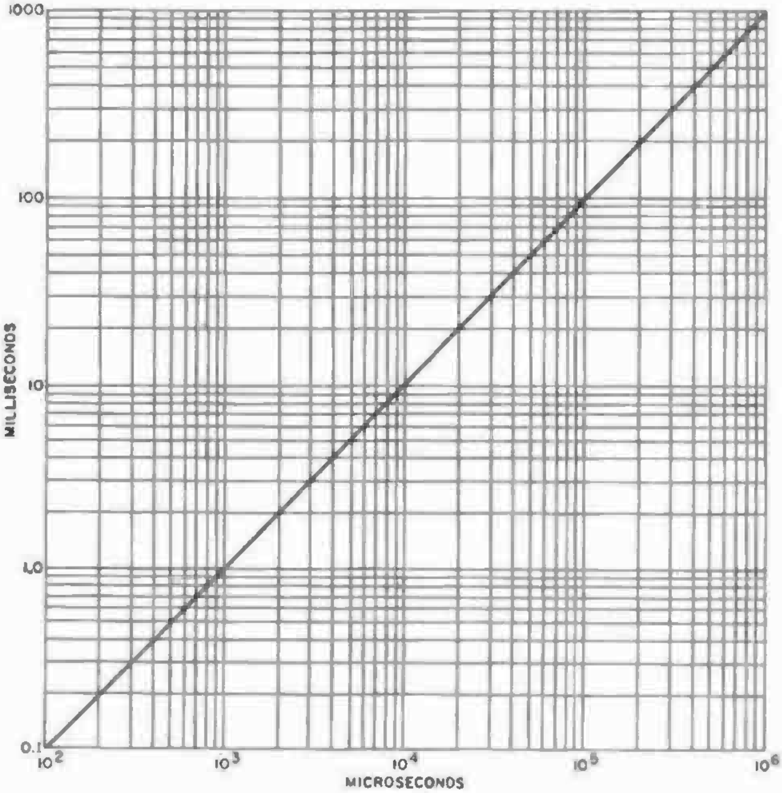


Fig. 25-118B. Conversion chart, milliseconds to microseconds.

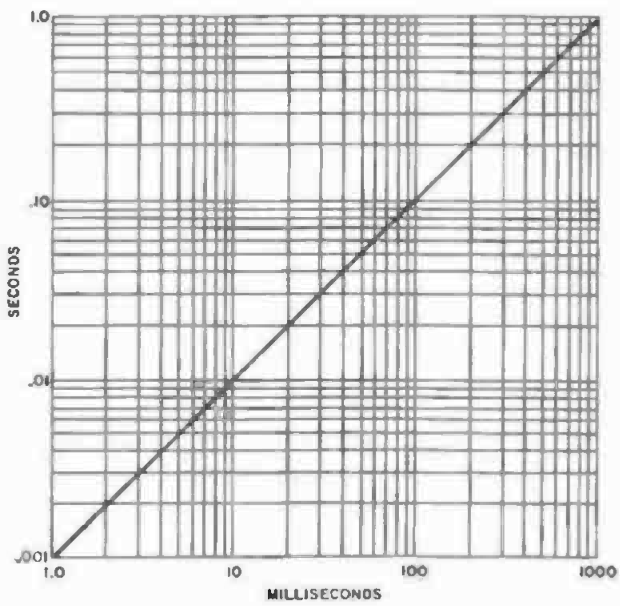


Fig. 25-118C. Conversion chart, seconds to milliseconds.

between a power line and the chassis of a piece of electrical equipment?—To meet the Underwriters' specifications, it may be calculated:

$$C \text{ (single phase)} = \frac{860}{F \times E}$$

$$C \text{ (three-phase)} = \frac{500}{F \times E}$$

where,

C is the capacitance in μF,
F is the frequency of the power source,
E is the voltage of the power source.

25.121 ***What is a synchroscope?***—A special type cathode-ray oscilloscope designed for the precise study of periodic and nonperiodic pulses such as are found in radar circuits. The name is derived from the fact the horizontal sweep is generated only when a synchronizing signal is present. The sweep is made variable and calibrated in microseconds for measurement of the pulse length.

25.122 ***Define the right-hand rule.***—The right-hand rule is a method devised for determining the direction of a magnetic field around a conductor carrying a direct current. The conductor is grasped in the right hand with the thumb extended along the conductor as shown in Fig. 25-122. The thumb points in the direction of the current. If the fingers are partly closed as shown, the finger tips will point in the direction of the magnetic field.

Maxwell's rule states: If the direction of travel of a right-handed corkscrew represents the direction of the current in a straight conductor, the direction of rotation of the corkscrew will represent the direction of the magnetic lines of force.

25.123 ***How can the formation of static charges on plastic meter cases be prevented?***—The inner and outer surfaces of the cases can be wiped with a detergent used for washing dishes, or with one of several antistatic solutions developed for this purpose. Meters of recent manufacture have antistatic covers.

25.124 ***What is electrolysis and what is its cause?***—Electrolysis is defined as the process of chemical decomposition caused by the influence of electricity. Electrolysis is caused by the leakage of electricity in the earth, setting up a chemical action leading to the decomposition of protective sheaths on cables, pipes, rails, and other metallic objects buried in the earth. The principal cause of electrolysis is the presence of stray currents from power, direct-current arc light, and electrical railway systems. Little damage is caused by the alternating currents through the earth.

Damage from electrolysis is, as a rule, due to direct current from street railway systems causing stray currents from the trolley car through the ground and back to the generator. The area covered by the stray currents varies with the car movement as it travels to

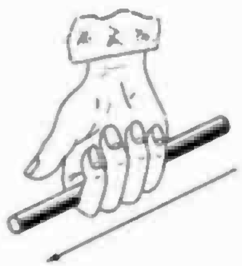


Fig. 25-122. The right-hand rule for determining the direction of a magnetic field around a conductor in which direct current is flowing.

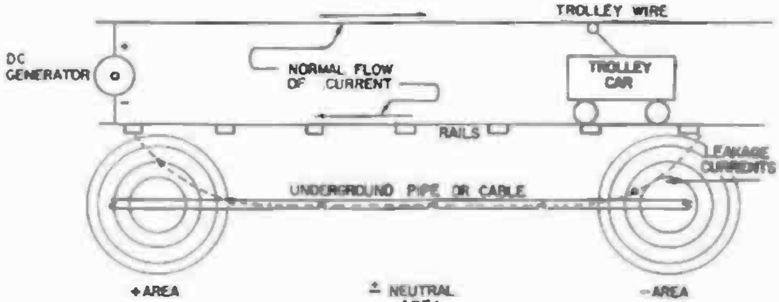


Fig. 25-124. Stray current between a street railway system and pipe or cable buried in the earth.

and from the generator end. If cables or pipes are in the path of the stray currents, the currents will flow through the pipes rather than the ground because of lower resistance.

Tests indicate that the rails are positive to the pipes in the positive area of the pipes, and the current alternates in the neutral areas. The pipes are positive in the negative areas. This latter point is where the stray currents leave the cable and return to the rails. It is also the area where the greatest damage occurs.

Electrolysis also attacks aerial cables, causing crystallization of the lead sheath and, ultimately, damage to the cable pairs contained in the sheath. The only solution to the problem of electrolysis is to ground all equipment, conduits, pipes, and circuits where practical. Fig. 25-124 illustrates how stray currents flow outward from a railway generator station, through pipes buried in the ground, and back to the generator again.

Gas pipes should never be used for ground connections because an insulated coupling is generally used at the meter. Also, the thread-sealing compound acts as an insulator. If a water pipe is used for a ground connection, a heavy copper strap should be connected around the meter to make good connection to the feed pipe.

Sound installation grounds must not be connected to a neutral wire or ground connection for the lighting system, because of the ac potential developed by leakage current and current back to ground through the neutral wire. Grounding systems are discussed in Section 24.

25.125 State Thevenin's theorem.—Thevenin's theorem is a method used for reducing complicated networks to a simple circuit consisting of a voltage source and a series impedance. The theorem is applicable to both ac and dc circuits under steady-state conditions.

The theorem states: The current in a terminating impedance connected to any network is the same as if the network were replaced by a generator with a voltage equal to the open-circuit voltage of the network, and whose impedance is the impedance seen by the termination looking back into the network. All generators in the network are replaced with impedances equal to the internal impedances of the generators.

25.126 What are the equations used for the solution of capacitor problems?
Two capacitors in series:

$$C_t = \frac{C_1 \times C_2}{C_1 + C_2}$$

$$C_x = \frac{C_1 \times C_t}{C_1 - C_t}$$

$$C_x = \frac{10^6 \times I}{2\pi fE}$$

Three or more capacitors in series:

$$C_t = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}}$$

Capacitors in parallel:

$$C_t = C_1 + C_2 + C_3 + \ldots$$

where,
C_t is the total capacitance,
C_x is an unknown capacitance.

25.127 What are the equations used for the solution of resistance problems?
Two resistances in parallel:

$$R_t = \frac{R_1 \times R_2}{R_1 + R_2}$$

$$R_x = \frac{R_1 \times R_t}{R_1 - R_t}$$

Three or more resistors in parallel:

$$R_t = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \ldots}$$

Resistors in series:

$$R_t = R_1 + R_2 + R_3 \ldots$$

where,
R_t is the total resistance,
R_x is an unknown resistance.

25.128 What are the equations used for the solution of inductance problems?
Inductors in series, with *multiple* coupling zero:

$$L_t = L_1 + L_2 + L_3 + \ldots$$

$$L_x = \frac{X_L}{2\pi f}$$

Two inductors in parallel:

$$L_t = \frac{L_1 \times L_2}{L_1 + L_2}$$

$$L_x = \frac{L_1 \times L_t}{L_1 - L_t}$$

Three or more inductors in parallel:

$$L_t = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3} + \cdots}$$

where,
- L_t is the total inductance,
- L_x is the unknown inductance.

25.129 What are the equations used for the solution of resonant-circuit problems?

$$F = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi C X_c} = \frac{X_L}{2\pi L}$$

$$L = \frac{X_L}{2\pi f} = \frac{1}{(2\pi f)^2 C}$$

$$C = \frac{1}{2\pi f X_c} = \frac{1}{(2\pi f)^2 L}$$

$$X_L = 2\pi f L$$

$$X_c = \frac{1}{2\pi f C}$$

$$Z = R \text{ when } X_L = X_c$$

where,
- F is the frequency,
- L is the inductance,
- C is the capacitance,
- X_L is the inductive reactance,
- X_c is the capacitive reactance,
- Z is the impedance,
- R is the dc resistance.

25.130 What is the equation for calculating the Q of a capacitor?

$$Q = \frac{1}{2\pi f CR} = \tan \phi \approx \frac{1}{PF}$$

where,
- F is the frequency,
- C is the value of capacitance,
- R is the internal resistance of the capacitance,
- PF is the power factor.

25.131 Is the inductance of a coil affected by the type of wire used for its construction?—No; the inductance will not be affected, regardless of the type of wire employed, for a given number of turns and other factors. However, the Q of the coil will be governed by the ohmic resistance of the wire. Coils wound with silver or gold wire have the highest Q for a given design.

25.132 What are network transfer functions?—They are the ratio of the input to output voltage for a given type of network containing resistive and reactive elements. The transfer functions for networks consisting of resistance and capacitance are given in Fig. 25-132. The expressions for the transfer functions of the networks are: A equals 6.28f; B equals RC; C equals R_1C_1; D equals R_2C_2; "n" is a positive multiplier. The values of C are expressed in farads, of f in Hz, and of R in ohms.

25.133 What is a percentage change chart?—A chart as shown in Fig. 25-133 for converting percentage change of voltage, current, power, resistance, and impedance to decibels, or vice versa. To illustrate how the chart is used, suppose it is desired to find the percentage change for a loss of 6 dB. Using the 100-percent point to represent the original value of voltage, follow the 6-dB line at the right to where it intercepts the diagonal line, then read downward to 50 percent. For a change of 20 dB, the percentage of the voltage left is 10 percent, or a change of 90 percent down from the original value. Values of percentage may be determined by reversing the procedure.

The chart is also useful in determining the percentage of modulation for a light modulator relative to a given deflection. The level of the harmonics for a given amount of percentage of distortion may be found by entering the chart at the percentage of distortion and following it upward to the diagonal line. The level in decibels is read from the vertical margin.

Example: What is the level of the harmonics for a percentage of distortion of 1.5 percent? Answer: 36.5 dB below the fundamental frequency. If the problem is one involving power, the answer is divided by 2. Thus, a change of 50 percent in power is equal to a change in level of 3 dB.

25.134 Define the term "waveform in quadrature."—It is the phase relationship between two periodic quantities of the same period when the phase difference between them is 90 degrees or one-quarter of a period. This may be seen by referring to Fig. 25-134. Here are seen two waveforms of the same frequency, but one-quarter period or 90 degrees apart. The displacement is taken at the peak of the waveform.

25.135 What are the equations for converting Fahrenheit to centigrade, and vice versa?

$$\text{Degrees C} = \frac{(F - 32) \times 5}{9}$$

$$\text{Degrees F} = \frac{C \times 9}{5} + 32$$

25.136 How may the temperature rise of a generator or motor be mea-

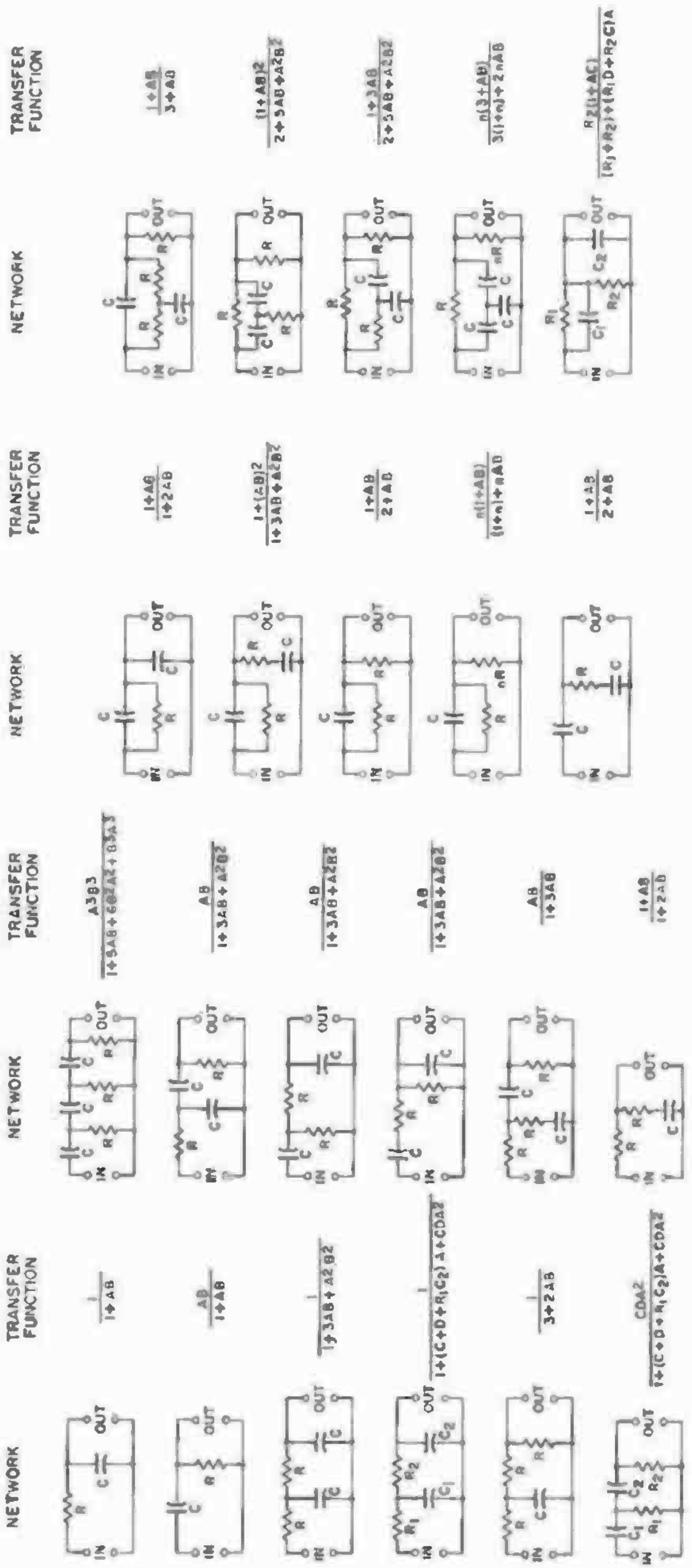


Fig. 25-132. RC network transfer functions.

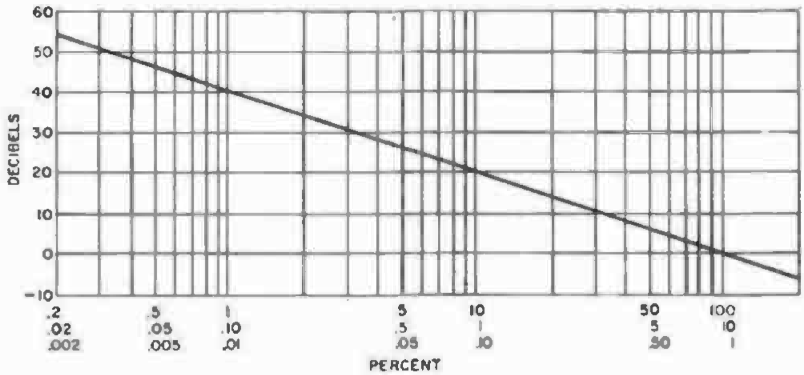


Fig. 25-133. Decibel/percentage ratio chart. Add 20 dB for each decade below 100 percent.

sured?—Internal temperature rise of a motor or generator is measured by measuring the resistance of either the field or armature windings of the machine.

If the temperature rise of a dc generator is to be measured, it is done in the following manner: Two convenient segments of the commutator are selected and marked for identification. The brushes are removed and, while the machine is cold, the resistance of the armature winding between the two commutator bars is accurately measured with a resistance bridge. The brushes are then replaced and the machine operated at its rated load and duty cycle. A thermometer is taped to the outside case and the machine run until the temperature stabilizes. The resistance of the previously identified winding is then measured. From these tests, the temperature rise may be computed:

$$T_2 = \frac{R_2}{R_1}(234.5 + T_1) - 234.5$$

where,

- T_1 (in degrees centigrade) is the measurement of the temperature at the time the first resistance measurement (R_1) was taken,
- T_2 is the temperature in degrees centigrade at the time the second resistance measurement (R_2) was made.

The value 234.5 is a constant.

25.137 What is the relationship of the turns in a coil to its inductance?

1. The inductance is proportional to the square of the turns.
2. The inductance is increased as the permeability of the core material is increased.

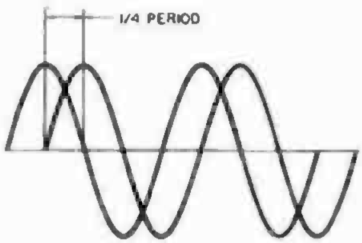


Fig. 25-134. Waveforms in quadrature. The displacement is measured from peak to peak.

3. The inductance increases as the cross-sectional area of the core material is increased.
4. The inductance is increased as the length of the winding is increased.

25.138 When is the maximum voltage induced in a conductor moving in a magnetic field?—The voltage is proportional to the number of magnetic lines of force cut by the conductor moving in the field. A conductor moving parallel to the lines of force cuts no lines of force; therefore, no current is generated in the conductor. A conductor moving at right angles to the lines of force will cut the maximum number of lines per inch per second; therefore, the voltage will be at the maximum.

A conductor moving at any angle to the lines of force cuts a number of lines of force proportional to the sine of the angles. Thus:

$$E = B \times L \times V \times \sin\theta \times 10^{-8}$$

where,

- L is the length of the conductor in centimeters,
- V is the velocity in centimeters per second of the conductor moving at an angle θ,
- B is the flux density.

The direction of the induced emf is in the direction in which the axis of a right-hand screw, when turned with the velocity vector, moves through the smallest angle toward the flux density vector. This is called the right-hand rule. (See Question 25.122.)

25.139 ***What is an electronic sound absorber?***—An electronic device developed by Dr. Harry F. Olson of RCA Acoustical Research Laboratory in 1954, with further development by J. C. Bleazy in 1962. Sound waves generated by noise are turned back against themselves and reduce the intensity of the original noise from 10 to 25 dB. Counter waveforms reduce the noise waveforms or cancel them entirely.

The electronic sound absorber consists of three elements: an electronic microphone (described in Question 4.69), an amplifier, and a specially designed speaker unit. In operation, the microphone picks up the noise and reacts instantaneously to pressure changes, and then translates these pressure changes into electrical impulses which are then amplified. The impulses emerge from the speaker directly behind the microphone as counter waves of equal and opposite pressure. The effect within a few feet of the speaker is a substantial leveling of the changes in air pressure —hence, a reduction or elimination of the oncoming sound.

25.140 ***How is the current-carrying capacity of heater wiring calculated for a printed circuit?***—The current-carrying capacity is based on the cross-sectional area as for wire. Fig. 25-140 shows graphically the current-carrying capacities for different widths and thicknesses of a printed circuit.

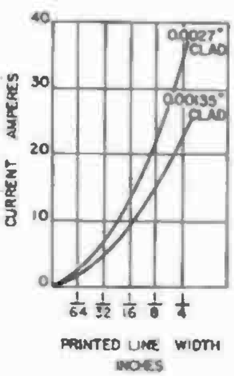


Fig. 25-140. Current-carrying capacity for printed circuits. Values are based on the resistance of printed wiring at 20°C for 100% conductivity copper.

25.141 ***What is a ferroresonant circuit?***—A circuit used in computers containing only resistance, capacitance, and inductance. It is designed to perform as an amplifier, trigger, switch, flip-flop, gate, or oscillator.

25.142 ***What is an electrometer?***—An advanced form of dc vacuum-tube voltmeter with an extremely high input impedance. The input impedance of an electrometer is usually around 10^{16} megohms, as compared with the input of 2 to 10 megohms of the conventional vacuum-tube voltmeter. Standard vacuum-tube voltmeters are not satisfactory for electrometer use because of the low input resistance between the control grid and cathode circuits.

An electrometer can be used for measuring currents of 10^{-16} microamperes and in circuits which require practically zero current drain.

25.143 ***What is a sound-powered telephone?***—A telephone transmitter which requires no external source of power. The acoustic waves impinging on the diaphragm of the transmitter cause a small armature attached to the diaphragm to move in a coil placed in a magnetic field. The movement of the armature generates currents which have the same frequency as the sound waves striking the diaphragm. These currents are sent over a wire to a receiver which is constructed in a similar manner.

The received currents cause the armature to move in the magnetic field. Moving the armature in the magnetic field causes acoustic waveforms to be generated which, in turn, are heard by the listener. Thus, a single unit may be used for both transmitting and receiving. A cross-section of a typical sound-powered telephone is shown in Fig. 25-143.

Sound-powered telephones are also constructed using the principle of a dynamic speaker, whereby a small voice coil is suspended from the diaphragm in a strong magnetic field. As in the armature type, a single unit is used for both transmitting and receiving. The average output voltage when transmitting is from 20 to 30 millivolts.

25.144 ***Give the nomenclature for identifying vacuum tubes.***—Tube types and numbers are divided into four

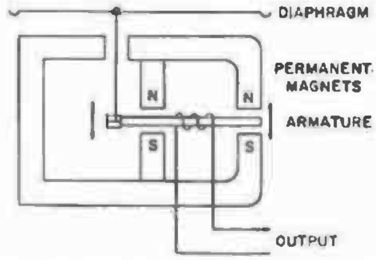


Fig. 25-143. Cross-sectional view of an armature-type, sound-powered telephone.

parts. The first part consists of one or more digits and designates the heater or filament voltage. Following the digit(s), one or more letters designate the type or function of the tube. Then, a number designates the number of elements. The final letter or letters designate the envelope size or construction.

A typical example is the 12AX7A tube, a high-mu twin diode requiring a heater voltage of 12.6 volts, and which is an amplifier with seven useful elements. The final letter designates those designed for military use, or a special base or construction.

Originally, rectifier tubes used only the letters U and Z; however, this is no longer true, as present-day tubes use no less than eleven different letters of the alphabet. The letter "G" indicates that the tube employs a glass envelope.

25.145 ***Describe the basic principles of electronic musical instruments.***—The musical engineer is confronted with two types of musical electronics: (1) musical amplification of the natural instrument, and (2) music created primarily with electronic circuitry. A third category might be considered a combination of the two, such as the amplification of the primary tones created by the instrument, with electronically added embellishments that were not present in the original produced tones.

Electronic music must approach a tonal quality pleasing to the human ear, with attention given to the frequency, intensity, waveform, and timing. Such music is often referred to as "synthetic music." Usually oscillators are used to provide the required frequencies in the form of a pure sine wave; harmonics are added to form a complex waveshape and produce timbre in the tone. Growth, duration, and decay time lend a natural quality to the tone. Variations in loudness provide the dynamics or expression to music. Portamento is required to supply a smooth gliding effect between tones and is achieved at various speeds and in various waveforms. Vibrato is a cyclic pulsating effect adding warmth and beauty to a tone. Rocking the finger on a string of an instrument produces this effect. Synthetically, it is produced by modulating either the frequency or amplitude. Tremolo is the rapid fluttering of a tone or chord without apparent break and is produced by a cyclic change in intensity.

The electronic carillon might well be used as a basic example of musical amplification. Here the operator makes use of a keyboard to actuate hammers (usually metal to similate bell tones) to strike the vibrating member. The resultant sound is electrostatically or electromagnetically picked, np amplified, and fed to speakers. Another form uses an electromagnetic mechanism to strike the vibrators. The amplifier may be in a remote position or near the keyboard, but it must be capable of producing considerable power to feed heavy-duty multiple speakers arranged for the desired distribution.

The theremin is an electronic instrument named for its inventor, Dr. Leon Theramin, and has tone qualities distinctive to itself. The device consists of two rf oscillators which operate similar to the beat-frequency oscillators described in Question 22.51. A free movement of the musician's hands in a space surrounding a tone rod changes the pitch and volume by altering the capacitance of a third oscillator whose output is rectified and used as a bias control on the amplifier feeding the speaker.

Amplifiers used for guitars may well be considered a combination of the natural and synthetic music, as they make use of the vibrations generated in the strings by the performer, and provide a means of adding tremelo, vibrato, and reverberation. Earlier amplifier systems usually consisted of a pickup mounted in contact with the bridge to pick up the vibrations; however, this arrangement often picked up unwanted sounds such as the rubbing of the instrument by clothing. Today magnetic pickups responding only to the motion of the metal strings in a magnetic field are used. A coil around a permanent magnet is placed under each string, and the physical motion of the string is that of a conductor moving in a magnetic field. Other models make use of a single, long, flat coil in a metal-shielded case clamped to the guitar. Where individual coils are used for each string, they may be connected in series or in parallel. The signal is fed to the preamplifier through a low-capacitance shielded coaxial line. High- and low-pass filters are used to alter the bass response. Tremolo may be accomplished by varying the bias on an amplifier stage with a low-frequency variable oscillator. Vibrato may be achieved electronically by means of phase-shift circuitry, but most often a mechanical lever action of the *tailpiece* of the guitar is used to alter the tension on the strings, thereby changing the frequency of the chord slightly. Both mechanical and time-delay networks are used to add reverberation. Power amplifiers and speakers complete the circuitry.

Electronic organs have come into wide use and are used frequently in studio production work. Several designs are used. One design makes use of a scalloped edge rotating in a flux supplied by a permanent magnet which is inserted in a coil. Undulations of the disc produce a change in flux, resulting in an alternating voltage. The output of this generator is then amplified and fed to a speaker. Oscillator circuits instead of a tone wheel are widely used to generate the tone. Other designs make use of air-driven reeds with electrostatic pickups, while yet another design employs electrostatic generators.

Fig. 25-145A. Electro-Voice Series-D electronic organ.

Of interest is the Electro-Voice Series D organ (Fig. 25-145A), which makes use of stators and scanners to reproduce the actual sounds created by the pipes of selected organs with regard to their timbre. This organ is based on the work of Dr. Jean Dereau, a French physicist, who developed an ingenious method of visually translating the complex waveforms to the stator plates. Twelve pairs of these plates are used on each organ generator. When the organist selects a tab and depresses a key, a single waveform area on one stator is electrostatically charged. The scanner operating in the charged area generates vibrations by changes in capacitance and when these vibrations are amplified accurately, they recreate the original sound of the organ, without embellish-

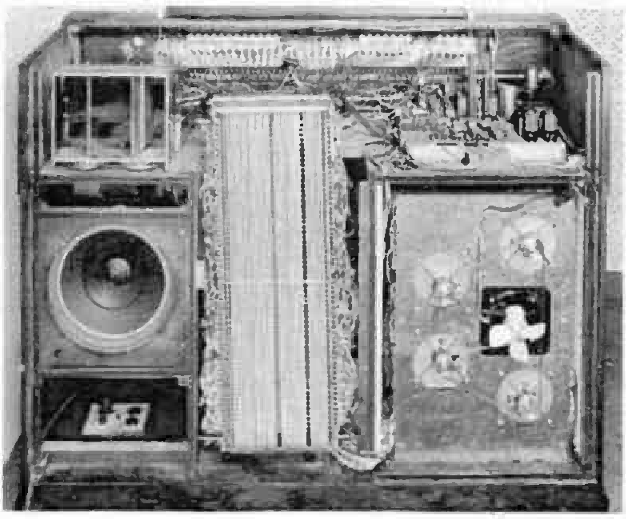


Fig. 25-145B. Rear interior view of Electro-Voice Series-D electronic organ.

ments or mixing. A rear view of the console is shown in Fig. 25-145B, and the tone generators are shown in Fig. 25-145C. A massive 30-inch dynamic speaker is used to realistically reproduce the 16 hertz of the 32-foot low C. Other speakers, midrange and very high frequency horns, provide reproduction beyond the range of audibility.

The electronic piano has no sound board, but does use the usual keyboard. String vibrations are converted into electrical voltages which are picked up and amplified in a manner similar to that of the guitar. These instruments are often used in studios by composers, who are required to work in relatively quiet areas, where the output may be tape recorded and played back over headphones or speakers for analysis and editing. A synthesizer, developed by Robert Moog, offers ultramodern facilities available to the composer, where multiple parts may be prerecorded and fed through a computer into a mixing center and reproduced for analysis of the composition. Much valuable time is conserved when the composer has such equipment available. However, due to cost of such facilities, few have been established.

Other amplification systems have been developed for such instruments as the reeds, horns, and percussions. Some of the problems have been to maintain the instrument in their natural form and not to create just another electronic device. Jean Selmer of Henri Selmer et Cie, Paris, experimented with the saxophone, and determined that the microphone should be placed within the saxophone, but found that the proper positioning within the bell of the instrument was critical. Working in conjunction with Electro-Voice, Jean Selmer developed a ceramic pressure-sensitive microphone along with its electronic circuitry, which added tremolo, reverberation, and a suboctave effect termed the "Octamatic." The resultant interplay of sounds approaches that of a duet.

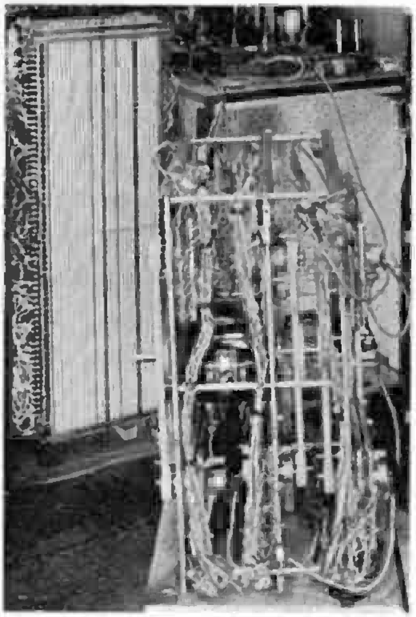


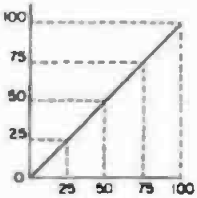
Fig. 25-145C. Rear interior view of Electro-Voice electronic organ with tone generators removed.

The amplifiers used for musical amplification must be capable of heavy power requirements, up to 100 watts of output. Speakers are selected for the ranges of the instruments involved, and some instruments may use multiple speaker systems.

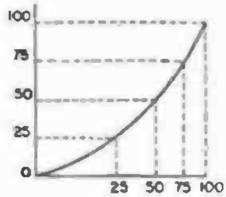
25.146 What is the difference between a linear and a nonlinear curve or response?—Three curves are shown in Fig. 25-146. Curve (a) is a linear curve; that is, if the input (the vertical figures) is increased 25 percent, the output (the figures along the bottom of the graph) increases in the same proportion. Such a relationship between the input and output of a device is said to be *linear* and appears as a straight line when plotted.

Curve (b) is a nonlinear curve. It will be noted this curve has no straight portion, but is constantly changing. For an increase of 25 percent at the input (vertical figures) the output voltage does not increase in the same proportion. This is also true for other values of input voltage.

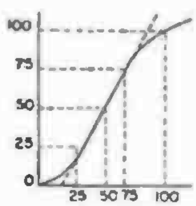
Curve (c) shows another type of nonlinear operation. It is similar to that of a vacuum-tube characteristic curve. It will be noted there is a curvature at both the top and the bottom and, also, that near the center the slope is fairly flat. A dotted line has been drawn in to indicate what would be called the straight-line portion of the curve. If this represents the characteristic curve of a vacuum tube, the negative bias voltage would be adjusted to center the operating point in the flattest portion of the center part of the curve. The tube would not be permitted to operate



(a) *Linear.*



(b) *Nonlinear.*



(c) *Nonlinear.*

Fig. 25-146. Linear and nonlinear curves.

beyond this portion. This would result in the 25- and 75-percent points being the same distance apart, producing a linear output voltage.

25.147 ***Describe a reed relay.***—Reed relays were developed by the Bell Telephone Laboratories in 1960 for use in the Bell System central offices. The device (Fig. 25-147A) consists of two magnetic reeds in a glass capsule with a nitrogen atmosphere. The glass envelope is surrounded by an electromagnetic coil connected to a control circuit. Although originally developed for the telephone company, such devices have found many uses in the electronics industry.

Fig. 25-147B shows the concept of the Ferreed relay in a simplified form. Windings around the glass envelope enclosed reed are arranged in such a way

Fig. 25-147A. Reed relay encased in an envelope of glass, filled with nitrogen. Contact capacitance is 0.1125 picofarad. (Courtesy, Bell Telephone Laboratories)

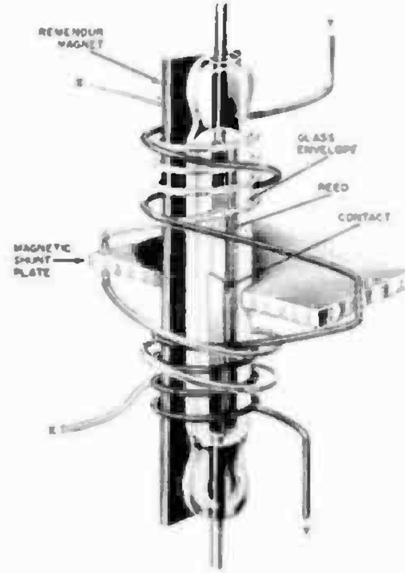


Fig. 25-147B. Bell Telephone Laboratories Ferreed relay, showing the coils and the Remendur magnet. The actual size of the relay is 1 × 0.150 inch. (Courtesy, Bell Telephone Laboratories Record)

that the contact is open or closed in response to pulses of current in the X and Y leads. For the closed state shown, simultaneous pulses in both the X and Y leads effectively cause the Remendur to become one magnet. The two reeds are now magnetically attracted and the contact closed. To open, a pulse is applied to either the X or the Y winding, and effectively divides the Remendur into two magnets at the magnetic shunt plate. Since the ends of the Remendur are both north (or both south) poles they repel each other and the contact is opened.

Remendur is a cobalt-iron-vanadium alloy, developed by Bell Telephone Laboratories, with a square hysteresis loop with values of coercive force between those of soft magnetic material and permanent magnet. The reed may be closed in ½ millisecond and has an intercapacitance of 0.1125 picofarads and an ac contact resistance of 50 milliohms. The advantages of such relays are their low internal capacitance, low contact resistance, and speed of operation.

Reed relays are also manufactured without the magnet and when the coils are energized, the reed snaps closed. The coil actuation is limited to low dc ripple current. However, an ac control voltage may be used in conjunction with a diode and filter capacitor. Depending on the design, the reeds will follow an ac coil input or dc pulses up to 400 Hz, provided the time on does not exceed 50 percent. Because of the fast action of the reed relays, they are ideal for use with semiconductor devices. Coils range from 1700 turns to 30,000 turns, with a dc resistance of 100 to 10,000 ohms. Operating voltages range from 6 to 110 volts. Up to three internal contacts are available; thus, single-pole, double-throw action can be obtained. The relays may be obtained with dry or wetted contacts and will withstand shock up to about 40 *g*'s, without false operation. Typical contact ratings are: 15 volt-amperes, 1-ampere switching, and 3 amperes carry.

25.148 Define the term "random noise."—It is the noise generated within a vacuum tube or transistor or by means of a random-noise generator. In a vacuum tube the noise increases somewhat uniformly with frequency. However, in a transistor the noise decreases with frequency up to a point and then rises again quite rapidly. However, today's transistors have noise factors that equal those of the quietest vacuum tube and are even less in some instances.

Random-noise generators as described in Question 22.56 are used in making tests on speakers, for the measurement of acoustic characteristics of auditoriums, and for many other different types of testing.

25.149 What is the mathematical relationship of rms, peak-to-peak, and average voltage?—The relationship is shown graphically in Fig. 25-149 and mathematically below:

rms $= 0.707 \times$ peak voltage
rms $= 1.11 \times$ average voltage
peak $= 1.414 \times$ rms voltage
peak $= 1.57 \times$ average voltage
average $= 0.637 \times$ peak voltage
average $= 0.9 \times$ rms voltage
peak-to-peak $=$ rms $\times 2.828$
rms $=$ peak-to-peak $\times 0.3535$

Effective voltage equals the rms voltage. Root mean square (rms) is defined as the square root of the average of the squares over a given angle.

25.150 What are the basic design principles used in dc relays and how are they classified?—Basically, there are two types of relays: ac and dc. A relay is an electrically operated switch connected to or actuated by a remote circuit. The relay causes a second circuit or group of circuits to operate. The relay may have any number of circuits connected to it. These circuits may consist of motors, bells, lights, etc., or the relay may be nsed to switch a number of other circuits.

Regardless of whether the relay is ac or dc, it will consist of an actuating coil, core, armature, and a group of contact springs which are connected to the circuit or circuits to be controlled. Associated with the armature are mechanical adjustments and springs. The mechanical arrangement of the contacts may be such that when the relay is at rest, certain circuits are either open or closed. If the circuits are closed when the relay is at rest, the relay is said to be normally closed. If the circuits are open, it is called a normally open relay.

The mechanical design of the contact springs is such that when the contacts are closed they slide for a short distance over the surfaces of each other before coming to rest. This is called a wiping contact and it ensures good electrical contact.

Telephone relays are wound in many different manners. Among them are the single-wound, double-wound, parallel-wound, tandem-wound, and noninductive-wound.

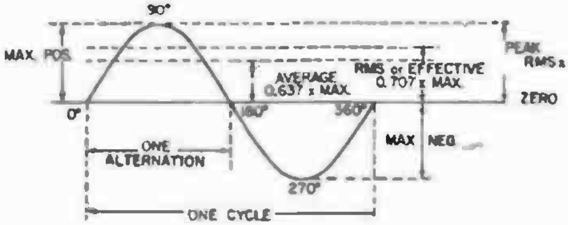


Fig. 25-149. Sine-wave terminology and voltage relations.

Relays differ in the amount of current and voltage required to operate them. Also, the dc resistance of the actuating coils may vary from a few ohms to several thousand ohms. In addition, they may be of many different types including the marginal, quick-operate, slow-operate, and polarized varieties.

A marginal relay is one which operates when the current through its winding reaches a specified value, and releases when the current falls to a given value. In the quick-operate type, the armature is attracted immediately to the pole piece of the electromagnet when the circuit is closed.

Slow-operate relays have a time-delay characteristic; the armature is not attracted immediately to the pole piece of the electromagnet when the circuit is closed. These relays employ a copper collar around the armature end of the pole piece and differ from the slow-release variety in that the latter type has the copper collar around the end of the pole piece opposite from the armature.

A polarized relay is designed to react to a given direction of current and magnitude. Polarized relays use a permanent-magnet core. Current in a given direction increases the magnetic field, and in the opposite direction it decreases the field. Thus, the relay will only operate for a given direction of current through the coil.

25.151 What are the basic principles of design in an alternating-current relay?—Alternating-current relays are similar in construction to the dc relays described in Question 25.150. Since alternating current has a zero value every half cycle, the magnetic field of an ac-operated relay will have corresponding zero values in the magnetic field every half cycle.

At and near the instants of zero current, the armature will leave the core, unless some provision is made to hold it in position. One method consists of using an armature of such mass that its inertia will hold it in position. Another method makes use of two windings on separate cores. These windings are connected so that their respective currents are out of phase with each other. Both coils effect a pull on the armature when current flows in both windings.

A third type employs a split pole piece of which one part is surrounded by a copper ring acting as a shorted turn. Alternating current in the actuating coil winding induces a current in the copper coil. This current is out of phase with the current in the actuating coil and does not reach the zero value at the same instant as the current in the actuating coil. As a result, there is always enough pull on the armature to hold it in the operating position.

An ac differential relay employs two windings exactly alike, except they are wound in opposite directions. Such relays operate only when one winding is energized. When both windings are energized in opposite directions, they produce an aiding magnetic field, since the windings are in opposite directions. When the current through the actuating coils is in the same direction, they produce opposite magnetic fields. If the current through the two coils is equal, the magnetic fields neutralize each other and the relay is nonoperative.

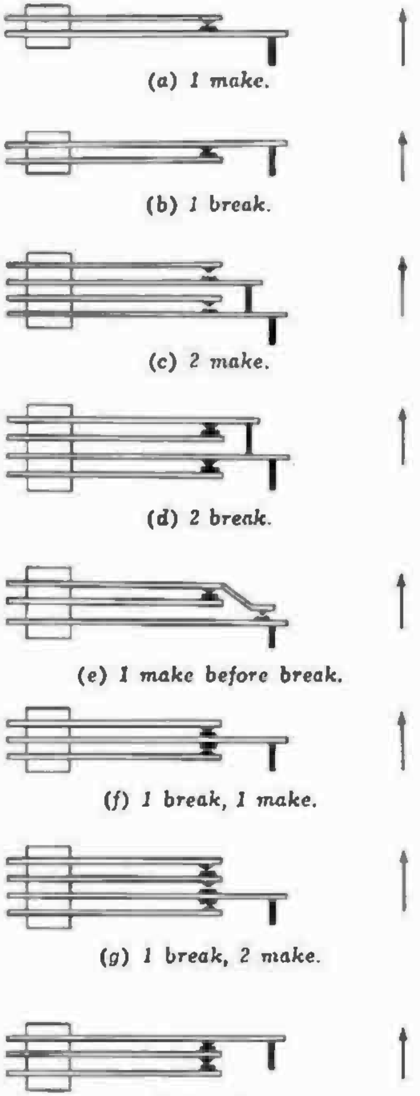
A differential polar relay employs a split magnetic circuit consisting of two windings on a permanent magnet core. In reality a differential polar relay is a combination of a differential and a polarized relay.

25.152 What are spring piles on a relay?—They are various combinations of contact springs making up the circuits which are operated by the action of the relay. Typical spring piles are shown in Fig. 25-152.

25.153 How are the actuating coils of a relay wound?—The different types of windings employed in relays are shown in Fig. 25-153.

25.154 What is a bimetal thermostat control?—A strip consisting of two dissimilar metals welded together. Because the metals selected have different temperature coefficients of expansion, they are caused to bend with changes of temperature.

Fig. 25-154A shows a bimetal thermostatic control manufactured by Thomas A. Edison, Inc. A typical application is the control of temperature in a crystal oscillator oven where the frequency of oscillation must be maintained to within five parts in ten million. In such applications, the temperature must be held to within 0.20 degree centigrade to permit frequency readings over long periods without recalibration. The interior



(a) *1 make.*

(b) *1 break.*

(c) *2 make.*

(d) *2 break.*

(e) *1 make before break.*

(f) *1 break, 1 make.*

(g) *1 break, 2 make.*

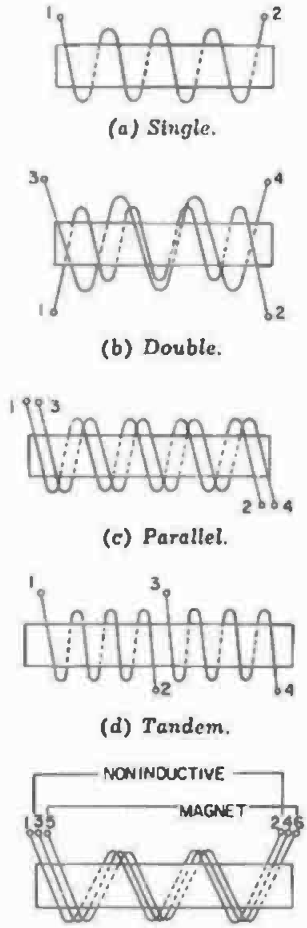
(h) *Break 2.*

Fig. 25-152. Relay contact spring piles.

construction of a typical thermostatic control is shown in the illustration.

Basically, the device is a single-pole, single-throw switch incorporating a bimetallic member actuated by thermal energy. To obtain operating stability and accuracy the complete structure is enclosed in a glass envelope containing a protective gas. The slow-make, slow-break principle of design results in small operating differentials that approach zero under no-load conditions. The gas atmosphere minimizes the effect of contact arcing under heavy current loads. The unit is calibrated to the final operating temperature setting by external means, permitting the adjustment while the element is sealed in the glass envelope.

The free end of the heat-sensitive bimetal deflects in the direction which separates the normally closed contacts as the temperature rises. After the stabilizing adjustment has been made, the



(a) *Single.*

(b) *Double.*

(c) *Parallel.*

(d) *Tandem.*

(e) *Noninductive.*

Fig. 25-153. Relay-coil windings.

bimetal will always assume a definite deflection for a given temperature. The temperature at which the contacts will open depends on the position of the calibration screw. By reversing the position of the bimetal in the assembly, the control can be made to close the external circuit as the temperature rises. The bimetal, because of its relatively long arm, is highly sensitive to small temperature changes.

A typical thermostatically controlled oven circuit is shown at (a) in Fig. 25-154B, and a typical thermoelectric

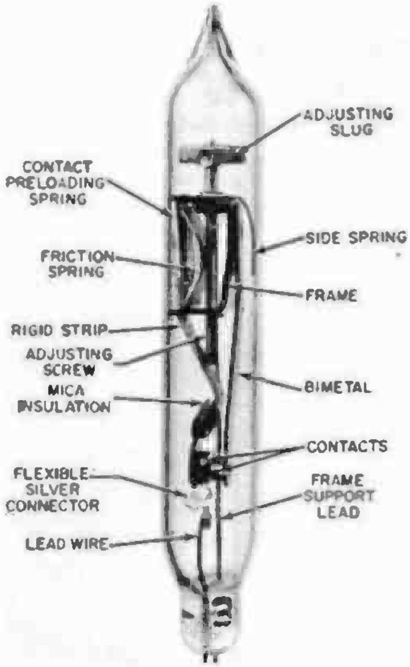


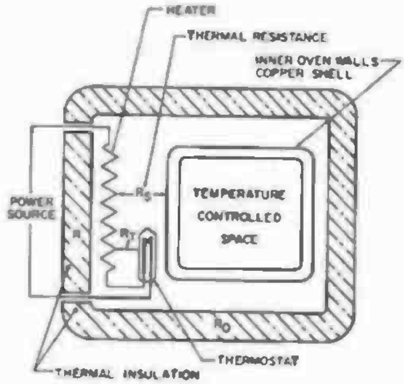
Fig. 25-154A. View of a Thomas A. Edison Instrument Division Model S1-1A thermostatic relay.

oven cycle is given at **(b)** in Fig. 25-154B.

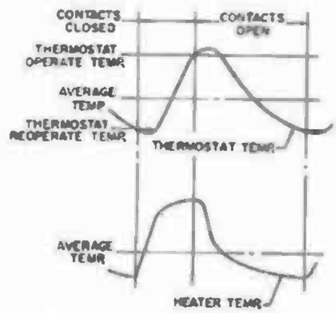
25.155 What is a thermal time-delay relay and how is it constructed?—A relay with a thermal unit so designed that a delay period of from 2 seconds to 5 minutes may be induced in an electrical circuit. Fig. 25-155A shows a thermal delay relay manufactured by Thomas A. Edison, Inc. It consists of two bimetallic strips, each rigidly supported at one end and carrying a contact at the other. An electrical heater is wound around the primary bimetal member, which is deflected by the application of heater current to mate with the contact mounted on the other bimetal strip. The use of the second bimetal strip which supports the preloaded spring contact provides compensation for ambient temperature changes from minus 60 to plus 85 degrees centigrade.

One contact is mounted on a spring attached to the compensating bimetal strip. This spring is restrained until the primary bimetal is deflected sufficiently to cause contact. After the contacts close the spring lifts from its restraining support and applies force to hold the contacts firmly together.

Factors controlling the deflection of the bimetal are stable and the original setting of the contacts permanently fixes the timing characteristics of the relay. The structure of the elements is such that the contacts will not close under vibration (1/16 inch at 55 Hz) with the heater unexcited, nor will the contacts open under the same condition at saturation temperature.



(a) Circuit for operation with a crystal oven.



(b) Oven operating cycle.

Fig. 25-154B. Circuit and operating characteristics for the Thomas A. Edison, Inc. thermostatic relay.

Applications of this device include the control of plate voltage to large vacuum tubes after the heaters have reached the proper operating temperature, overcurrent protection, overvoltage protection, motor starting, holdover circuits, and the integration of pulses or intermittent current into accumulated heat energy which will operate its contacts after a predetermined rate of input has been reached. The operating characteristics of this device are given in Fig. 25-155B, and typical operating circuits are shown in Fig. 25-155C.

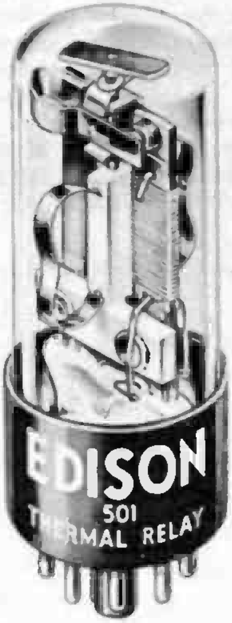
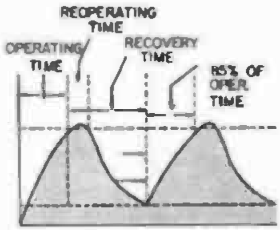


Fig. 25-155A. Interior view of a Thomas A. Edison, Instrument Division Model 501 thermal time-delay relay.

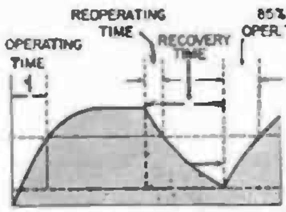
25.156 Describe the construction and operation of a sensitive direct-current relay.—A sensitive dc relay, manufactured by Thomas A. Edison, Inc., designed to operate directly from a thermocouple or photocell output, is shown in Fig. 25-156. It may also be used in the grid or plate circuit of a vacuum tube, or as a polarized relay for use as a null detector in a servo bridge circuit.

The relay mechanism is a fixed-coil, moving-magnet type having the characteristics of a d'Arsonval meter movement. The fixed coils may be obtained with resistance values between 0.5 and 24,000 ohms. The rotor shaft carries a small alnico V-shaped magnet vane at the lower end and a moving contact at the upper end. The rotor assembly is pivoted on jeweled bearings so that the magnet vane rotates between the stationary coils. Because of the small size and small mass of the rotor assembly, the relay may be used in applications having high shock values.

The contact circuit is completed through a hairspring that also provides counter torque or restoring force. Counterweights on a crossarm provide rotor balance to permit operation in any plane and to minimize the effects of vibration. The magnetic field is shielded

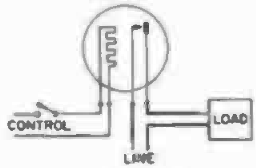


(a) Operation when the heater voltage is removed immediately after contact.

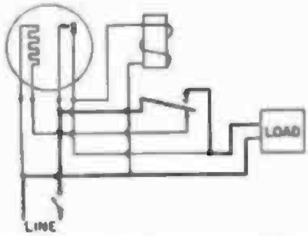


(b) Operation when the heater voltage is maintained until the bimetals reach saturation temperature.

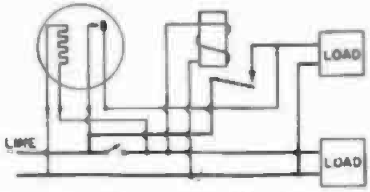
Fig. 25-155B. Characteristics of the Thomas A. Edison thermal time-delay relay.



(a) As a relay.



(b) As a slave unit to provide rapid resetting.



(c) Delayed opening of load circuit.

Fig. 25-155C. Circuit connections for the Thomas A. Edison thermal time-delay relay.

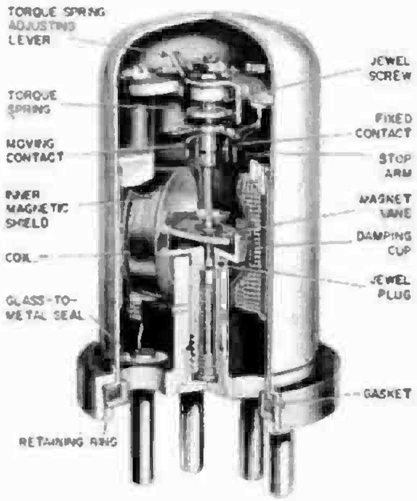


Fig. 25-156. Interior view of a Thomas A. Edison, Instrument Division Model 501-S1-1A sensitive dc relay.

from external fields by a high-permeability shield and by a second external shield cover. This shielding also provides a return for the magnetic flux.

Such relays are calibrated in terms of current to avoid variations induced by temperature. The relay is a current-sensitive instrument operating on the same principle as a tangent galvanometer. In the instrument shown, if the operating current is changed gradually, the moving contact will follow the rate of change until it touches one of the stationary contacts. The speed of operation depends on the rate of change of the operating current and its magnitude. As a rule, initial contact is made in approximately 150 milliseconds when the rated current is instantaneously applied. Increasing the operating current increases the speed of contact closure exponentially. At ten times the nominal closing current, contact is made in approximately 40 milliseconds. The normal deflection is 45 degrees; however, by presetting the angular position of the fixed contact, the operating power can be varied from 0 to 70 microwatts.

25.157 ***What are the prefixes used for metric units?***—The standard symbols adopted by the Institute of Electrical and Electronic Engineers (IEEE) for metric units are given below.

Numerical Value	*Prefix*	*Symbol*
10^{12}	tera	T
10^{9}	giga	G
10^{6}	mega	M
10^{4}	myria	My
10^{3}	kilo	k
10^{2}	hecto	h
10	deka	da
10^{-1}	deci	d
10^{-2}	centi	c
10^{-3}	milli	m
10^{-6}	micro	μ
10^{-9}	nano	n
10^{-12}	pico	p
10^{-15}	femto	f
10^{-18}	atto	a

Examples of usage:

decibel	dB
gigahertz	GHz
hertz (cps)	Hz
kilohertz	kHz
picofarad ($\mu\mu$F)	pF

For further use of symbols and adoption, IEEE Standard Symbols for Units, No. 260, should be consulted.

The metric conversion table (Fig. 25-157) provides a fast and easy method of convertiug from one metric notation to another, including the preceding designated units. The table aids in restating a value. After locating the prefix in the Original Value column, it may be converted to another desired value of different prefix. The number indicates the number of places the decimal is to be

ORIGINAL VALUE	DESIRED VALUE													
	Tera-	Giga-	Mega-	Myria-	Kilo-	Hekto-	Deka-	Units	Deci-	Centi-	Milli-	Micro-	Nano-	Pico-
Tera-		3→	6→	8→	9→	10→	11→	12→	13→	14→	15→	18→	21→	24→
Giga-	←3		3→	5→	6→	7→	8→	9→	10→	11→	12→	15→	18→	21→
Mega-	←6	←3		2→	3→	4→	5→	6→	7→	8→	9→	12→	15→	18→
Myria-	←8	←5	←2		1→	2→	3→	4→	5→	6→	7→	10→	13→	16→
Kilo-	←9	←6	←3	←1		1→	2→	3→	4→	5→	6→	9→	12→	15→
Hekto-	←10	←7	←4	←2	←1		1→	2→	3→	4→	5→	8→	11→	14→
Deka-	←11	←8	←5	←3	←2	←1		1→	2→	3→	4→	7→	10→	13→
Units	←12	←9	←6	←4	←3	←2	←1		1→	2→	3→	6→	9→	12→
Deci-	←13	←10	←7	←5	←4	←3	←2	←1		1→	2→	5→	8→	11→
Centi-	←14	←11	←8	←6	←5	←4	←3	←2	←1		1→	4→	7→	10→
Milli-	←15	←12	←9	←7	←6	←5	←4	←3	←2	←1		3→	6→	9→
Micro-	←18	←15	←12	←10	←9	←8	←7	←6	←5	←4	←3		3→	6→
Nano-	←21	←18	←15	←13	←12	←11	←10	←9	←8	←7	←6	←3		3→
Pico-	←24	←21	←18	←16	←15	←14	←13	←12	←11	←10	←9	←6	←3	

Fig. 25-157. Metric conversion table.

moved, and the arrow indicates in which direction. As an example, if 0.15 kilowatts is to be converted to watts, taking kilo from the Original Value column aud following the horizontal line to the unit column, read a 3 with the arrow pointing right. Therefore the decimal is to be moved right three places, and it may now read as 150 watts. If 4500 kilohertz is to be converted to megahertz, read kilo in the left-hand column and under mega read 3 with a left-pointing arrow. The decimal is shifted three points in that direction, and 4500 kHz is equivalent to 4.5 MHz.

25.158 What is a guard circuit and where is it used?—A guard circuit is one which is used to remove the effects of stray leakage and capacitance or any other stray parameter that might affect the measurement of the actual value of the element under measurement. Guard circuits are used with many different types of electrical measuring circuits and equipment, particularly with capacitance bridges. (See Question 25.116.)

25.159 Describe a latching relay.—It is a relay containing two separate actuating coils (Fig. 25-159). Actuating one coil latches the relay in one position, where it remains until it is unlatched by energizing the other coil. Relays of this design are used frequently in sound installations, such as in the virgin looping system described in Question 17.227.

25.160 Give a table showing the relationship of density to percent of transmission or reflection.—In the process of recording on photographic film, it is necessary to know the percent of transmission relative to a given density. To simplify the computation, the table of density related to percent transmission or reflection is given in Fig. 25-160. It will be observed that for a density of 1.50 (used in optical film recording), the transmission is 3.162, or 3.162 percent transmission or reflection; for a density of 4.0 the transmission is 0.01 percent.

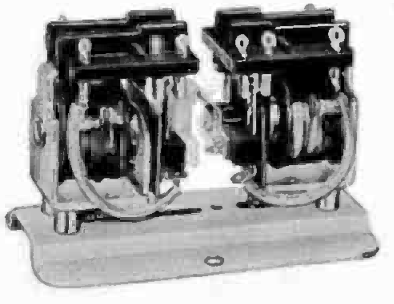


Fig. 25-159. Latching relay manufactured by Potter and Brumfield (Canada).

$$\text{Density} = \text{Log}_{10}\frac{1}{T}$$

where,

T is the percent transmission.

25.161 What is the relationship of footage and running time for 16-mm motion picture film?—This relationship is given in Fig. 25-161.

25.162 What is the relationship between 16-mm and 35-mm motion-picture film footage?—The relationship is 2.5:1; that is, 400 feet of 16-mm film running at 36 feet per minute equals 1000 feet of 35-mm film running at 90 feet per minute. (See Fig. 25-162.)

25.163 Show a wire table for wire sizes between 1 and 44, with the circular-mil area and current-carrying capacities.—Such a table is given in Fig. 25-163.

25.164 Show graphically the relationship between farads and microfarads.—Such a graph is shown in Fig. 25-164.

25.165 Show graphically the relationship between microfarads and picofarads.—Such a graph is given in Fig. 25-165. The term *micromicrofarads* has now been replaced by the term *picofarads*.

25.166 What is electronic gating?—A vacuum-tube amplifier that is cut off and on electronically by energizing the control grid or cathode circuit by the application of a square wave. When so actuated by the external-control source voltage, the output of the gating tube will pass a signal at the rate at which the control circuit operates. This system is used in computers and electronic control equipment.

25.167 What is decibel notation and how is it used?—Decibel notation is a method of equating power levels involving the transmission of electrical energy in the audio- and radio-frequency bands. To measure the acoustic or electrical power associated with the generation or transmission of audio frequencies, special instruments and measuring techniques are required. In a commercial power transmission system at 60 Hz, little of the energy is lost because of the relatively low fre-

Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection
0.00	100.00								
0.01	97.72	0.41	38.90	0.81	15.49	1.21	6.166	1.61	2.455
0.02	95.50	0.42	38.02	0.82	15.14	1.22	6.026	1.62	2.399
0.03	93.33	0.43	37.15	0.83	14.79	1.23	5.888	1.63	2.344
0.04	91.20	0.44	36.31	0.84	14.45	1.24	5.754	1.64	2.291
0.05	89.13	0.45	35.48	0.85	14.13	1.25	5.623	1.65	2.239
0.06	87.10	0.46	34.67	0.86	13.80	1.26	5.495	1.66	2.188
0.07	85.11	0.47	33.88	0.87	13.49	1.27	5.370	1.67	2.138
0.08	83.18	0.48	33.11	0.88	13.18	1.28	5.248	1.68	2.089
0.09	81.28	0.49	32.36	0.89	12.88	1.29	5.129	1.69	2.042
0.10	79.43	0.50	31.62	0.90	12.59	1.30	5.012	1.70	1.995
0.11	77.62	0.51	30.90	0.91	12.30	1.31	4.898	1.71	1.950
0.12	75.86	0.52	30.20	0.92	12.02	1.32	4.786	1.72	1.905
0.13	74.13	0.53	29.51	0.93	11.75	1.33	4.677	1.73	1.862
0.14	72.44	0.54	28.84	0.94	11.48	1.34	4.571	1.74	1.820
0.15	70.79	0.55	28.18	0.95	11.22	1.35	4.467	1.75	1.778
0.16	69.18	0.56	27.54	0.96	10.96	1.36	4.365	1.76	1.738
0.17	67.61	0.57	26.92	0.97	10.72	1.37	4.266	1.77	1.698
0.18	66.07	0.58	26.30	0.98	10.47	1.38	4.169	1.78	1.660
0.19	64.57	0.59	25.70	0.99	10.23	1.39	4.074	1.79	1.622
0.20	63.10	0.60	25.12	1.00	10.00	1.40	3.981	1.80	1.585
0.21	61.66	0.61	24.55	1.01	9.772	1.41	3.890	1.81	1.549
0.22	60.26	0.62	23.90	1.02	9.550	1.42	3.802	1.82	1.514
0.23	58.88	0.63	23.44	1.03	9.333	1.43	3.715	1.83	1.479
0.24	57.54	0.64	22.91	1.04	9.120	1.44	3.631	1.84	1.445
0.25	56.23	0.65	22.39	1.05	8.913	1.45	3.548	1.85	1.413
0.26	54.95	0.66	21.88	1.06	8.710	1.46	3.467	1.86	1.380
0.27	53.70	0.67	21.38	1.07	8.511	1.47	3.388	1.87	1.349
0.28	52.48	0.68	20.89	1.08	8.318	1.48	3.311	1.88	1.318
0.29	51.29	0.69	20.42	1.09	8.128	1.49	3.236	1.89	1.288
0.30	50.12	0.70	19.95	1.10	7.943	1.50	3.162	1.90	1.259
0.31	48.98	0.71	19.50	1.11	7.762	1.51	3.090	1.91	1.230
0.32	47.86	0.72	19.05	1.12	7.586	1.52	3.020	1.92	1.202
0.33	46.77	0.73	18.62	1.13	7.413	1.53	2.951	1.93	1.175
0.34	45.71	0.74	18.20	1.14	7.244	1.54	2.884	1.94	1.148
0.35	44.67	0.75	17.78	1.15	7.079	1.55	2.818	1.95	1.122
0.36	43.65	0.76	17.38	1.16	6.918	1.56	2.754	1.96	1.096
0.37	42.66	0.77	16.98	1.17	6.761	1.57	2.692	1.97	1.072
0.38	41.69	0.78	16.60	1.18	6.607	1.58	2.630	1.98	1.047
0.39	40.74	0.79	16.22	1.19	6.457	1.59	2.570	1.99	1.023
0.40	39.81	0.80	15.85	1.20	6.310	1.60	2.512	2.00	1.000

Fig. 25-160. Conversion table. Optical density versus

Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection	Density	Percent Transmission or Reflection
2.01	0.9772	2.41	0.3890	2.81	0.1549	3.21	0.0617	3.61	0.0245
2.02	0.9550	2.42	0.3802	2.82	0.1514	3.22	0.0603	3.62	0.0239
2.03	0.9333	2.43	0.3715	2.83	0.1479	3.23	0.0588	3.63	0.0234
2.04	0.9120	2.44	0.3631	2.84	0.1445	3.24	0.0575	3.64	0.0229
2.05	0.8913	2.45	0.3548	2.85	0.1413	3.25	0.0562	3.65	0.0223
2.06	0.8710	2.46	0.3467	2.86	0.1380	3.26	0.0550	3.66	0.0218
2.07	0.8511	2.47	0.3388	2.87	0.1349	3.27	0.0537	3.67	0.0214
2.08	0.8318	2.48	0.3311	2.88	0.1318	3.28	0.0529	3.68	0.0218
2.09	0.8128	2.49	0.3236	2.89	0.1288	3.29	0.0513	3.69	0.0204
2.10	0.7943	2.50	0.3162	2.90	0.1259	3.30	0.0501	3.70	0.0199
2.11	0.7762	2.51	0.3090	2.91	0.1230	3.31	0.0489	3.71	0.0195
2.12	0.7586	2.52	0.3020	2.92	0.1202	3.32	0.0478	3.72	0.0190
2.13	0.7413	2.53	0.2951	2.93	0.1175	3.33	0.0468	3.73	0.0186
2.14	0.7244	2.54	0.2884	2.94	0.1148	3.34	0.0457	3.74	0.0182
2.15	0.7079	2.55	0.2818	2.95	0.1122	3.35	0.0446	3.75	0.0178
2.16	0.6918	2.56	0.2754	2.96	0.1096	3.36	0.0436	3.76	0.0174
2.17	0.6761	2.57	0.2692	2.97	0.1072	3.37	0.0426	3.77	0.0169
2.18	0.6607	2.58	0.2630	2.98	0.1047	3.38	0.0417	3.78	0.0166
2.19	0.6457	2.59	0.2570	2.99	0.1023	3.39	0.0407	3.79	0.0162
2.20	0.6310	2.60	0.2512	3.00	0.1000	3.40	0.0398	3.80	0.0158
2.21	0.6166	2.61	0.2455	3.01	0.0977	3.41	0.0389	3.81	0.0155
2.22	0.6026	2.62	0.2399	3.02	0.0955	3.42	0.0380	3.82	0.0152
2.23	0.5888	2.63	0.2344	3.03	0.0933	3.43	0.0371	3.83	0.0148
2.24	0.5754	2.64	0.2291	3.04	0.0912	3.44	0.0363	3.84	0.0145
2.25	0.5623	2.65	0.2239	3.05	0.0891	3.45	0.0355	3.85	0.0141
2.26	0.5495	2.66	0.2188	3.06	0.0871	3.46	0.0347	3.86	0.0138
2.27	0.5370	2.67	0.2138	3.07	0.0851	3.47	0.0339	3.87	0.0135
2.28	0.5248	2.68	0.2089	3.08	0.0832	3.48	0.0331	3.88	0.0132
2.29	0.5129	2.69	0.2042	3.09	0.0813	3.49	0.0324	3.89	0.0129
2.30	0.5012	2.70	0.1995	3.10	0.0794	3.50	0.0316	3.90	0.0126
2.31	0.4898	2.71	0.1950	3.11	0.0776	3.51	0.0309	3.91	0.0123
2.32	0.4786	2.72	0.1905	3.12	0.0759	3.52	0.0302	3.92	0.0120
2.33	0.4677	2.73	0.1862	3.13	0.0741	3.53	0.0295	3.93	0.0117
2.34	0.4571	2.74	0.1820	3.14	0.0724	3.54	0.0288	3.94	0.0114
2.35	0.4467	2.75	0.1778	3.15	0.0708	3.55	0.0289	3.95	0.0112
2.36	0.4365	2.76	0.1738	3.16	0.0692	3.56	0.0275	3.96	0.0109
2.37	0.4266	2.77	0.1698	3.17	0.0676	3.57	0.0269	3.97	0.0107
2.38	0.4169	2.78	0.1660	3.18	0.0661	3.58	0.0263	3.98	0.0105
2.39	0.4074	2.79	0.1622	3.19	0.0646	3.59	0.0257	3.99	0.0102
2.40	0.3981	2.80	0.1585	3.20	0.0631	3.60	0.0251	4.00	0.0100

percent transmission or reflection.

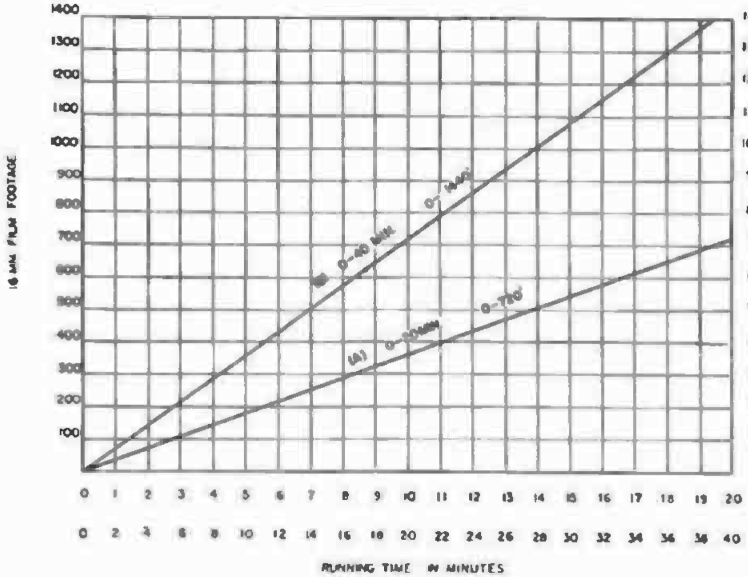


Fig. 25-161. Time versus footage for 16-mm film.

quency. However, power used for voice and music transmission over the same distance will suffer a considerable loss because of the higher frequencies involved and the complexity of the waveforms. The longer the transmission line, the greater are the power losses which must be compensated for during transmission.

In the early days of telephone and radio transmission over telephone lines, a unit of measurement called "miles of loss" was used. This expressed the transmission loss of the line in terms of a standard-loop mile of No. 19 telephone wire. (A loop mile is, in reality, two miles of wire, one going and one returning.) Later this term was replaced with a logarithmic term called the "transmission unit." Some years later, this term was changed to "bel" in honor of Alexander Graham Bell. The unit bel being too large for practical use, it was changed to decibel or one-tenth of a bel, abbreviated dB. The fundamental equation for the bel is:

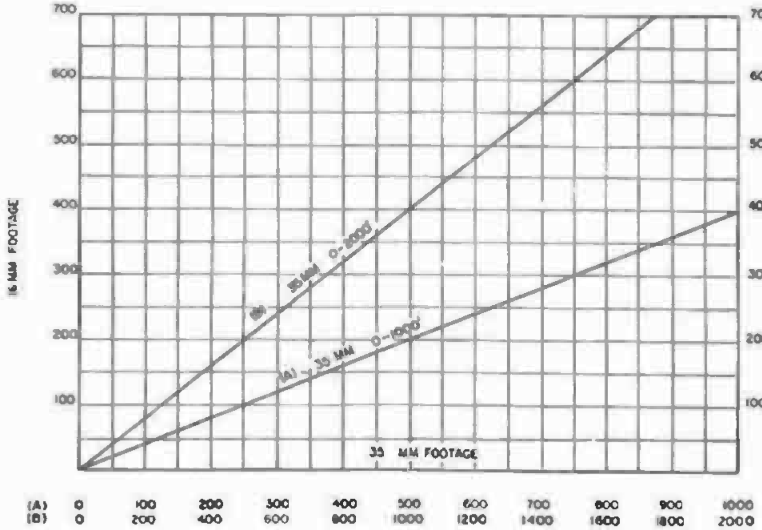


Fig. 25-162. 35-mm to 16-mm footage conversion chart.

$$\text{bel} = \text{Log}_{10}\frac{P_1}{P_2}$$

Changing to decibels, the equation becomes:

$$dB = 10\,\text{Log}_{10}\frac{P_1}{P_2}$$

The decibel has no actual numerical value but is used only to express a ratio between two powers, voltages, or currents. The decibel may also be used to express ratios dealing with pressure, impedance, and other basic units.

It is customary, when using decibels, to employ a reference voltage or power level to which the measurement is referred. The standard reference now in use by the broadcasting, recording, and electronics industries is 1 milliwatt and is written "0 dBm" (0.001 watt). Whenever the letters dBm appear behind the number 0, it means that the reference power is 1 milliwatt. Therefore, when such a reference level is indicated, the power in parenthesis may be left off. The 1-milliwatt reference power was adopted by the electronic industry in 1939 and represents 1 milliwatt of power in a 600-ohm circuit which has a sine-wave rms value of 0.773 volt across it.

Several different power-reference levels have been used over the past years. They are: 6, 10, 12.5, and 50 milliwatts. These have been replaced with the present reference power level of 1 milliwatt (600 ohms). Previously the most common reference power level was 6 milliwatts. This level is still used in some types of equipment used for purposes other thau recording and broadcasting. It is based on 6 milliwatts of power in a 500-ohm circuit, with an rms sine-wave voltage of 1.73 volts. In 1939 it was agreed by manufacturers of electronic equipment that all future equipment would be rated with reference to 1 milliwatt.

Because of the difference in the power of 1 milliwatt and 6 milliwatts, a conversion factor must be used when converting or referring from one reference level to the other. The difference in decibels between the two reference powers is 7.78 dB. This may be found as follows:

$$dB = 10\,\text{Log}_{10}\frac{6}{1}$$
$$= 10 \times 0.778$$
$$= 7.78\ dB$$

Thus, when a piece of equipment is rated for a given amount of power, the numerical difference in decibels for the 1-milliwatt reference level will be 7.78 dB higher. A typical example of this is an amplifier stated to have a power output of 20 watts. The output level in decibels for 20 watts, using the 6-milliwatt reference level, is a plus 35.26 dB, while for the 1-milliwatt reference power, the level in decibels is a plus 43 dBm. This may appear confusing because both decibel ratings are for the same amount of power output. Therefore, it is mandatory that the reference level be stated when rating an amplifier in decibels only.

The transmission loss or gain of a circuit may be expressed either in power or in voltage. When expressed in power, the following equation is used:

$$dB = 10\,\text{Log}_{10}\frac{P_1}{P_2}$$

If expressed in current or voltage, use the equation:

$$dB = 20\,\text{Log}_{10}\frac{V_1}{V_2} \text{ or } \frac{V_2}{V_1} \text{ or } \frac{I_1}{I_2} \text{ or } \frac{I_2}{I_1}$$

where,

V_1, V_2, and I_2 are the input and output voltages or currents, respectively.

As an example, what is the loss in decibels for an attenuator which has 6 volts at its input and 2 volts at the output, the input and output impedances being the same and properly terminated? Using the equation:

$$dB = 20\,\text{Log}_{10}\frac{6}{2} = 20 \times 0.477 = -9.54\ dB.$$

Knowing that the device presents a loss and not a gain, a minus sign (−) is placed before the number value in decibels. If the device is an amplifier and represents a gain, no sign is used. The answer is simply stated 9.54 dB.

For unequal input and output impedances, the equation becomes:

$$dB = 20\,\text{Log}_{10}\frac{E_1}{E_2} + 10\,\text{Log}_{10}\frac{Z_1}{Z_2}$$

where,

Z_1 and Z_2 are the input and output impedances.

For convenience of computation, E_1 and E_2, or Z_1 and Z_2, may be inverted, with the larger value placed above.

In Section 23 many different equations and methods of using decibels are

A.W.G. No. (B.&S.)	Diam. in mils.*	Diam. in mm.	Cir. mils	Cross-sectional area Sq. Inches	Sq. mm.	Turns per linear inch** D.S.C. or D.C.C.	S.C.C.	Enamel	S.S.C.
1	289.3	7.348	83690	.06573	42.41	—	—	—	—
2	257.6	6.544	66370	.05213	33.63	—	—	—	—
3	229.4	5.827	52640	.04134	26.67	—	—	—	—
4	204.3	5.189	41740	.03278	21.15	—	—	—	—
5	181.9	4.621	33100	.02600	16.77	—	—	—	—
6	162.0	4.115	26250	.02062	13.3	—	—	—	—
7	144.3	3.665	20820	.01635	10.55	—	—	—	—
8	128.5	3.264	16510	.01297	8.36	7.1	7.4	7.6	—
9	114.4	2.906	13090	.01028	6.63	7.8	8.2	8.6	—
10	101.9	2.588	10380	.008155	5.26	8.9	9.3	9.6	—
11	90.74	2.305	8234	.006467	4.17	9.8	10.3	10.7	—
12	80.81	2.053	6530	.005129	3.31	10.9	11.5	12.0	—
13	71.96	1.828	5178	.004067	2.62	12.0	12.8	13.5	—
14	64.08	1.628	4107	.003225	2.08	13.3	14.2	15.0	—
15	57.07	1.450	3257	.002558	1.65	14.7	15.8	16.8	—
16	50.82	1.291	2583	.002028	1.31	16.4	17.9	18.9	18.9
17	45.26	1.150	2048	.001609	1.04	18.1	19.9	21.2	21.2
18	40.30	1.024	1624	.001276	.82	19.8	22.0	23.6	23.6
19	35.89	.9116	1288	.001012	.65	21.8	24.4	26.4	26.4
20	31.96	.8118	1022	.0008023	.52	23.8	27.0	29.4	29.4
21	28.46	.7230	810.1	.0006363	.41	26.0	29.8	33.1	32.7
22	25.35	.6438	642.4	.0005046	.33	30.0	34.1	37.0	36.5
23	22.57	.5733	509.5	.0004002	.26	31.6	37.6	41.3	40.6
24	20.10	.5106	404.0	.0003173	.20	35.6	41.5	46.3	45.3
25	17.90	.4547	320.4	.0002517	.16	38.6	45.6	51.7	50.4
26	15.94	.4049	254.1	.0001996	.13	41.8	50.2	58.0	55.6
27	14.20	.3606	201.5	.0001583	.10	45.0	55.0	64.9	61.5
28	12.64	.3211	159.8	.0001255	.08	48.5	60.2	72.7	68.6
29	11.26	.2859	126.7	.00009953	.064	51.8	65.4	81.6	74.8
30	10.03	.2546	100.5	.00007894	.051	55.5	71.5	90.5	83.3
31	8.928	.2268	79.70	.00006260	.040	59.2	77.5	101.	92.0
32	7.950	.2019	63.21	.00004964	.032	62.6	83.6	113.	101.
33	7.080	.1798	50.13	.00003937	.0254	66.3	90.3	127.	110.
34	6.305	.1601	39.75	.00003122	.0201	70.0	97.0	143.	120.
35	5.615	.1426	31.52	.00002476	.0159	73.5	104.	158.	132.
36	5.000	.1270	25.00	.00001964	.0127	77.0	111.	175.	143.
37	4.453	.1131	19.83	.00001557	.0100	80.3	118.	198.	154.
38	3.965	.1007	15.72	.00001235	.0079	83.6	126.	224.	166.
39	3.531	.0897	12.47	.000009793	.0063	86.6	133.	248.	181.
40	3.134	.0799	9.888	.000007766	.0050	89.7	140.	282.	194.
41	2.75	.0711	7.841	.000006160	.0040	—	—	—	—
42	2.50	.0633	6.220	.000004885	.0032	—	—	—	—
43	2.25	.0564	4.933	.000003873	.0025	—	—	—	—
44	2.00	.0502	3.910	.000003073	.0020	—	—	—	—

*A mil is 1/1000 inch.

Fig. 25-163. Wire table for sizes between

discussed. Conversion tables and other pertinent data are given in Section 10. Several different equations that are common in decibel notation appear in Fig. 25-39.

25.168 How are telephone lines classified?—Open-wire, field-wire, and cable.

25.169 What are open-wire lines?—Single bare-wire conductors sup-

Turns per Sq. In.**			Feet per pound				Current-carrying capacity (amperes)		
S.C.C.	Enamel	D.C.C.	D.C.C.	S.C.C.	Bare	Ohms per 1000 ft.	At 1000 C.M. per amp	At 1500 C.M. per amp	Nearest British S.W.G.
—	—	—	—	—	3.947	.1260	83.7	55.7	1
—	—	—	—	—	4.977	.1592	66.4	44.1	3
—	—	—	—	—	6.276	.2004	52.6	35.0	4
—	—	—	—	—	7.914	.2536	41.7	27.7	5
—	—	—	—	—	9.980	.3192	33.1	22.0	7
—	—	—	—	—	12.58	.4028	26.3	17.5	8
—	—	—	—	—	15.87	.5080	20.8	13.8	9
—	58	—	19.6	19.9	20.01	.6045	16.5	11.0	10
—	74	—	24.6	25.1	25.23	.8077	13.1	8.7	11
87.5	92	80.0	30.9	31.6	31.82	1.018	10.4	6.9	12
110	114	95.5	38.8	39.8	40.12	1.284	8.2	5.5	13
136	144	121	48.9	50.2	50.59	1.619	6.5	4.4	14
170	182	150	61.5	63.2	63.80	2.042	5.2	3.5	15
211	225	183	77.3	79.6	80.44	2.575	4.1	2.7	16
262	282	223	97.3	100	101.4	3.247	3.3	2.2	17
321	357	271	119	124	127.9	4.094	2.6	1.7	17-18
397	450	329	150	155	161.3	5.163	2.0	1.3	18
493	558	399	188	196	203.4	6.510	1.6	1.1	19
592	708	479	237	247	256.5	8.210	1.3	.86	20
775	868	625	298	311	323.4	10.35	1.0	.68	21
940	1090	754	370	389	407.8	13.05	.81	.54	22
1150	1368	910	461	491	514.8	16.46	.64	.43	23
1400	1640	1080	584	624	648.4	20.76	.51	.34	24
1700	2140	1260	745	778	817.7	26.17	.41	.27	25
2060	2530	1510	903	958	1031	33.00	.32	.21	26
2500	3340	1750	1118	1188	1300	41.62	.25	.17	27-28
3030	4145	2020	1422	1533	1639	52.48	.20	.13	29
3670	5250	2310	1759	1903	2067	66.17	.16	.11	30
4300	6510	2700	2207	2461	2607	83.44	.13	.084	31-32
5040	8175	3020	2534	2893	3287	105.20	.10	.067	33
5920	10220	—	2768	3483	4145	132.70	.079	.053	34-35
7060	12650	—	3137	4414	5227	167.30	.063	.042	36
8120	16200	—	4697	5688	6591	211.00	.050	.033	36-37
9600	19950	—	6168	6400	8310	266.00	.039	.026	37-38
10900	25000	—	6737	8393	10480	335.00	.032	.021	38-39
12200	31700	—	7877	9846	13210	423.00	.025	.017	39-40
14000	39600	6510	9309	11636	16660	533.40	.020	.013	41
16600	49100	6950	10666	13848	21010	672.60	.016	.010	42-43
18000	62600	7450	11907	18286	26500	848.10	.012	.008	43
—	77600	—	14222	24381	33410	1069.00	.009	.006	44
—	97500	—	17920	30610	42130	1323.00	.008	.005	45
—	122000	—	22600	38700	53100	1667.00	.006	.004	45-46
—	152000	—	28410	48600	66970	2105.00	.005	.003	46-47
—	190000	—	35950	61400	84460	2655.00	.004	.0025	47

**Approximate only—thickness of insulation varies.

1 and 44. (Courtesy, Radio Electronics)

ported in insulators on a telephone pole crossarm. The wires may be of hard-drawn copper, steel, copper-galvanized steel, or iron. Two wires constitute a transmission line. The spacing between conductors is generally 8 inches. When more than one pair of wires is strung on one pole, the spacing between wires is increased to 10 or 12 inches. Wire diameters most frequently used are

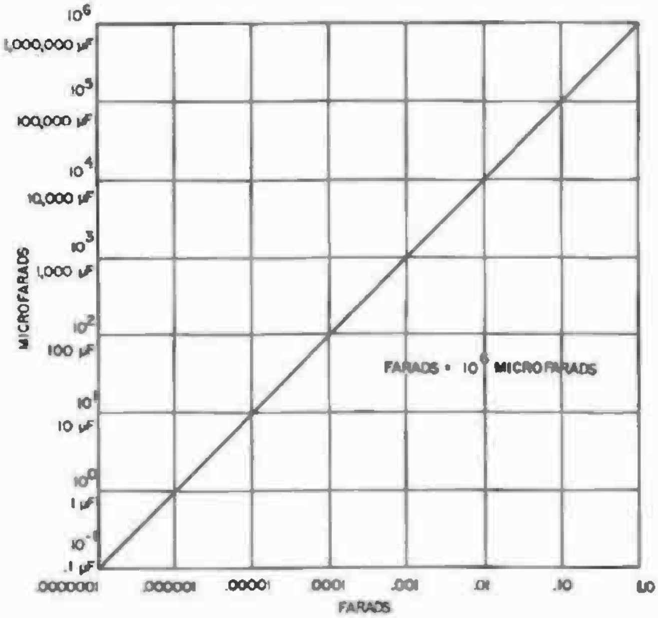


Fig. 25-164. Conversion chart—microfarads to farads.

from 80 to 165 mils. This is equivalent to a wire size of 12 to 6 gauge.

25.170 What is a cable pair?—A cable consisting of more than two pairs. Each wire is individually insulated. The wires of each pair are twisted together and the entire group is given a long spiral twist and then covered with a lead sheath or a plastic coating. The conductors are annealed copper. Standard gauges are 16 and 19.

25.171 What is a toll cable?—A cable made in the manner described in Question 25.170, except it is always covered with a lead sheath. It may be suspended from poles or buried in the ground. It is used for long-distance toll calls, hence its name.

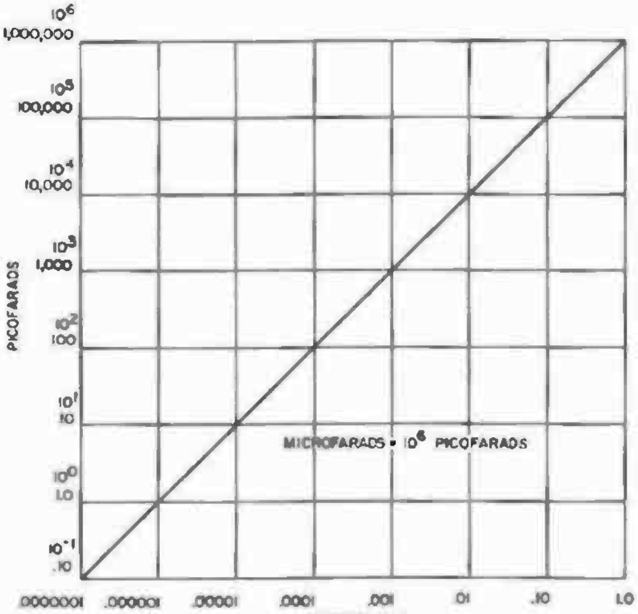


Fig. 25-165. Conversion chart—picofarads to microfarads.

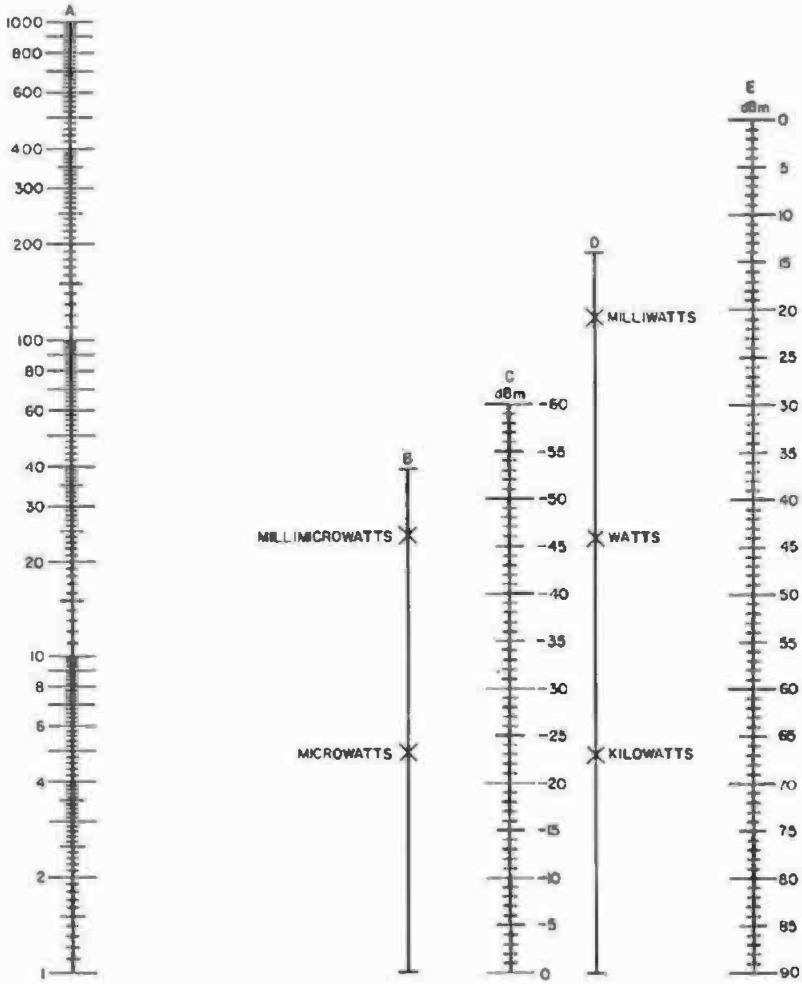


Fig. 25-174. Nomograph for converting absolute power units to dBm. (Courtesy, Electronics & Communications, Canada)

25.172 What are field wires?—Field wire, because of its high transmission loss, is used only for military and emergency purposes and for short temporary lines. It consists of two conductors made up of seven strands, of which four are copper and three are steel, to give strength. Each of the seven-strand conductors is covered with polyethylene insulation with an outer covering of nylon.

25.173 What is a loop mile?—It is the total wire in a 1-mile line consisting of two conductors, or 2 miles of wire.

25.174 Give a nomograph for converting absolute units of power to decibels.—The nomograph in Fig. 25-174 may be used to convert absolute units of power, such as watts, to decibels. To convert nanowatts or microwatts to dBm, position a straightedge between the appropriate number on scale A and the applicable cross-point on scale B. The resultant level in dBm is read from scale C.

To convert milliwatts watts, or kilowatts, scales A and D are employed, and the answer is read on scale E. For example, a power of 2 watts is equal to a level of plus 33 dBm as read on scale E. The level in dBm for a power of 10 microwatts (scales A and B) is minus 20 dBm as read on scale C.

25.175 What is considered to be a short line?—A short line is one in which the electrical length of the line is considerably shorter than the wavelength of the transmitted signal. An electrically short line represents a loop mile of line. If a signal of 1000 Hz is sent through the line, and the velocity of the signal in the line is assumed to be 180,000 miles per second, the wavelength of the signal is 180 miles, or the line is 1/180 of a wavelength. A second example: If the propagation velocity is 60,000 miles per second, the line is 1/60 of a wavelength. Although both lines are the same length physically, the slower circuit is electrically three times as long as the faster circuit.

25.176 What is considered to be a long line?—A long line is one in which the length is approximately equal to, or greater than, the wavelength of the transmitted signal. A line that would be considered electrically long might be 360 miles. If a 1000-Hz signal is applied to this line, the wavelength will he 180 miles. Under different circumstances, the same line may behave as either an electrically short or long line.

If the short line discussed in Question 25.175 were to be energized by a signal of 200,000 Hz, corresponding to a wavelength of 0.9 mile, this would be considered a long line. On the other hand, if it were energized by a signal of 60 Hz, corresponding to a wavelength of 3000 miles, it would he considered a short line.

25.177 What is an equivalent circuit of a long line?—A long line may be considered to be a series of low-pass filters connected in series with a series dc resistance. An equivalent circuit is shown in Fig. 25-177.

25.178 What is the equivalent circuit for a short line?—It is one section between the dotted lines of Fig. 25-177.

25.179 What is the electrical length of a telephone transmission line?—It is the relationship of the length of the line to the wavelength of the transmitted signal. The wavelength may be calculated:

$$\text{Wavelength} = \frac{v}{f}$$

where,

v is the velocity of propagation of the line at 1000 Hz,
f is the frequency in hertz.

For open wire, this figure is 176,000 to 180,000 miles per second. For nonloaded cable, it varies from 47,000 to 66,000 miles per second.

25.180 What are line parameters?—The constants of the line comprising the series resistance, series inductance, shunt capacitance, and the shunt leakage inductance. These elements are designated R, L, C, and G, respectively, and are based on 1 mile of loop line.

The numerical value of the constants depends on the size of the conductors, their spacing, insulation, the frequency of the transmitted signal, and weather conditions. The previously mentioned parameters are distributed along the entire length of the line. (See Question 25.177.)

25.181 How is the attenuation for a loop mile of line calculated?

dB attenuation =

$$\frac{R}{2}\sqrt{\frac{C}{L}} + \frac{G}{2}\sqrt{\frac{L}{C}} \times 8.686$$

where,

$\frac{R}{2}\sqrt{\frac{C}{L}}$ is the series loss,

$\frac{G}{2}\sqrt{\frac{L}{C}}$ is the shunt loss.

Generally, the series loss exceeds the shunt loss. Increasing the series loading inductance decreases the series loss but increases the shunt loss.

25.182 What is an artificial line?—A unit containing the equivalent constants of 1 mile of standard telephone line or cable. The parameters may be represented as shown in Fig. 25-177. A long line may be considered to be made up of a series of unit sections. Thus a 360-mile line would consist of five 72-mile units. Such setups are used in the laboratory to study the behavior of transmission lines.

25.183 What is the characteristic impedance of a transmission line?—The characteristic impedance of a transmission line is equal to the impedance that must be used to terminate the line in order to make the input impedance

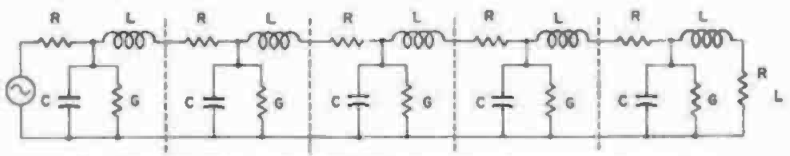


Fig. 25-177. The equivalent circuit for a long transmission line.

equal to the terminating impedance. For a long line (see Question 25.176) the input impedance will equal the characteristic impedance of the line, irrespective of the terminating impedance.

The characteristic impedance will also depend on the parameters of the pair and the applied frequency. The resistive component of the characteristic impedance is generally high at the low frequencies, falling off with an increase of frequency. The reactive component is high at the low frequencies and falls off as the frequency is increased.

The impedance of a uniform line is the impedance obtained for a long line (infinite length). It is apparent, for a long line, the current in the line is little affected by the value of the terminating impedance at the far end of the line. If the line has an attenuation of 20 dB and the far end is short circuited, the characteristic impedance as measured at the sending end will not be affected by more than 2 percent.

In practice, a transmission line is seldom uniform. The terminating equipment impedance will vary from that of the line. This may be corrected by the use of a repeat coil which may be matched to the line on one side and to the terminating equipment on the other.

25.184 ***What causes delay in a long transmission line?***—The parameters of the line, causing it to induce a time constant in the transmission.

25.185 ***What causes the velocity of propagation to vary with different transmission lines?***—A transmission line possesses three inherent properties which are the characteristics of all electrical circuits: resistance, inductance, and capacitance.

All three of these properties will exist regardless of how the line is constructed. Lines cannot be constructed to eliminate any of these characteristics.

Under the foregoing conditions, the velocity of the electrical pulses applied to the line are slowed down in their transmission. The elements of the line are distributed evenly and are not localized or present in a lumped quantity.

The velocity of propagation will vary from 10,000 to about 90,000 miles per second, depending on the construction of the line.

25.186 ***What does the phrase "loading a transmission line" mean?***—It is a method of loading a transmission line at stated intervals by connecting an inductance in series with the line. Two types of loading are in general usage, lumped and continuous. Loading a line increases the impedance of the line, thereby decreasing the series loss because of the conductor resistance.

Although loading decreases the attenuation and distortion and permits a more uniform frequency characteristic, it also increases the shunt losses caused by leakage. Loading also causes the line to have a cutoff frequency above which the loss becomes excessive.

25.187 ***What is the difference between a continuously loaded line and one which is lump loaded?***—In the continuously loaded line, loading is obtained by wrapping the complete cable with a high-permeability magnetic tape or wire. In the lumped loading method, toroidally wound coils are placed at equally spaced intervals along the line.

25.188 ***What is the effect of continuously loading a line?***—The inductance is distributed evenly along the line, causing it to behave as a line with distributed constants.

25.189 ***How are loading coils connected in a telephone transmission line?***—The coils are of toroidal construction and are connected in the lines as shown in Fig. 25-189. Each coil has an inductance on the order of 88 millihenries. The insulation between the line conductors and ground must be extremely good if the coils are to function properly. Loading coils will increase the talking distance by 35 to 90 miles for the average line.

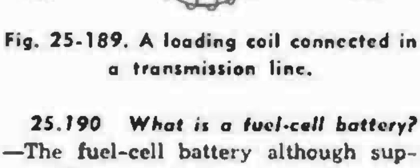


Fig. 25-189. A loading coil connected in a transmission line.

25.190 ***What is a fuel-cell battery?***—The fuel-cell battery although supposed by many to be a new discovery, was discovered by Sir William Grove in 1839. Later developments on the cell were made by Mond and Langer, who coined the name, "fuel cell." Work on this cell was resumed by Bacon in the

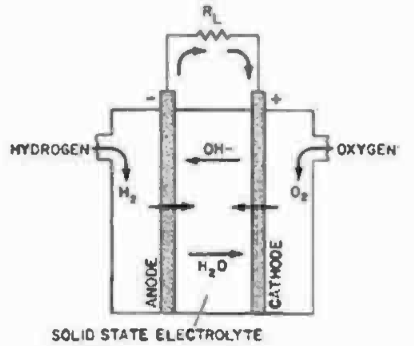


Fig. 25-190. Basic principle of operation of a fuel cell battery.

early 1930's. At present, such cells are used in spacecraft.

A fuel cell consists of an anode and cathode which are placed in contact with a solid-state electrolyte to permit the exchange of hydrogen ions between the electrodes. Fuel cells produce their power from a continuous supply of fuel (hydrogen and oxygen), hence their name. With the aid of a catalyst, the hydrogen atoms give out one electron each, forming ions which migrate through the solid electrolyte to the cathode. There they combine to create electricity and as a by-product produce drinkable water which is carried off by capillary action to a collection point.

Such cells have high efficiency, converting 50 to 70 percent of their energy to power. They require no charging and give off no fumes. The basic construction for such a cell is shown in Fig. 25-190.

25.191 Describe the piezoelectric effect. — Piezoelectricity is "pressure electricity" and is a property of certain crystals such as Rochelle salt, tourmaline, barium titanate, and quartz. When pressure is applied to any of these crystals, electricity is generated. If an electrical charge is applied to the crystal, it changes shape and can be used to impart motion. The piezoelectric effect was discovered in 1880 by Pierre and Jacques Curie. Present-day commercial materials have been especially developed for their piezoelectric qualities. Among them are ammonium dihydrogen phosphate (ADP), lithium sulphate (LN), dipotassium tartrate (DKT), po-

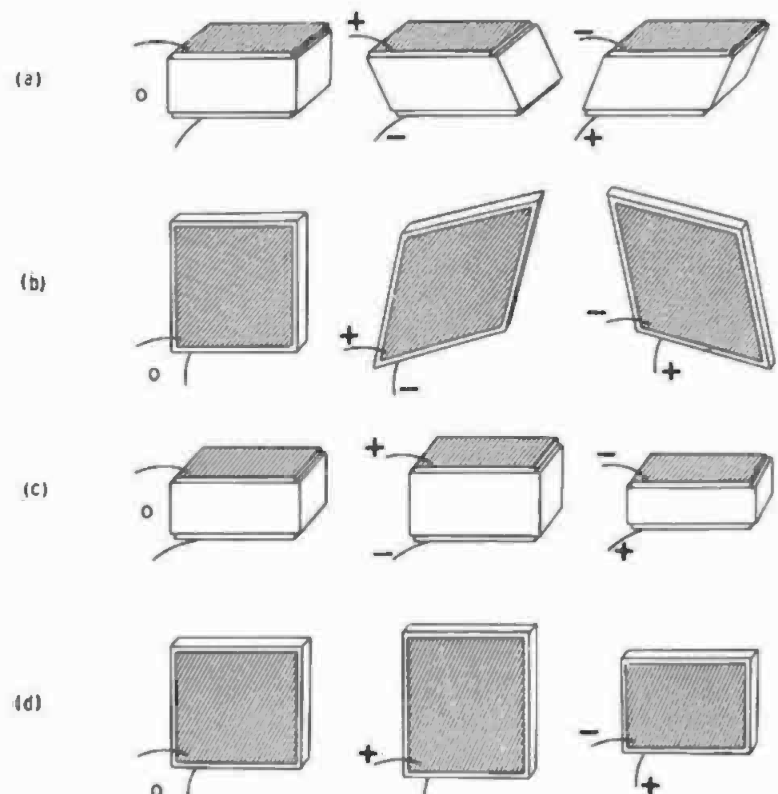


Fig. 25-191A. Basic deformations of piezoelectric plates.

tassium dihydrogen phosphate (KDP), lead zirconate, and lead titanate (PZT). Ceramics do not have piezoelectric characteristics in their original state, but the characteristics are introduced in the materials by a polarizing process. In piezoelectric ceramic materials the direction of the electrical and mechanical axes depend on the direction of the original dc polarizing potential. During polarization a ceramic element experiences a permanent increase in dimensions between the poling electrodes, and a permanent decrease in dimension parallel to the electrodes.

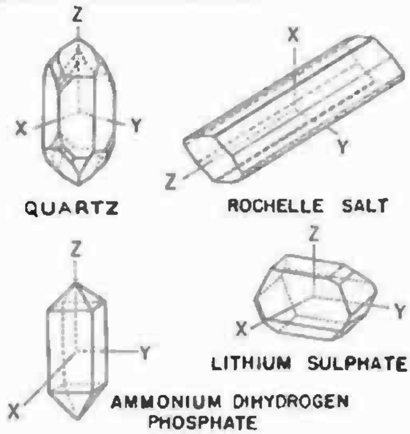


Fig. 25-191B. Different crystals and their axes. (Courtesy, Clevite Corp. Piezoelectric Div.)

When a dc voltage of the same polarity as the polarizing voltage, but smaller in magnitude, is applied between the polarizing terminals, the element experiences a temporary expansion in the polarizing direction and a contraction parallel to the electrodes. Conversely, when a dc voltage of the opposite polarity is applied, the element contracts in the polarizing direction and expands parallel to the electrodes. In either case, the element returns to its original poled

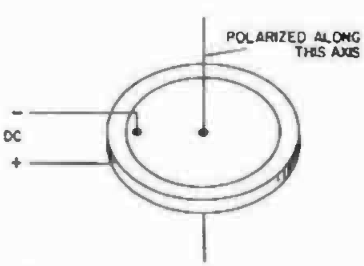


Fig. 25-191C. Expansion of ceramic disc during polarization.

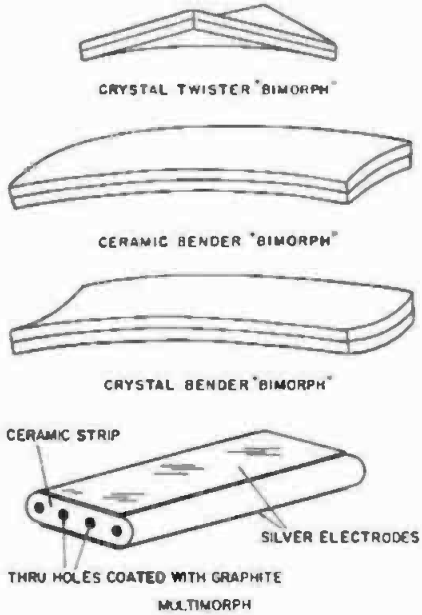


Fig. 25-191D. Curvatures of Bimorphs and Multimorph. (Courtesy, Clevite Corp. Piezoelectric Div.)

dimensions when the voltage is removed from the electrodes. These effects are shown, greatly exaggerated, in Fig. 25-191A. The thickness tranverse effects are not of equal magnitude, so accordingly there is a small change in volume when a voltage is applied between the terminals.

Deformation of a crystal that results from the application of an electric current and the nature of the deforming force depends on the type of piezoelectric material. Crystals and their axes are shown in Fig. 25-191B. Piezoelectric crystal elements are made from slabs cut from the whole crystal rather than by using the crystal itself. In the instance of piezoelectric ceramics, the axes are established with reference to

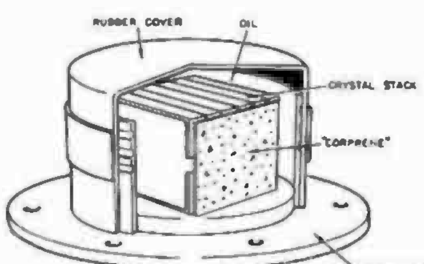


Fig. 25-191E. Underwater sound transducer. (Courtesy, Clevite Corp., Piezoelectric Div.)

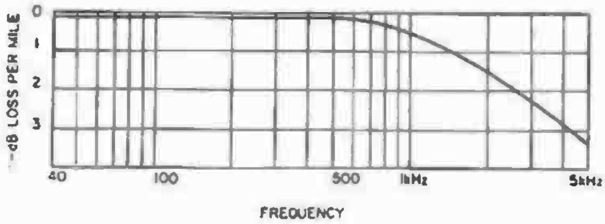


Fig. 25-193. The frequency characteristics of a nonloaded No. 9 cable pair, loss per loop mile in decibels.

the direction of the electric field used originally during the polarizing process. Basic deformation and the polarizing polarities for ceramic piezoelectric plates are shown in Fig. 25-191C. Curvatures for ceramic and crystal slabs (also greatly exaggerated) are shown in Fig. 25-191D. Combinations of these slabs are known as Bimorph and Multimorph, a tradename of Clevite Corp., Piezoelectric Division. Crystals of this type are used in the manufacture of microphones, pickups, speakers, disc recording heads, underwater sound communication devices, and many others. A typical ceramic underwater sound transducer is shown in Fig. 25-191E.

Rochelle salt crystals are the most commonly used types for microphones, pickups, and headphones, because of their greater sensitivity over ceramic crystals. However, in speakers the ceramic is preferred because of its greater stability and ability to handle greater power put out by modern amplifiers. (See Question 4.13.)

25.192 What is ferroelectricity?—In the search for control devices which might be substituted for a vacuum tube or transistor, considerable research has been conducted in the field of ferroelectricity, one of whose properties is nonlinear dielectric action. Ferroelectricity is an unusual characteristic exhibited by certain dielectric materials which alter their dielectric constant with the applied voltage. Linear dielectrics such as insulation commonly used in electronic devices exhibit no such properties.

Voltage - sensitive dielectrics that exhibit ferroelectric properties are barium titanate, titaninm dioxide, quanadine aluminum sulphate hexahydrate (GASH), and triglycene sulphate. Certain of these materials are used for phonograph pickups, microphones, earphones, and similar devices.

25.193 Describe an electrolytic switch.—It is a device consisting of two load-connected electrodes and a grid-element immersed in an electrolytic bath which is housed in a sealed container. A signal voltage applied to the acid-resistant grid-element makes the two load-electrodes permeable to ions and instantaneously changes the state of the electrodes from nonconducting to conducting. Large quantities of electrons flow from one electrode to the other through the electrolyte of acid and free metal ions. The actuating signal may be an on-off or modulating signal. When the switch is properly constructed, several amperes of current can be handled.

25.194 What is a simplex circuit?—A circuit in which a ground-return telephone or telegraph circuit is superimposed on a full-metallic circuit to obtain an extra channel. Provision is made to prevent interaction or interference between the two systems. An elementary diagram showing how a

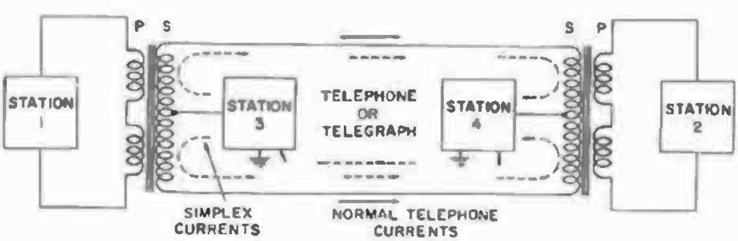


Fig. 25-194. Elementary diagram, showing the current paths of a simplex circuit.

simplex circuit functions is given in Fig. 25-194.

25.195 ***What is a duplex circuit?***—A circuit which will permit electrical communication between two points in both directions simultaneously. (The term full-duplex is synonymous with duplex.)

25.196 ***What is a half-duplex circuit?***—A circuit which will permit unidirectional electrical communication between two points. (Technical arrangements may permit communication in either direction, but not simultaneously.)

25.197 ***Give the names and abbreviations of technical societies and international commissions of interest to the audio engineer.***—Numerous technical societies have been formed for the dissemination of knowledge, and international commissions for the creation of world-wide standards. Among these organizations of interest to the Audio Engineer are:

ASA Acoustical Society of America
ASA American Standards Association (Now USASI)
ASI American Standards Institute (See USASI. Formerly ASA)
AES Audio Engineering Society
AIEE American Institute of Electrical Engineers (Now combined with IEEE)
CCIF International Telephonic Consultive Committee
CCIR International Radio Consultive Committee
CEE International Commission on Rules for Approval of Electrical Equipment
CSA Canadian Standards Association
DIN Deutsche Industrie Normen (Standards; can be obtained from USASI)
EIA Electronic Industries Association (Formerly RETMA)
EIAC Electronic Industries Association of Canada
IEEE Institute of Electrical and Electronic Engineers (Formerly AIEE)
IEC International Electrotechnical Commission
IHF Institute of High Fidelity
IRIG Inter Range Instrument Group
IRE Institute of Radio Engineers (now combined with IEEE)
ISO International Organization for Standardization
JAN Joint Army-Navy Specifications
MIRA Magnetic Recording Industries Association (Now EIA)
NAB National Association of Broadcasters (Formerly NARTB)
NARTB National Association of Radio and Television Broadcasters (Now NAB)
NEMA National Electrical Manufacturers Association
RETMA Radio Electronics Television Manufacturers Association (Now EIA)
RIAA Record Industries Association of America
SMPE Society of Motion Picture Engineers (Now SMPTE)
SMPTE Society of Motion Picture and Television Engineers (Formerly SMPE)
UL Underwriters' Laboratories (USA)
USASI United States of America Standards Institute (Formerly ASA)
USNBS United States National Bureau of Standards

The CCIF and CCIR international commissions have set international standards for the recording and reproduction of magnetic tape and records. Certain manufacturers of recording equipment provide plug-in equalization in tape recorders to meet these standards, as is discussed in Question 17.172.

25.198 ***Describe the Hall effect.***—In 1897, Edwin Hall, a physicist at Johns Hopkins University, while conducting basic research on the nature of electrical conduction discovered that if a very thin strip of gold foil was passed through the field of a powerful electromagnet carrying longitudinal currents, and a sensitive galvanometer was connected across the strip as shown in Fig. 25-198A, a difference of potential could be measured between the two edges. The equipotential lines running at right angles to the edges are skewed into an oblique position and the lines of current are deflected to one side. The Hall effect, when discovered, was of little interest except it added to the store of knowledge on the conduction of electric currents. However, since the extensive use of solid-state devices, the Hall effect has become of great interest, since semiconductor devices can develop as much as 1 volt of Hall voltage and can be used to operate a relay without further amplification.

Fig. 25-198B shows a modern version of the Hall effect, which produces an output voltage proportional to the

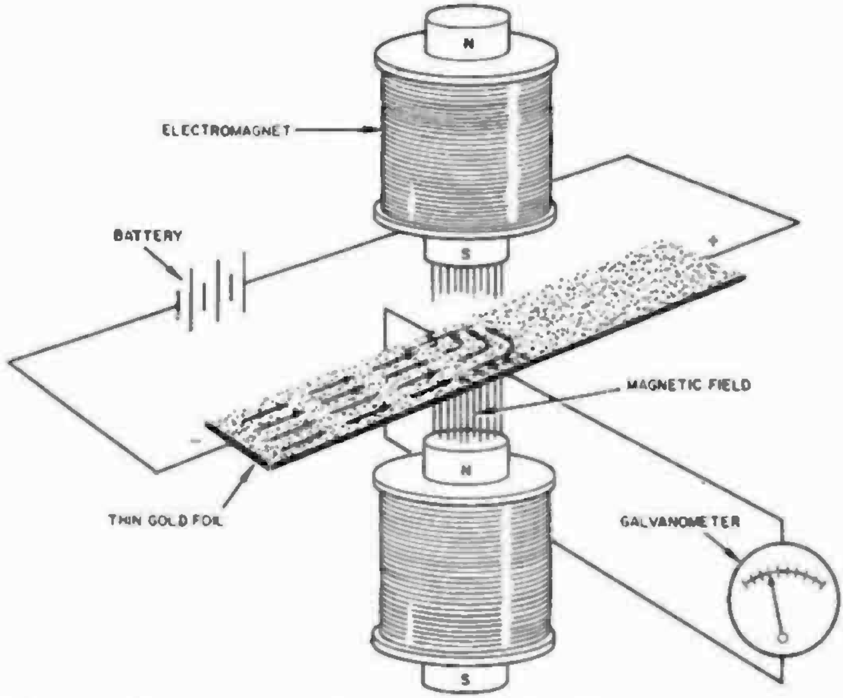


Fig. 25-198A. Basic principles of the generator used by Hall in his experiments which established the Hall effect. (Courtesy, Lenkurt Electric Co.)

product of the magnetic field strength and input current. The inherent Hall effect in semiconductors has a frequency response from dc to about 10^{12} Hz, or a 1000-GHz bandwidth. Since the advent of semiconductors, many uses for this effect have been found, such as in ammeters, voltmeters, wattmeters, multipliers, dividers, modulators, memory devices, and in a host of others.

25.199 What is a servomechanism? —An electromechanical system used for the control of a device at a remote point. H. L. Hazen, in 1934, proposed the term "servomechanism," which may be defined as "a power amplifying device in which the amplifying element driving the output is actuated by the difference between the input and the output." A typical example is a synchro generator at a given location which, when actuated, causes the shaft of a remote unit to turn to a given angle with power amplification. If the rotation of the shaft in the remote unit differs from that of the control unit, an error voltage is generated. This voltage is used to correct the difference in the angles of the two shafts and bring the remote unit to the exact angle of the control unit.

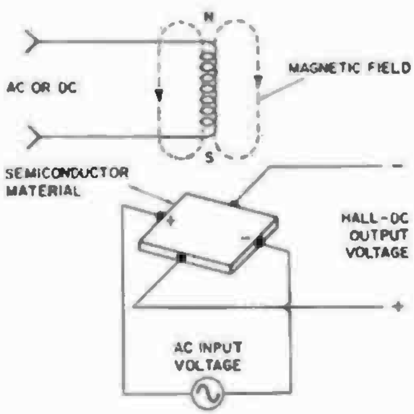


Fig. 25-198B. Modern version of a Hall-effect generator.

Selsyn interlock systems such as are described in Question 3.48 are often referred to as servo systems, although they do not meet the exact definition of such a system.

25.200 What is a skewed winding? —A special winding developed by the General Radio Co. for use with their variable delay lines. The wire is wound on a flat card in the shape of a letter D as shown in Fig. 25-200. After winding, the card is bent into a circle and placed

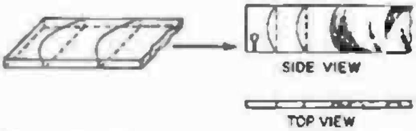


Fig. 25-200. The General Radio Co. skewed winding, for variable delay line use.

on a standard circular mounting used for a potentiometer or rheostat.

The use of skewed winding provides a more nearly constant effective inductance of the distributed winding. Skewing the inductance produces an effective inductance which remains nearly constant up to a critical value of phase change per turn. In effect, skewing offers a means of control over the mutual inductance between turns of a distributed-winding delay line.

Several different forms of skewed winding have been developed; however, the D-shaped winding on a flat mandrel card appears to be the most satisfactory one, since it results in a smooth winding of constant characteristic impedance which can be curved to fit a standard potentiometer mounting.

25.201 What is a reactance chart and how is it used?—A reactance chart developed by the Bell Telephone Lab-

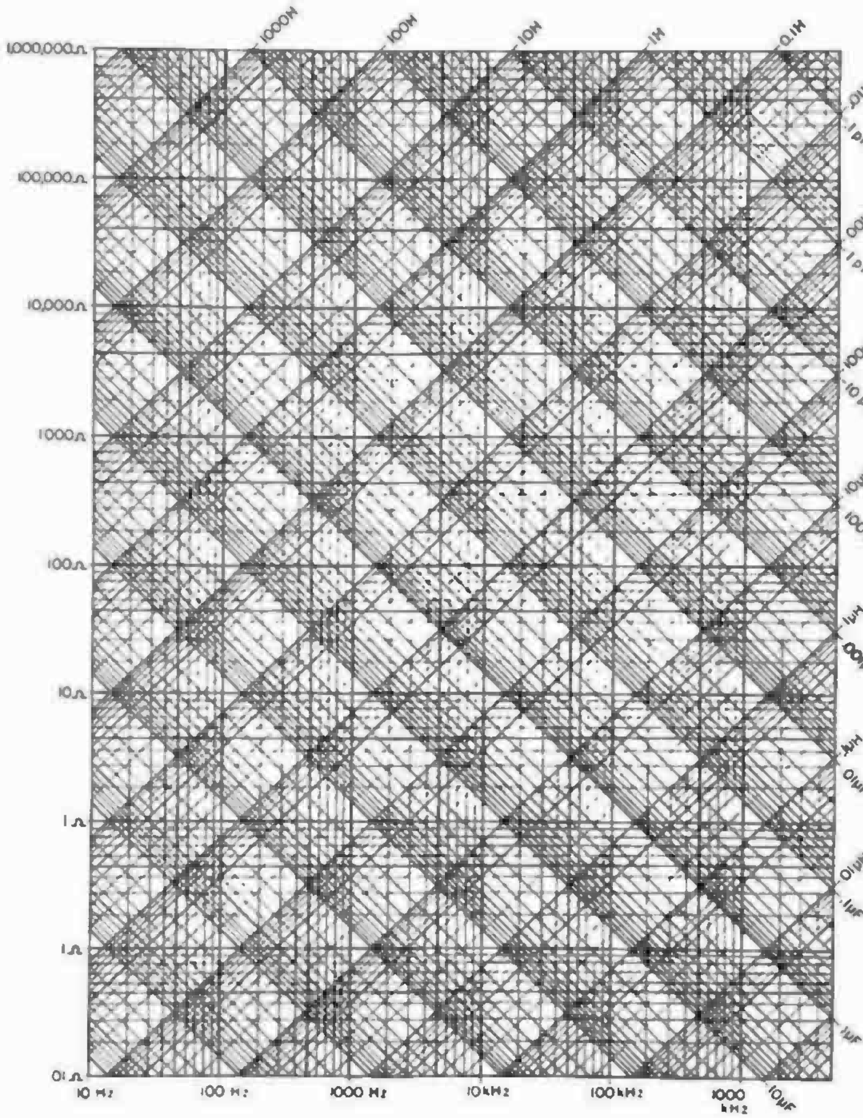


Fig. 25-201. A reactance-frequency chart. (Courtesy, Bell Telephone Laboratories)

MOLDED PAPER CAPACITOR CODES (CAPACITANCE GIVEN IN PF)			
COLOR	DIGIT	MULTI-PLIER	TOLER-ANCE
BLACK	0	1	20%
BROWN	1	10	
RED	2	100	
ORANGE	3	1000	
YELLOW	4	10000	
GREEN	5	100000	5%
BLUE	6	1000000	
VIOLET	7		
GRAY	8		
WHITE	9		10%
GOLD			5%
SILVER			10%
NO COLOR			20%

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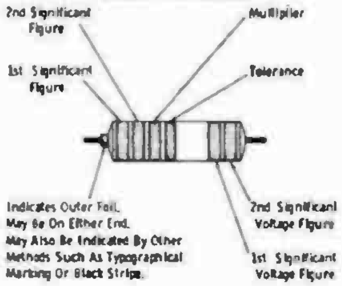
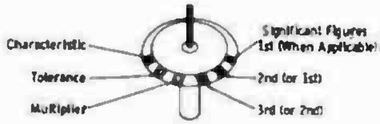
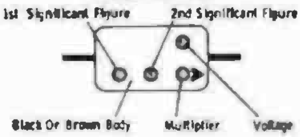


Add Two Zeros To Significant Voltage Figures.
One Band Indicates Voltage Ratings Under 1000 Volts.

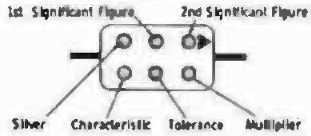
SILVERED MICA BUTTON CAPACITORS



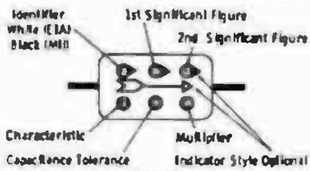
MOLDED FLAT PAPER CAPACITORS (COMMERCIAL CODE)



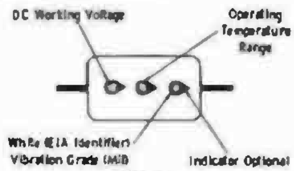
MOLDED FLAT PAPER CAPACITORS (MILITARY CODE)



CURRENT EIA AND MILITARY COLOR CODE FOR MOLDED MICA CAPACITORS



A (FRONT)



B (REAR)

NOTES:

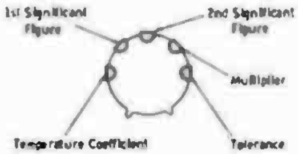
1. The multiplier is the factor by which the two significant figures are multiplied to yield the nominal capacitance.
2. "A" illustrates standard six-dot system used for "N" temperature range capacitors manufactured according to EIA Standard RS-153-A.
3. Drawings "A" and "B" combined illustrate standard nine-dot system used for "O" temperature range capacitors manufactured according to EIA Standard RS-153-A, and for all units manufactured according to Military Specification MIL-C-5C.

MICA CAPACITOR COLOR CODE							
COLOR	CHARAC-TERISTIC*	CAPACITANCE: 1ST AND 2ND SIGNIFICANT FIGURES	CAPACITANCE: MULTIPLIER	CAPACITANCE TOLERANCE	DC WORKING VOLTAGE	OPERATING TEMPERATURE RANGE	VIBRATION GRADE (MIL)
Black	A (EIA)	0	1	±20% (EIA)		- 55° to + 70°C (MIL)	10-55 Hz
Brown	B	1	10	±1%	100 (EIA)		
Red	C	2	100	±2%		- 55° to + 85°C	
Orange	D	3	1000		300		
Yellow	E	4	10000 (EIA)			- 55° to + 125°C	10-2000 Hz
Green	F	5		±5%	500		
Blue		6				- 55° to + 150°C (MIL)	
Purple (violet)		7					
Gray		8					
White		9					
Gold			0.1	±1/2% (EIA)	1000 (EIA)		
Silver			0.01 (EIA)	±10%			

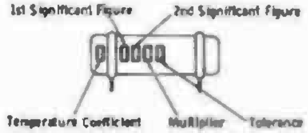
* Denotes specifications of design involving Q factors, temperature coefficients, and production test requirements.
† Or ±0.5 pf, whichever is greater. All others are specified tolerance or ±1.0 pf, whichever is greater.

Fig. 25-202. Color codes for resistors

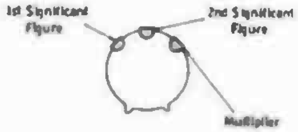
DISC CERAMICS (5-DOT SYSTEM)



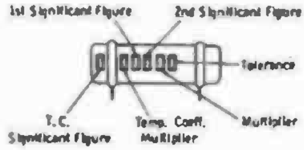
TEMPERATURE COMPENSATING TUBULAR CERAMICS



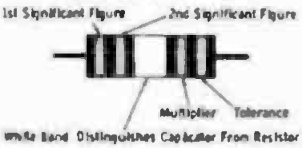
DISC CERAMICS (3-DOT SYSTEM)



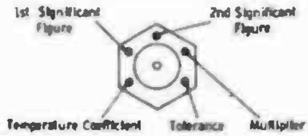
EXTENDED RANGE T.C. TUBULAR CERAMICS



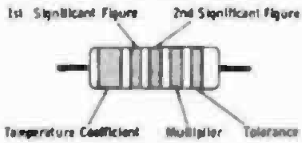
MOLDED CERAMICS
Using Standard Resistor Color Code



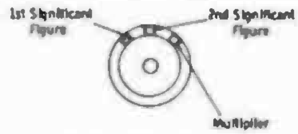
FEED-THRU CERAMICS



MOLDED-INSULATED AXIAL LEAD CERAMICS

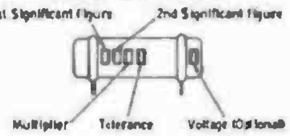


BUTTON CERAMICS

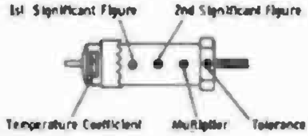


Viewed From Soldered Surface

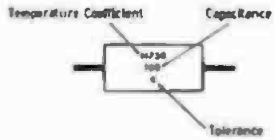
HIGH-CAPACITANCE TUBULAR CERAMICS
INSULATED OR NON-INSULATED



STAND-OFF CERAMICS



TYPOGRAPHICALLY MARKED CERAMICS



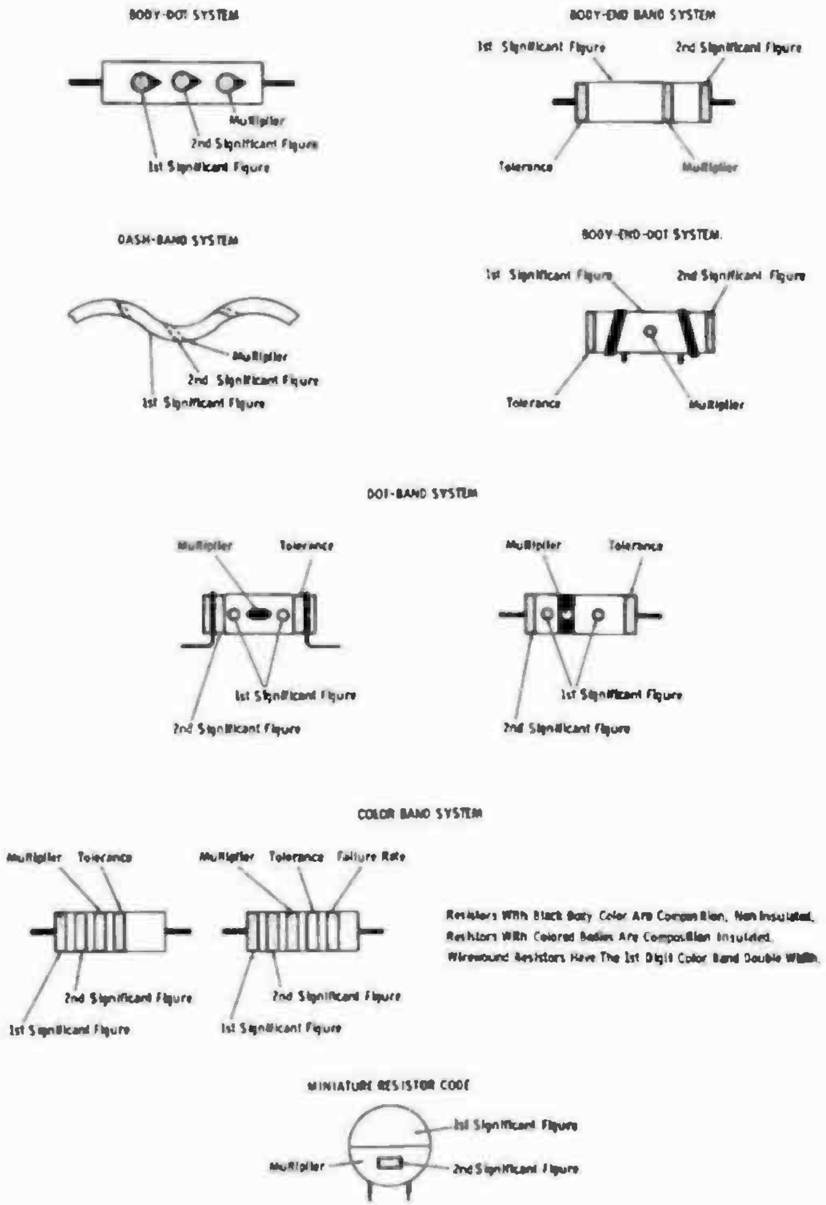
MIL LETTER	TOLERANCE	
	10PF OR LESS	OVER 10PF
C	± 0.25pf	
D	± 0.5pf	
F	± 1.0pf	± 1%
G	± 2.0pf	± 2%
J		± 5%
K		±10%
M		±20%

CERAMIC CAPACITOR CODES (CAPACITANCE GIVEN IN PF)

COLOR	DIGIT	MULTIPLIER	TOLERANCE		TEMPERATURE COEFFICIENT PPM °C	EXTENDED RANGE TEMP. COEFF.	
			10 PF OR LESS	OVER 10 PF		SIGNIFICANT FIGURE	MULTIPLIER
BLACK	0	1	± 2.0pf	± 20%*	0	0.0	-1
BROWN	1	10	± 0.1pf	± 1%	-33		-10
RED	2	100		± 2%	-75	1.0	-100
ORANGE	3	1000	± 0.25pf	± 2.5%*	-150	1.5	-1000
YELLOW	4	10000			-220	2.2	-10000
GREEN	5		± 0.5pf	± 5%	-330	3.3	+1
BLUE	6				-470	4.7	+10
VIOLET	7				-750	7.5	+100
GRAY	8	.01	± 0.25pf*		+30		+1000
WHITE	9	.1	± 1.0pf	± 10%	General Purpose		+10000
SILVER					Bypass & Coupling		
GOLD					+100 (MFG)		

*EIA only. Ceramic Capacitor capacitor voltage ratings are standard 500 volts, for some manufacturers, 1000 volts for other manufacturers, unless otherwise specified.

and capacitors (Continued next page).



RESISTOR COLOR CODE				
COLOR	1ST - 2ND SIGNIFICANT FIGURES	MULTIPLIER	TOLERANCE	FAILURE RATE*
Black	0	1	-	-
Brown	1	10	±1%	1.0
Red	2	100	±2%	0.1
Orange	3	1000	±3%	0.01
Yellow	4	10000	±4%	0.001
Green	5	100000	-	-
Blue	6	1000000	-	-
Violet	7	10000000	-	-
Gray	8	100000000	-	-
White	9	-	-	Solderable*
Gold	-	0.1	±5%	-
Silver	-	0.01	±10%	-
No Color	-	-	±20%	-

*When used on composition resistors, indicates percent failure per 1,000 hours. On film resistors, a white fifth band indicates solderable terminal.

Fig. 25-202. Color codes for resistors and capacitors (Cont'd).

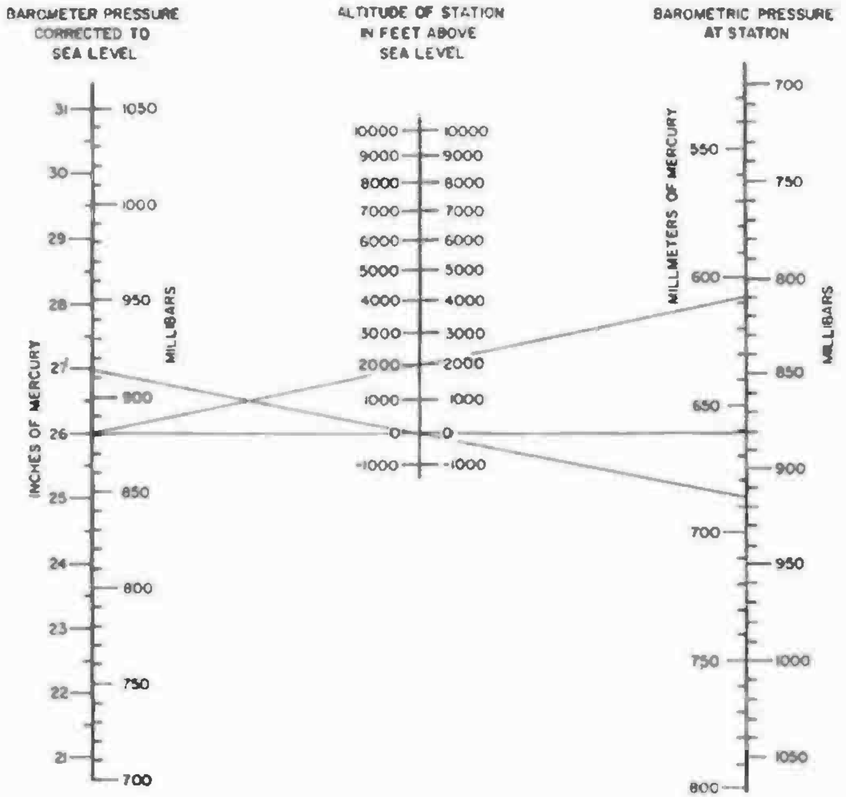


Fig. 25-203A. Nomograph for supplying altitude correction to barometric pressure.

oratories is shown in Fig. 25-201. This chart can be used for solving problems of inductance, capacitance, frequency, and impedance. If two of the values are known, the third and fourth may be found with its use. As an example, what is the value of capacitance and inductance required to resonate at a frequency of 1000 Hz in a circuit having an impedance of 500 ohms? Entering the chart on the 1000-Hz vertical line and following it to the 500-ohm line (impedance is shown along the left-hand margin) the value of inductance is indicated by the diagonal line running upward as 0.08 henries (80 millihenries), and the capacitance read at the right-hand margin is 0.3 μF.

A practical problem involving the use of this chart is given in Question 6.43. The chart may also be used for determining the reactance of an inductance or a capacitor and many other problems.

25.202 ***Give the color code used for capacitors and resistors.*** — Color codes for both resistors and various types of capacitors are given in the illustration of Fig. 25-202.

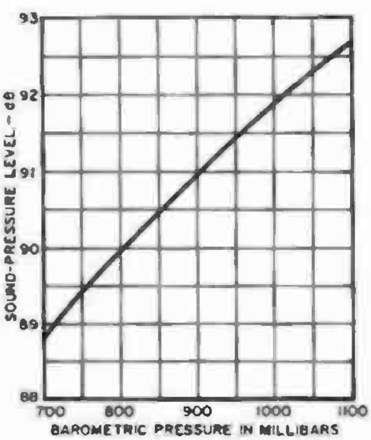


Fig. 25-203B. Relationship of decibels to barometric pressure.

25.203 ***What effect does barometric pressure have on sound levels?***—Altitude and barometric pressures are significant to the sound engineer particularly for the calibration of microphones. Most barometers are calibrated to read pressure in terms of sea level. If the altitude is not known, it may be obtained from the local weather bureau and correc-

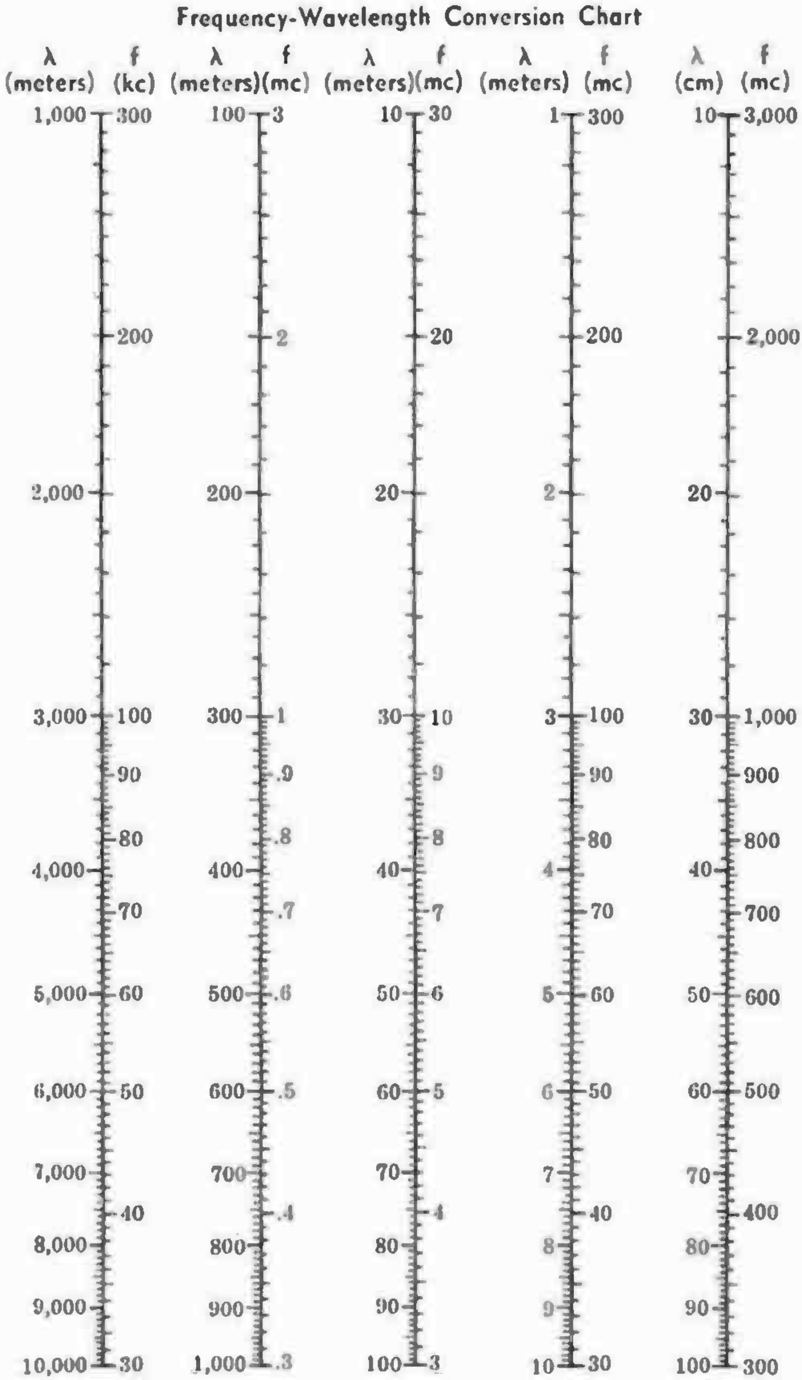


Fig. 25-204. Frequency to wavelength conversion nomograph.

Fraction	64ths	Decimal	Millimeters	Fraction	64ths	Decimal	Millimeters
	1	.015625	.0397		33	.515625	13.097
1/32	2	.03125	.794	17/32	34	.53125	13.494
	3	.046875	1.191		35	.546875	13.891
1/16	4	.0625	1.588	9/16	36	.5625	14.288
	5	.078125	1.984		37	.578125	14.684
3/32	6	.09375	2.381	19/32	38	.59375	15.081
	7	.109375	2.778		39	.609375	15.478
1/8	8	.125	3.175	5/8	40	.625	15.875
	9	.140625	3.572		41	.640625	16.272
5/32	10	.15625	3.969	21/32	42	.65625	16.669
	11	.171875	4.366		43	.671875	17.066
3/16	12	.1875	4.763	11/16	44	.6875	17.463
	13	.203125	5.159		45	.703125	17.859
7/32	14	.21875	5.556	23/32	46	.71875	18.256
	15	.234375	5.953		47	.734375	18.653
1/4	16	.250	6.350	3/4	48	.750	19.050
	17	.265625	6.747		49	.765625	19.447
9/32	18	.28125	7.144	25/32	50	.78125	19.844
	19	.296875	7.541		51	.796875	20.241
5/16	20	.3125	7.938	13/16	52	.8125	20.638
	21	.328125	8.334		53	.828125	21.034
11/32	22	.34375	8.731	27/32	54	.84375	21.431
	23	.359375	9.128		55	.859375	21.828
3/8	24	.375	9.525	7/8	56	.875	22.225
	25	.390625	9.922		57	.890625	22.622
13/32	26	.40625	10.319	29/32	58	.90625	23.019
	27	.421875	10.716		59	.921875	23.416
7/16	28	.4375	11.113	15/16	60	.9375	23.813
	29	.453125	11.509		61	.953125	24.209
15/32	30	.46875	11.906	31/32	62	.96875	24.606
	31	.484375	12.303		63	.984375	25.003
1/2	32	.500	12.700	1	64	1.000	25.400

Fig. 25-205. Decimal and millimeter equivalents, in 64th's of an inch.

tions made for altitudes other than sea level. These corrections are a function of altitude, temperature, and pressure, the principal factor being *altitude*, which amounts to 1 inch of mercury per 1000 feet above sea level.

Fig. 25-203A shows a conversion nomograph for inches of mercury to millibars. While the pressure should be reasonably accurate, an error of 34 millibars (1 inch of mercury) in barometric pressure will cause only an error of approximately 0.15 dB in microphone calibration, and 0.3 dB in sound-level calibrations.

Referring to the nomograph, barometric corrections are shown in millibars for altitudes from 1000 feet below to 10,000 feet above sea level. By selecting the altitude of the measuring location on the center scale, pressure may be read in either inches of mercury or millibars at the left. Connecting these points by a straightedge, changes are determined from the right-hand scale. For example, with a barometric pressure of 26 inches at sea level, the reading on the right-hand scale is 880 millibars. But, if the altitude is 2000 feet the barometric pressure is reduced to approximately 808 millibars. With a rise of 1 inch of pressure at sea level the pressure at the station rises to 914, as read on the right-hand scale.

The relationship of decibels to barometric pressure is shown in the graph in Fig. 25-203B.

25.204 ***Give a frequency-to-wavelength conversion nomograph.*** — The wavelength for any frequency between 30 kHz and 3000 MHz can be read directly by means of the nomograph given in Fig. 25-204. If only the frequency is known, it may be converted to wave-

MM	*Inches*	*MM*	*Inches*	*MM*	*Inches*
.01	.0004	.45	.0177	.89	.0350
.02	.0008	.46	.0181	.90	.0354
.03	.0012	.47	.0185	.91	.0358
.04	.0016	.48	.0189	.92	.0362
.05	.0020	.49	.0193	.93	.0366
.06	.0024	.50	.0197	.94	.0370
.07	.0028	.51	.0201	.95	.0374
.08	.0031	.52	.0205	.96	.0378
.09	.0035	.53	.0209	.97	.0382
.10	.0039	.54	.0213	.98	.0386
.11	.0043	.55	.0217	.99	.0390
.12	.0047	.56	.0221	1.00	.0394
.13	.0051	.57	.0224	2.00	.0787
.14	.0055	.58	.0228	3.00	.1181
.15	.0059	.59	.0232	4.00	.1575
.16	.0063	.60	.0236	5.00	.1969
.17	.0067	.61	.0240	6.00	.2362
.18	.0071	.62	.0244	7.00	.2756
.19	.0075	.62	.0248	8.00	.3150
.20	.0079	.64	.0252	9.00	.3543
.21	.0083	.65	.0256	10.00	.3937
.22	.0087	.66	.0260	11.00	.4331
.23	.0091	.67	.0264	12.00	.4724
.24	.0094	.68	.0268	13.00	.5118
.25	.0098	.69	.0272	14.00	.5512
.26	.0102	.70	.0276	15.00	.5906
.27	.0106	.71	.0280	16.00	.6299
.28	.0110	.72	.0284	17.00	.6693
.29	.0114	.73	.0287	18.00	.7087
.30	.0118	.74	.0291	19.00	.7480
.31	.0122	.75	.0295	20.00	.7847
.32	.0126	.76	.0299	21.00	.8268
.33	.0130	.77	.0303	22.00	.8661
.34	.0134	.78	.0307	23.00	.9055
.35	.0138	.79	.0311	24.00	.9449
.36	.0142	.80	.0315	25.00	.9843
.37	.0146	.81	.0319	26.00	1.0236
.38	.0150	.82	.0323	27.00	1.0630
.39	.0154	.83	.0327	28.00	1.1024
.40	.0158	.84	.0331	29.00	1.1417
.41	.0161	.85	.0335	30.00	1.1811
.42	.0165	.86	.0339	31.00	1.2205
.43	.0169	.87	.0343	32.00	1.2598
.44	.0173	.88	.0347	33.00	1.2992

Fig. 25-206. Millimeter equivalents in inches.

Unit	*Equivalents*	*Unit*	*Equivalents*
1 Hp.	746 watts	1 Hp.-hr.	0.746 kw. hours
1 Hp.	0.746 kw.	1 Hp.-hr.	1,980,000 ft-lbs.
1 Hp.	33,000 ft.-lbs. per minute	1 Hp.-hr.	2,545 Btu.
1 Hp.	550 ft.-lbs. per second	1 Hp.-hr.	273,740 kilogram meters
1 Hp.	2,544 Btu. per hour	1 Hp.-hr.	2.64 lbs. water evaporated from and at 212°F.
1 Hp.	42.4 Btu. per minute		
1 Hp.	0.707 Btu. per second	1 Hp.-hr.	17.0 lbs. water raised from 62°F. to 212°F.
1 Hp.	2.64 lbs. water evaporated per hour from and at 212°F.		
1 Kw.	1,000 watts	1 Kw.-hr.	1,000 watt hours
1 Kw.	1.34 horsepower	1 Kw.-hr.	1.34 horsepower hours
1 Kw.	2,654,200 ft.-lbs. per hour	1 Kw.-hr.	2,654,200 ft.-lbs.
1 Kw.	44,240 ft.-lbs. per minute	1 Kw.-hr.	3,600,000 joules
1 Kw.	737.3 ft.-lbs. per second	1 Kw.-hr.	3,413 Btu.
1 Kw.	3,413 Btu. per hour	1 Kw.-hr.	367,000 kilogram meters
1 Kw.	56.9 Btu. per minute	1 Kw.-hr.	3.53 lbs. water evaporated from and at 212°F.
1 Kw.	0.948 Btu. per second		
1 Kw.	3.53 lbs. water evaporated per hour from and at 212° F.	1 Kw.-hr.	22.75 lbs. water raised from 62°F. to 212°F.
1 Watt	1 joule per second		
1 Watt	0.00135 horsepower		
1 Watt	3.413 Btu. per hour	1 Ft.-lb.	1.356 joules
1 Watt	0.7373 ft. lb. per second	1 Ft.-lb.	0.1383 k.g.m.
1 Watt	0.0035 lb. water evaporated per hour	1 Ft.-lb.	0.000000377 kw. hours
		1 Ft.-lb.	0.001285 Btu.
1 Watt	44.24 ft.-lbs. per minute	1 Ft.-lb.	0.0000005 hp. hour

Fig. 25-207. Mechanical, electrical, and heat equivalents.

length for values of 10 centimeters to 1000 meters. The value of frequency or wavelength is found on one of the vertical lines and the desired converted value is read from the opposite side of the line.

25.205 ***Give a fractional, decimal, and millimeter equivalent table.***—Such a table is shown in Fig. 25-205.

25.206 ***Give a millimeter-inch equivalent table.***—Such a table appears in Fig. 25-206.

25.207 ***Give a table of mechanical, electrical, and heat equivalents.***—Such a table is given in Fig. 25-207.

25.208 ***Give a general conversion-factor table.***—Such a table is given in Fig. 25-208.

25.209 ***Give a table of common logarithms.***—Such a table is given in Fig. 25-209.

25.210 ***Give a table of natural trigonometric functions.***—Such a table appears in Fig. 25-210.

25.211 ***Give a table of dielectric constants.***—The dielectric constant of materials is generally affected by both temperature and frequency, except for quartz, Styrofoam, and Teflon, whose dielectric constants remain essentially constant. Small differences in the composition of a given material will also affect the dielectric constant. A table of dielectric constants is given in Fig. 25-211.

25.212 ***Give the capacitance per foot for most frequently used coaxial cable.***—Coaxial cable is used quite extensively with various types of test equipment. When such cable is replaced the capacitance per foot must be taken into consideration, particularly for oscilloscope probes. The capacitance value per foot is given in Fig. 25-212. The various types of cable may be identified from their coding:

R Radio frequeucy
G Made for US Government
— Number assigned by Government approval
U Universal specification.

Thus, a cable marked, RG-11a/U means radio frequency, Government, approval number, with a universal specification.

To Convert	*Multiply By*
Bars to dynes per square centimeter	1.00
British thermal units to foot-pounds	778.00
British thermal units to joules	1055.00
British thermal units to watt-hours	0.293
British thermal units per minute to horsepower	0.02356
Centimeters to feet	0.03281
Centimeters to inches	0.3937
Centimeters to meters	0.01
Centimeters to mils	393.70
Centimeters to millimeters	10.00
Circular mils to square centimeters	5.067×10^{-6}
Circular mils to square inches	7.854×10^{-7}
Circular mils to square millimeters	5.066×10^{-4}
Circular mils to square mils	0.7854
Degrees to minutes	60.00
Degrees to radians	0.01745
Dynes per square centimeter to bars	1.00
Ergs to joules	10^{-7}
Feet to centimeters	30.48
Feet to meters	0.3048
Foot-pounds to British thermal units	1.285×10^{-3}
Foot-pounds to joules	1.356
Foot-pounds to kilogram meters	0.1383
Foot-pounds per minute to horsepower	3.03×10^{-5}
Foot-pounds per minute to kilowatts	2.260×10^{-5}
Foot-pounds per minute to watts	0.0226
Foot-pounds per second to horsepower	1.818×10^{-3}
Foot-pounds per second to kilowatts	1.356×10^{-3}
Foot-pounds per second to watts	1.356
Gram calories to joules	4.186
Horsepower to foot-pounds per minute	33,000.00
Horsepower to foot-pounds per second	550.00
Horsepower to kilowatts	0.746
Horsepower to watts	746.00
Inches to centimeters	2.54
Inches to meters	0.0254
Inches to millimeters	25.40
Inches to mils	1,000.00
Joules to British thermal units	9.47×10^{-4}
Joules to ergs	10^{7}
Joules to foot-pounds	0.7375
Joules to gram-calories	0.2388
Joules to kilogram-meters	0.10198
Joules to watt-hours	2.778×10^{-4}
Kilograms to pounds	2.205
Kilogram-meters to foot-pounds	7.233
Kilogram-meters to joules	9.8117
Kilogram-meters per second to watts	9.807
Kilograms per kilometer to pounds per 1,000 feet	0.6719

Fig. 25-208. Conversion factors for area, length,

To Convert	*Multiply By*
Kilometers to feet	3,281.00
Kilometers to miles	0.6214
Kilometers to yards	1,093.60
Kilowatts to horsepower	1.341
Meters to feet	3.2808
Meters to inches	39.3701
Meters to yards	1.0936
Miles to kilometers	1.6093
Millimeters to inches	0.03937
Millimeters to mils	39.3701
Mils to centimeters	2.54×10^{-3}
Mils to inches	0.001
Mils to millimeters	0.0254
Ohms per kilometer to ohms per 1,000 feet	0.3048
Ohms per 1,000 feet to ohms per kilometer	3.2808
Ohms per 1,000 yards to ohms per kilometer	1.0936
Pounds to kilograms	0.4536
Pounds per 1,000 feet to kilograms per kilometer	1.488
Pounds per 1,000 yards to kilograms per kilometer	0.4960
Pounds per 1,000 yards to pounds per kilometer	1.0936
Radians to degrees	57.30
Radians to minutes	3438.00
Resistivity in microhm centimeters to ohms CMF	6.0153
Resistivity in ohms CMF to microhm centimeters	0.166
Specific gravity to pounds per cubic inch	0.0361
Square centimeters to circular mils	1.973×10^{5}
Square centimeters to square feet	1.076×10^{-3}
Square centimeters to square inches	0.155
Square feet to square centimeters	929.00
Square feet to square inches	144.00
Square feet to square meters	0.0929
Square inches to circular mils	1,273,240.00
Square inches to square centimeters	6.4516
Square inches to square feet	6.944×10^{-3}
Square inches to square mils	10^{6}
Square inches to square millimeters	645.16
Square meters to square feet	10.764
Square millimeters to circular mils	1,973.51
Square millimeters to square inches	1.55×10^{-3}
Square mils to circular mils	1.2732
Square mils to square centimeters	6.452×10^{-6}
Square mils to square inches	10^{-6}
Watts to foot-pounds per minute	44.25
Watts to foot-pounds per second	0.7375
Watts to horsepower	1.341×10^{-3}
Watts to kilogram-meters per second	0.1020
Watt-hours to British thermal units	3.4126
Yards to centimeters	91.44
Yards to meters	0.9144

COMMON LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9	N
10	0000	0043	0086	0128	0170	0212	0253	0294	0334	0374	10
11	0414	0453	0492	0531	0569	0607	0645	0682	0719	0755	11
12	0792	0828	0864	0899	0934	0969	1004	1038	1072	1106	12
13	1139	1173	1206	1239	1271	1303	1335	1367	1399	1430	13
14	1461	1492	1523	1553	1584	1614	1644	1673	1703	1732	14
15	1761	1790	1818	1847	1875	1903	1931	1959	1987	2014	15
16	2041	2068	2095	2122	2148	2175	2201	2227	2253	2279	16
17	2304	2330	2355	2380	2405	2430	2455	2480	2504	2529	17
18	2553	2577	2601	2625	2648	2672	2695	2718	2742	2765	18
19	2788	2810	2833	2856	2878	2900	2923	2945	2967	2989	19
20	3010	3032	3054	3075	3096	3118	3139	3160	3181	3201	20
21	3222	3243	3263	3284	3304	3324	3345	3365	3385	3404	21
22	3424	3444	3464	3483	3502	3522	3541	3560	3579	3598	22
23	3617	3636	3655	3674	3692	3711	3729	3747	3766	3784	23
24	3802	3820	3838	3856	3874	3892	3909	3927	3945	3962	24
25	3979	3997	4014	4031	4048	4065	4082	4099	4116	4133	25
26	4150	4166	4183	4200	4216	4232	4249	4265	4281	4298	26
27	4314	4330	4346	4362	4378	4393	4409	4425	4440	4456	27
28	4472	4487	4502	4518	4533	4548	4564	4579	4594	4609	28
29	4624	4639	4654	4669	4683	4698	4713	4728	4742	4757	29
30	4771	4786	4800	4814	4829	4843	4857	4871	4886	4900	30
31	4914	4928	4942	4955	4969	4983	4997	5011	5024	5038	31
32	5051	5065	5079	5092	5105	5119	5132	5145	5159	5172	32
33	5185	5198	5211	5224	5237	5250	5263	5276	5289	5302	33
34	5315	5328	5340	5353	5366	5378	5391	5403	5416	5428	34
35	5441	5453	5465	5478	5490	5502	5514	5527	5539	5551	35
36	5563	5575	5587	5599	5611	5623	5635	5647	5658	5670	36
37	5682	5694	5705	5717	5729	5740	5752	5763	5775	5786	37
38	5798	5809	5821	5832	5843	5855	5866	5877	5888	5899	38
39	5911	5922	5933	5944	5955	5966	5977	5988	5999	6010	39
40	6021	6031	6042	6053	6064	6075	6085	6096	6107	6117	40
41	6128	6138	6149	6160	6170	6180	6191	6201	6212	6222	41
42	6232	6243	6253	6263	6274	6284	6294	6304	6314	6325	42
43	6335	6345	6355	6365	6375	6385	6395	6405	6415	6425	43
44	6435	6444	6454	6464	6474	6484	6493	6503	6513	6522	44
45	6532	6542	6551	6561	6571	6580	6590	6599	6609	6618	45
46	6628	6637	6646	6656	6665	6675	6684	6693	6702	6712	46
47	6721	6730	6739	6749	6758	6767	6776	6785	6794	6803	47
48	6812	6821	6830	6839	6848	6857	6866	6875	6884	6893	48
49	6902	6911	6920	6928	6937	6946	6955	6964	6972	6981	49
50	6990	6998	7007	7016	7024	7033	7042	7050	7059	7067	50
51	7076	7084	7093	7101	7110	7118	7126	7135	7143	7152	51
52	7160	7168	7177	7185	7193	7202	7210	7218	7226	7235	52
53	7243	7251	7259	7267	7275	7284	7292	7300	7308	7316	53
54	7324	7332	7340	7348	7356	7364	7372	7380	7388	7396	54
N	0	1	2	3	4	5	6	7	8	9	N

Fig. 25-209. Table of

COMMON LOGARITHMS (Continued)

N	0	1	2	3	4	5	6	7	8	9	N
55	7404	7412	7419	7427	7435	7443	7451	7459	7466	7474	55
56	7482	7490	7497	7505	7513	7520	7528	7536	7543	7551	56
57	7559	7566	7574	7582	7589	7597	7604	7612	7619	7627	57
58	7634	7642	7649	7657	7664	7672	7679	7686	7694	7701	58
59	7709	7716	7723	7731	7738	7745	7752	7760	7767	7774	59
60	7782	7789	7796	7803	7810	7818	7825	7832	7839	7846	60
61	7853	7860	7868	7875	7882	7889	7896	7903	7910	7917	61
62	7924	7931	7938	7945	7952	7959	7966	7973	7980	7987	62
63	7993	8000	8007	8014	8021	8028	8035	8041	8048	8055	63
64	8062	8069	8075	8082	8089	8096	8102	8109	8116	8122	64
65	8129	8136	8142	8149	8156	8162	8169	8176	8182	8189	65
66	8195	8202	8209	8215	8222	8228	8235	8241	8248	8254	66
67	8261	8267	8274	8280	8287	8293	8299	8306	8312	8319	67
68	8325	8331	8338	8344	8351	8357	8363	8370	8376	8382	68
69	8388	8395	8401	8407	8414	8420	8426	8432	8439	8445	69
70	8451	8457	8463	8470	8476	8482	8488	8494	8500	8506	70
71	8513	8519	8525	8531	8537	8543	8549	8555	8561	8567	71
72	8573	8579	8585	8591	8597	8603	8609	8615	8621	8627	72
73	8633	8639	8645	8651	8657	8663	8669	8675	8681	8686	73
74	8692	8698	8704	8710	8716	8722	8727	8733	8739	8745	74
75	8751	8756	8762	8768	8774	8779	8785	8791	8797	8802	75
76	8808	8814	8820	8825	8831	8837	8842	8848	8854	8859	76
77	8865	8871	8876	8882	8887	8893	8899	8904	8910	8915	77
78	8921	8927	8932	8938	8943	8949	8954	8960	8965	8971	78
79	8976	8982	8987	8993	8998	9004	9009	9015	9020	9025	79
80	9031	9036	9042	9047	9053	9058	9063	9069	9074	9079	80
81	9085	9090	9096	9101	9106	9112	9117	9122	9128	9133	81
82	9138	9143	9149	9154	9159	9165	9170	9175	9180	9186	82
83	9191	9196	9201	9206	9212	9217	9222	9227	9232	9238	83
84	9243	9248	9253	9258	9263	9269	9274	9279	9284	9289	84
85	9294	9299	9304	9309	9315	9320	9325	9330	9335	9340	85
86	9345	9350	9355	9360	9365	9370	9375	9380	9385	9390	86
87	9395	9400	9405	9410	9415	9420	9425	9430	9435	9440	87
88	9445	9450	9455	9460	9465	9469	9474	9479	9484	9489	88
89	9494	9499	9504	9509	9513	9518	9523	9528	9533	9538	89
90	9542	9547	9552	9557	9562	9566	9571	9576	9581	9586	90
91	9590	9595	9600	9605	9609	9614	9619	9624	9628	9633	91
92	9638	9643	9647	9652	9657	9661	9666	9671	9675	9680	92
93	9685	9689	9694	9699	9703	9708	9713	9717	9722	9727	93
94	9731	9736	9741	9745	9750	9754	9759	9763	9768	9773	94
95	9777	9782	9786	9791	9795	9800	9805	9809	9814	9818	95
96	9823	9827	9832	9836	9841	9845	9850	9854	9859	9863	96
97	9868	9872	9877	9881	9886	9890	9894	9899	9903	9908	97
98	9912	9917	9921	9926	9930	9934	9939	9943	9948	9952	98
99	9956	9961	9965	9969	9974	9978	9983	9987	9991	9996	99
N	0	1	2	3	4	5	6	7	8	9	N

Degrees	Sin	Cos	Tan	Cot	
0° 00′	0.0000	1.0000	0.0000	∞	90° 00′
10	.0029	1.0000	.0029	343.77	50
20	.0058	1.0000	.0058	171.89	40
30	.0087	1.0000	.0087	114.59	30
40	.0116	.9999	.0116	85.940	20
50	.0145	.9999	.0145	68.750	10
1° 00′	0.0175	0.9998	0.0175	57.290	89° 00′
10	.0204	.9998	.0204	49.104	50
20	.0233	.9997	.0233	42.964	40
30	.0262	.9997	.0262	38.188	30
40	.0291	.9996	.0291	34.368	20
50	.0320	.9995	.0320	31.242	10
2° 00′	0.0349	0.9994	0.0349	28.636	88° 00′
10	.0378	.9993	.0378	26.432	50
20	.0407	.9992	.0407	24.542	40
30	.0436	.9990	.0437	22.904	30
40	.0465	.9989	.0466	21.470	20
50	.0494	.9988	.0495	20.206	10
3° 00′	0.0523	0.9986	0.0524	19.081	87° 00′
10	.0552	.9985	.0553	18.075	50
20	.0581	.9983	.0582	17.169	40
30	.0610	.9981	.0612	16.350	30
40	.0640	.9980	.0641	15.605	20
50	.0669	.9978	.0670	14.924	10
4° 00′	0.0698	0.9976	0.0699	14.301	86° 00′
10	.0727	.9974	.0729	13.727	50
20	.0756	.9971	.0758	13.197	40
30	.0785	.9969	.0787	12.706	30
40	.0814	.9967	.0816	12.251	20
50	.0843	.9964	.0846	11.826	10
5° 00′	0.0872	0.9962	0.0875	11.430	85° 00′
10	.0901	.9959	.0904	11.059	50
20	.0929	.9957	.0934	10.712	40
30	.0958	.9954	.0963	10.385	30
40	.0987	.9951	.0992	10.078	20
50	.1016	.9948	.1022	9.7882	10
6° 00′	0.1045	0.9945	0.1051	9.5144	84° 00′
10	.1074	.9942	.1080	9.2553	50
20	.1103	.9939	.1110	9.0098	40
30	.1132	.9936	.1139	8.7769	30
40	.1161	.9932	.1169	8.5555	20
50	.1190	.9929	.1198	8.3450	10
7° 00′	0.1219	0.9925	0.1228	8.1443	83° 00′
10	.1248	.9922	.1257	7.9530	50
20	.1276	.9918	.1287	7.7704	40
30	.1305	.9914	.1317	7.5958	30
40	.1334	.9911	.1346	7.4287	20
50	.1363	.9907	.1376	7.2687	10
8° 00′	0.1392	0.9903	0.1405	7.1154	82° 00′
10	.1421	.9899	.1435	6.9682	50
20	.1449	.9894	.1465	6.8269	40
30	.1478	.9890	.1495	6.6912	30
40	.1507	.9886	.1524	6.5606	20
50	.1536	.9881	.1554	6.4348	10
9° 00′	0.1564	0.9877	0.1584	6.3138	81° 00′
10	.1593	.9872	.1614	6.1970	50
20	.1622	.9868	.1644	6.0844	40
30	.1650	.9863	.1673	5.9758	30
40	.1679	.9858	.1703	5.8708	20
50	.1708	.9853	.1733	5.7694	10
	Cos	Sin	Cot	Tan	Degrees

Fig. 25-210. Table of natural

Degrees	Sin	Cos	Tan	Cot	
10° 00′	0.1736	0.9848	0.1763	5.6713	80° 00′
10	.1765	.9843	.1793	5.5764	50
20	.1794	.9838	.1823	5.4845	40
30	.1822	.9833	.1853	5.3955	30
40	.1851	.9827	.1883	5.3093	20
50	.1880	.9822	.1914	5.2257	10
11° 00′	0.1908	0.9816	0.1944	5.1446	79° 00′
10	.1937	.9811	.1974	5.0658	50
20	.1965	.9805	.2004	4.9894	40
30	.1994	.9799	.2035	4.9152	30
40	.2022	.9793	.2065	4.8430	20
50	.2051	.9787	.2095	4.7729	10
12° 00′	0.2079	0.9781	0.2126	4.7046	78° 00′
10	.2108	.9775	.2156	4.6382	50
20	.2136	.9769	.2186	4.5736	40
30	.2164	.9763	.2217	4.5107	30
40	.2193	.9757	.2247	4.4494	20
50	.2221	.9750	.2278	4.3897	10
13° 00′	0.2250	0.9744	0.2309	4.3315	77° 00′
10	.2278	.9737	.2339	4.2747	50
20	.2306	.9730	.2370	4.2193	40
30	.2334	.9724	.2401	4.1653	30
40	.2363	.9717	.2432	4.1126	20
50	.2391	.9710	.2462	4.0611	10
14° 00′	0.2419	0.9703	0.2493	4.0108	76° 00′
10	.2447	.9696	.2524	3.9617	50
20	.2476	.9689	.2555	3.9136	40
30	.2504	.9681	.2586	3.8667	30
40	.2532	.9674	.2617	3.8208	20
50	.2560	.9667	.2648	3.7760	10
15° 00′	0.2588	0.9659	0.2679	3.7321	75° 00′
10	.2616	.9652	.2711	3.6891	50
20	.2644	.9644	.2742	3.6470	40
30	.2672	.9636	.2773	3.6059	30
40	.2700	.9628	.2805	3.5656	20
50	.2728	.9621	.2836	3.5261	10
16° 00′	0.2756	0.9613	0.2867	3.4874	74° 00′
10	.2784	.9605	.2899	3.4495	50
20	.2812	.9596	.2931	3.4124	40
30	.2840	.9588	.2962	3.3759	30
40	.2868	.9580	.2994	3.3402	20
50	.2896	.9572	.3026	3.3052	10
17° 00′	0.2924	0.9563	0.3057	3.2709	73° 00′
10	.2952	.9555	.3089	3.2371	50
20	.2979	.9546	.3121	3.2041	40
30	.3007	.9537	.3153	3.1716	30
40	.3035	.9528	.3185	3.1397	20
50	.3062	.9520	.3217	3.1084	10
18° 00′	0.3090	0.9511	0.3249	3.0777	72° 00′
10	.3118	.9502	.3281	3.0475	50
20	.3145	.9492	.3314	3.0178	40
30	.3173	.9483	.3346	2.9887	30
40	.3201	.9474	.3378	2.9600	20
50	.3228	.9465	.3411	2.9319	10
19° 00′	0.3256	0.9455	0.3443	2.9042	71° 00′
10	.3283	.9446	.3476	2.8770	50
20	.3311	.9436	.3508	2.8502	40
30	.3338	.9426	.3541	2.8239	30
40	.3365	.9417	.3574	2.7980	20
50	.3393	.9407	.3607	2.7725	10
	Cos	Sin	Cot	Tan	Degrees

trigonometric functions (Continued next page).

Degrees	Sin	Cos	Tan	Cot	
20° 00′	0.3420	0.9397	0.3640	2.7475	70° 00′
10	.3448	.9387	.3673	2.7228	50
20	.3475	.9377	.3706	2.6985	40
30	.3502	.9367	.3739	2.6746	30
40	.3529	.9356	.3772	2.6511	20
50	.3557	.9346	.3805	2.6279	10
21° 00′	0.3584	0.9336	0.3839	2.6051	69° 00′
10	.3611	.9325	.3872	2.5826	50
20	.3638	.9315	.3906	2.5605	40
30	.3665	.9304	.3939	2.5386	30
40	.3692	.9293	.3973	2.5172	20
50	.3719	.9283	.4006	2.4960	10
22° 00′	0.3746	0.9272	0.4040	2.4751	68° 00′
10	.3773	.9261	.4074	2.4545	50
20	.3800	.9250	.4108	2.4342	40
30	.3827	.9239	.4142	2.4142	30
40	.3854	.9228	.4176	2.3945	20
50	.3881	.9216	.4210	2.3750	10
23° 00′	0.3907	0.9205	0.4245	2.3559	67° 00′
10	.3934	.9194	.4279	2.3369	50
20	.3961	.9182	.4314	2.3183	40
30	.3987	.9171	.4348	2.2998	30
40	.4014	.9159	.4383	2.2817	20
50	.4041	.9147	.4417	2.2637	10
24° 00′	0.4067	0.9135	0.4452	2.2460	66° 00′
10	.4094	.9124	.4487	2.2286	50
20	.4120	.9112	.4522	2.2113	40
30	.4147	.9100	.4557	2.1943	30
40	.4173	.9088	.4592	2.1775	20
50	.4200	.9075	.4628	2.1609	10
25° 00′	0.4226	0.9063	0.4663	2.1445	65° 00′
10	.4253	.9051	.4699	2.1283	50
20	.4279	.9038	.4734	2.1123	40
30	.4305	.9026	.4770	2.0965	30
40	.4331	.9013	.4806	2.0809	20
50	.4358	.9001	.4841	2.0655	10
26° 00′	0.4384	0.8988	0.4877	2.0503	64° 00′
10	.4410	.8975	.4913	2.0353	50
20	.4436	.8962	.4950	2.0204	40
30	.4462	.8949	.4986	2.0057	30
40	.4488	.8936	.5022	1.9912	20
50	.4514	.8923	.5059	1.9768	10
27° 00′	0.4540	0.8910	0.5095	1.9626	63° 00′
10	.4566	.8897	.5132	1.9486	50
20	.4592	.8884	.5169	1.9347	40
30	.4617	.8870	.5206	1.9210	30
40	.4643	.8857	.5243	1.9074	20
50	.4669	.8843	.5280	1.8940	10
28° 00′	0.4695	0.8829	0.5317	1.8807	62° 00′
10	.4720	.8816	.5354	1.8676	50
20	.4746	.8802	.5392	1.8546	40
30	.4772	.8788	.5430	1.8418	30
40	.4797	.8774	.5467	1.8291	20
50	.4823	.8760	.5505	1.8165	10
29° 00′	0.4848	0.8746	0.5543	1.8040	61° 00′
10	.4874	.8732	.5581	1.7917	50
20	.4899	.8718	.5619	1.7796	40
30	.4924	.8704	.5658	1.7675	30
40	.4950	.8689	.5696	1.7556	20
50	.4975	.8675	.5735	1.7437	10
	Cos	Sin	Cot	Tan	Degrees

Fig. 25-210. Table of natural

Degrees	Sin	Cos	Tan	Cot	
30° 00′	0.5000	0.8660	0.5774	1.7321	60° 00′
10	.5025	.8646	.5812	1.7205	50
20	.5050	.8631	.5851	1.7090	40
30	.5075	.8616	.5890	1.6977	30
40	.5100	.8601	.5930	1.6864	20
50	.5125	.8587	.5969	1.6753	10
31° 00′	0.5150	0.8572	0.6009	1.6643	59° 00′
10	.5175	.8557	.6048	1.6534	50
20	.5200	.8542	.6088	1.6426	40
30	.5225	.8526	.6128	1.6319	30
40	.5250	.8511	.6168	1.6212	20
50	.5275	.8496	.6208	1.6107	10
32° 00′	0.5299	0.8480	0.6249	1.6003	58° 00′
10	.5324	.8465	.6289	1.5900	50
20	.5348	.8450	.6330	1.5798	40
30	.5373	.8434	.6371	1.5697	30
40	.5398	.8418	.6412	1.5597	20
50	.5422	.8403	.6453	1.5497	10
33° 00′	0.5446	0.8387	0.6494	1.5399	57° 00′
10	.5471	.8371	.6536	1.5301	50
20	.5495	.8355	.6577	1.5204	40
30	.5519	.8339	.6619	1.5108	30
40	.5544	.8323	.6661	1.5013	20
50	.5568	.8307	.6703	1.4919	10
34° 00′	0.5592	0.8290	0.6745	1.4826	56° 00′
10	.5616	.8274	.6787	1.4733	50
20	.5640	.8258	.6830	1.4641	40
30	.5664	.8241	.6873	1.4550	30
40	.5688	.8225	.6916	1.4460	20
50	.5712	.8208	.6959	1.4370	10
35° 00′	0.5736	0.8192	0.7002	1.4281	55° 00′
10	.5760	.8175	.7046	1.4193	50
20	.5783	.8158	.7089	1.4106	40
30	.5807	.8141	.7133	1.4019	30
40	.5831	.8124	.7177	1.3934	20
50	.5854	.8107	.7221	1.3848	10
36° 00′	0.5878	0.8090	0.7265	1.3764	54° 00′
10	.5901	.8073	.7310	1.3680	50
20	.5925	.8056	.7355	1.3597	40
30	.5948	.8039	.7400	1.3514	30
40	.5972	.8021	.7445	1.3432	20
50	.5995	.8004	.7490	1.3351	10
37° 00′	.6018	.7986	.7536	1.3270	53° 00′
10	.6041	.7969	.7581	1.3190	50
20	.6065	.7951	.7627	1.3111	40
30	.6088	.7934	.7673	1.3032	30
40	.6111	.7916	.7720	1.2954	20
50	.6134	.7898	.7766	1.2876	10
38° 00′	0.6157	0.7880	0.7813	1.2799	52° 00′
10	.6180	.7862	.7860	1.2723	50
20	.6202	.7844	.7907	1.2647	40
30	.6225	.7826	.7954	1.2572	30
40	.6248	.7808	.8002	1.2497	20
50	.6271	.7790	.8050	1.2423	10
39° 00′	0.6293	0.7771	0.8098	1.2349	51° 00′
10	.6316	.7753	.8146	1.2276	50
20	.6338	.7735	.8195	1.2203	40
30	.6361	.7716	.8243	1.2131	30
40	.6383	.7698	.8292	1.2059	20
50	.6406	.7679	.8342	1.1988	10
	Cos	Sin	Cot	Tan	Degrees

trigonometric functions (Continued next page).

Degrees	Sin	Cos	Tan	Cot	
40° 00′	0.6428	0.7660	0.8391	1.1918	50° 00′
10	.6450	.7642	.8441	1.1847	50
20	.6472	.7623	.8491	1.1778	40
30	.6494	.7604	.8541	1.1708	30
40	.6517	.7585	.8591	1.1640	20
50	.6539	.7566	.8642	1.1571	10
41° 00′	0.6561	0.7547	0.8693	1.1504	49° 00′
10	.6583	.7528	.8744	1.1436	50
20	.6604	.7509	.8796	1.1369	40
30	.6626	.7490	.8847	1.1303	30
40	.6648	.7470	.8899	1.1237	20
50	.6670	.7451	.8952	1.1171	10
42° 00′	0.6691	0.7431	0.9004	1.1106	48° 00′
10	.6713	.7412	.9057	1.1041	50
20	.6734	.7392	.9110	1.0977	40
30	.6756	.7373	.9163	1.0913	30
40	.6777	.7353	.9217	1.0850	20
50	.6799	.7333	.9271	1.0786	10
43° 00′	0.6820	0.7314	0.9325	1.0724	47° 00′
10	.6841	.7294	.9380	1.0661	50
20	.6862	.7274	.9435	1.0599	40
30	.6884	.7254	.9490	1.0538	30
40	.6905	.7234	.9545	1.0477	20
50	.6926	.7214	.9601	1.0416	10
44° 00′	0.6947	0.7193	0.9657	1.0355	46° 00′
10	.6967	.7173	.9713	1.0295	50
20	.6988	.7163	.9770	1.0235	40
30	.7009	.7133	.9827	1.0176	30
40	.7030	.7112	.9884	1.0117	20
50	.7050	.7092	.9942	1.0058	10
45° 00′	0.7071	0.7071	1.0000	1.0000	45° 00′
	Cos	Sin	Cot	Tan	Degrees

Fig. 25-210. Table of natural trigonometric functions (Cont'd).

The letter appearing before the slash (/) is a specification modification.

25.213 Give an energy-level chart. —Energy levels are difficult to conceive without comparing energy levels from one scientific field to another. The energy chart prepared by J. R. Williams, Air Force Special Weapons Center (ARDC), appears in Fig. 25-213 and is designed to relate several commonly used units with that of real life. It will be observed, moving downward along the chart, that the energy level increases. Horizontally it is constant. For example, the energy equivalent of 1 gram of matter is 9×10^{20} ergs, which is approximately 2.5×10^{10} watt-hours, or the energy from 20 kilotons of TNT exploded, or the energy released in complete fission of 1 kilogram of radio active uranium U-235.

25.214 Give a decibel ratio table for gain or loss of voltage and power.—To simplify the computation for equivalent number of decibels for loss or gain ratios of power or voltage, the table in Fig. 25-214 may be used as follows:

(a) To find current or voltage loss or gain equivalent to a given number of decibels, find the required number of decibels in the Decibel Voltage column and read the corresponding ratio in loss or gain column.

(b) To find power loss or gain ratio equivalent to a given number of decibels, find the required number of decibels in the decibel power column and read the corresponding ratio in the loss or gain column.

(c) To find the number of decibels equivalent to a given loss or gain, find the required ratio in loss or gain, and read the corresponding number of decibels in the Decibel Voltage column.

(d) To find the number of decibels equivalent to a given power loss or gain, find the required ratio in the loss or gain column and read the corresponding number of decibels in the Decibel Power column.

$$\text{Decibels} = 10 \text{ Log}_{10} (P_1/P_2)$$

Material	Dielectric Constant (Approx.)	Material	Dielectric Constant (Approx.)
Air	1.0	Nylon	3.4-22.4
Amber	2.6-2.7	Paper (dry)	1.5-3.0
Bakelite (asbestos base)	5.0-22	Paper (paraffin coated)	2.5-4.0
Bakelite (mica filled)	4.5-4.8	Paraffin (solid)	2.0-3.0
Beeswax	2.4-2.8	Plexiglass	2.6-3.5
Cambric (varnished)	4.0	Polyethylene	2.3
Celluloid	4.0	Polystyrene	2.4-3.0
Cellulose Acetate	3.1-4.5	Porcelain (dry process)	5.0-5.5
Durite	4.7-5.1	Porcelain (wet process)	5.8-6.5
Ebonite	2.7	Quartz	5.0
Fiber	5.0	Quartz (fused)	3.78
Formica	3.6-6.0	Rubber (hard)	2.0-4.0
Glass (electrical)	3.8-14.5	Ruby Mica	5.4
Glass (photographic)	7.5	Shellac (natural)	2.9-3.9
Glass (Pyrex)	4.6-5.0	Silicone (glass) (molding)	3.2-4.7
Glass (window)	7.6	Silicone (glass) (laminate)	3.7-4.3
Gutta Percha	2.4-2.6	Slate	7.0
Isolantite	6.1	Steatite (ceramic)	5.2-6.3
Lucite	2.5	Steatite (low loss)	4.4
Mica (electrical)	4.0-9.0	Styrofoam	1.03
Mica (clear India)	7.5	Teflon	2.1
Mica (filled phenolic)	4.2-5.2	Vaseline	2.16
Micarta	3.2-5.5	Vinylite	2.7-7.5
Mycalex	7.3-9.3	Water (distilled)	34-78
Neoprene	4.0-6.7	Wood (dry)	1.4-2.9

Fig. 25-211. Dielectric constants of materials.

If the two voltages or currents under consideration are at the same impedance level, then:

$$\text{Decibels} = 20 \text{ Log}_{10} (E_1/E_2)$$

For values outside the range of the table, the rules that appear below the table apply.

25.215 ***Describe a device using ultrasonic sound waves for detecting flaws in metal.***—The use of ultrasonic sound waves has great industrial importance in the detection of flaws in castings, nonbonds, welding, and biological examinations. Although ultrasonic sound waves differ radically from electromagnetic waves, such as visible light, they can be manipulated in similar ways. They can be reflected, refracted, scattered, and absorbed, and the same physical laws apply approximately to both types of waveforms. This has led to the development of an ultrasonic camera, which has the capability of changing an ultrasonic image to an easily recognizable visual form. A device of this type, developed by James Electronics, Inc., with a tradename of *Ultra-Scan* is pictured in Fig. 25-215A.

All sound waves are basically a series of compressions and rarefactions moving through a medium, and unlike electromagnetic radiations they cannot travel through a vacuum. Ultrasonic frequencies range from 0.016 MHz to 10 MHz, and are used in nondestruct testing of various materials.

An ultrasonic waveform passing through a material may be attenuated in two ways: by reflection and refraction, and by scattering from discontinuities. Its energy may also be absorbed by the material. Similarly, a porous material gradually scatters the main portion of the beam and reduces its energy.

The basic principle of the James Electronics system is given in Fig. 25-215B. It is based on the idea that if an extended area of piezoelectric material is subjected to sound energy at a point on its entire surface, a potential will be produced at that point rather than over the entire surface. If this surface is scanned with a high-velocity electron beam, it emits secondary electrons which are displayed on a CRT. The important points of this system are that a secondary electron beam is modulated, and its intensity varies with the potential at each point on the piezoelectric target. Thus, ultrasonic energy passing through an object will produce varying potential in the surface of the piezoelec-

Type RG.../U	Imp. (ohms)	Cap. (mmf per ft.)	Diam. (inches)	Attenuation—db per 100 ft. 1 mc	10 mc	100 mc	400 mc	1000 mc	REMARKS
5	52.5	28.5	.332	.21	.77	2.9	6.5	11.5	Small, double braid
5A	50	29	.328	.16	.66	2.4	5.25	8.8	Small, low loss
6	76	20	.332	.21	.78	2.9	6.5	11.2	IF & video
8	52	29.5	.405	.16	.55	2.0	4.5	8.5	General purpose
9	51	30	.420	.12	.47	1.9	4.4	8.5	General purpose
9A	51	30	.420	.16	.59	2.3	5.2	8.6	Stable attenuation
11	75	20.5	.405	.18	.62	2.2	4.7	8.2	Community TV
13	74	20.5	.420	.18	.62	2.2	4.7	8.2	IF
14	52	29.5	.545	.10	.38	1.5	3.5	6.0	RF power
16	52	29.5	.630	—	—	—	—	—	RF power
17	52	29.5	.870	.06	.24	.95	2.4	4.4	RF power
19	52	29.5	1.120	.04	.17	.68	1.28	3.5	Low-loss RF
21	53	29	.332	1.4	4.4	14.0	29.0	46.0	Attenuating cable
22	95	16	.405	.41	1.3	4.3	8.8	—	Twin conductors
23	125	12	.65 X .945	—	.4	1.7	—	—	Twin conductors (balanced)
25	48	50	.565	—	—	—	—	—	Pulse
26	48	50	.525	—	—	—	—	—	Pulse
27	48	50	.675	—	—	—	—	—	Pulse
28	48	50	.805	—	—	—	—	—	Pulse
33	51	30	.470	—	—	—	—	—	Pulse
34	71	21.5	.625	.065	.29	1.3	3.3	6.0	Flexible, medium
35	71	21.5	.945	.064	.22	.85	2.3	4.2	Low-loss video
36	69	22	1.180	—	—	—	—	—	—
41	67.5	27	.425	—	—	—	—	—	Special twist
54A	58	26.5	.250	.18	.74	3.1	6.7	11.5	Flexible, small
55	53.5	28.5	.206	.36	1.3	4.8	10.4	17.0	Flexible, small
56	—	—	.535	—	—	—	—	—	Pulse
57	95	17	.625	.18	.71	3.0	7.3	13.0	Twin conductors
58	53.5	30	.195	.38	1.4	5.2	11.2	20.0	General purpose
58A	50	30	.195	.42	1.6	6.2	14.0	24.0	Test leads
59	73	21	.242	.30	1.1	3.8	8.5	14.0	TV lead-in
60	50	—	.425	—	—	—	—	—	Pulse cable
61	500	—	—	—	—	—	—	—	Special 500-ohm twin-lead
62	93	13.5	.242	.25	.83	2.7	5.6	9.0	Low capacity, small
63	125	10	.405	.19	.61	2.0	4.0	6.3	Low capacity
64	48	50	.495	—	—	—	—	—	Pulse
65	950	44	.405	—	—	—	—	—	Coaxial delay line
71	93	13.5	.250	.25	.83	2.7	5.6	9.0	Low capacity, small
77	48	50	.415	—	—	—	—	—	Pulse
78	48	50	.385	—	—	—	—	—	Pulse
87A	50	29.5	.425	.13	.52	2.0	4.4	7.6	Teflon dielectric
88	48	50	.490	—	—	—	—	—	Pulse
101	75	—	.588	—	—	—	—	—	—
102	140	—	1.088	—	—	—	—	—	—
108	76	25	.245	—	—	—	—	—	Twin conductors
114	185	6.5	.405	—	—	—	—	—	Extra flexible
117	50	29	.730	.05	.20	.85	2.0	3.6	Teflon & Fiberglas
119	50	29	.470	—	—	—	—	—	Teflon & Fiberglas
122	50	29.3	.160	.40	1.70	7.0	16.5	29.0	—
126	50	29	.290	3.20	9.0	25.0	47.0	72.0	Teflon & Fiberglas
140	73	21	.242	.33	1.03	3.3	6.9	11.7	Teflon & Fiberglas
141	50	29	.195	.35	1.12	3.8	8.0	13.8	Teflon & Fiberglas
142	50	29	.206	.35	1.12	3.8	8.0	13.8	Teflon & Fiberglas
143	50	29	.325	.24	.77	2.5	5.3	9.0	Teflon & Fiberglas
144	72	21	.395	.16	.53	1.8	3.9	7.0	Teflon & Fiberglas
174	50	30	.10	—	—	—	19.0	—	Miniature coaxial

Fig. 25-212. Characteristics of coaxial cable.

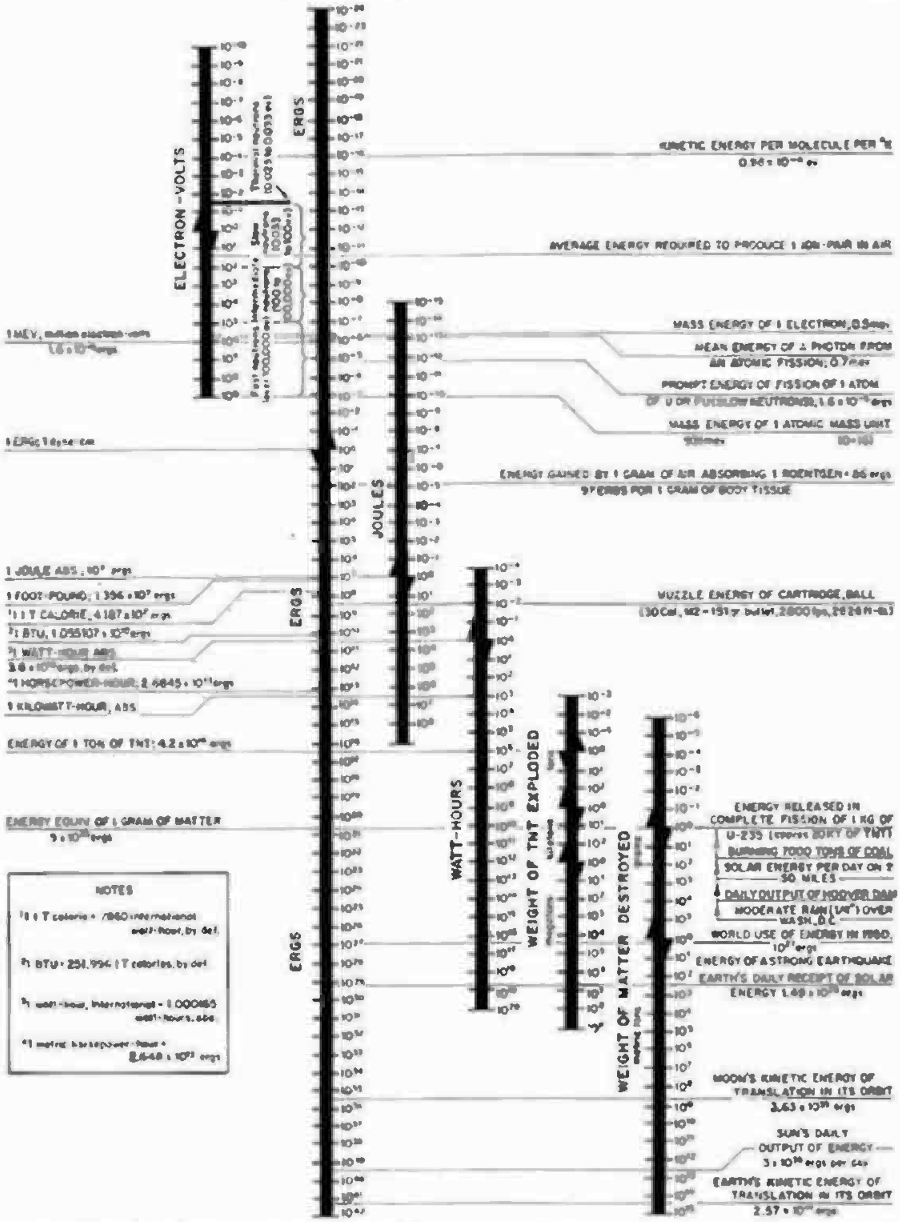


Fig. 25-213. Energy level chart. After J. R. Williams. (Courtesy, Missiles and Rockets.)

tric target, characteristic of the flaw of the object. The final display on the CRT is a visual presentation of the flaw. Flaws on the order of 0.010 inch may be thoroughly analyzed.

The Ultra-Scan, pictured in Fig. 25-215A consists of a control console, a sealed tube console, and a specimen tank. The system may be operated at frequencies between 1 and 10 MHz. Typical images as seen on the CRT are shown in Figs. 25-215C and D.

When the system is used for biological research, it is possible to identify the differences between blood and tissue, and between different types of tissue. Frozen tissues are easily identified. Cinesonograph (motion picture) of living tissue is also possible.

The system described can be used to nondestruct test primary metals and many nonmetallic materials for surface, subsurface, deep-seated cracks, hole porosity differeuces, foreign inclusion, flow and stress patterns, laminal flow in fluids, temperature differences, homogenity in potted assemblies, and many others.

25.216 ***Describe an ultrasonic thickness-measuring gauge.***—Ultrasonic thickness-measuring gauges utilize the resonance principle, responding to mechanical resonant frequencies provided by the test sample. An ultrasonic thickness gauge, Model SO-300, manufactured by Magnaflux Corp., is shown in Figs. 25-216A and B.

The instrument contains an ultrasonic oscillator which generates a continuously varying frequency. This frequency is fed to a crystal transducer. Ultrasonic vibrations are emitted by the crystal into the sample under test, through a coupling medium such as oil, glycerin, or a soap film. The mechanical resonant frequency of the test sample varies with its thickness. When the constantly varying frequency hits this resonance, there is an electrical change within the crystal which is sensed by the instrument and displayed as a logarithmically spaced pattern of neon lights in a viewing window, covering an arc of 180 degrees. A logarithmic function generator provides the basis for the direct readout capabilities.

Three overlapping ranges provide measurements ranging from 0.025 to 3.0 inches. To operate, the operator selects the correct range and applies a liquid couplant to the material to be measured, and positions the transducer. A harmonic scale disc is then rotated until the division lines accurately coincide with a flashing light pattern. A circular calibrated scale on the disc interprets the flashes in terms of the thickness of the material tested.

A number of transducers in various shapes are available for testing a variety

Decibel (Voltage)	Loss	Gain	Decibel (Power)
.0	1.0000	1.000	.0
.1	.9886	1.012	.05
.2	.9772	1.023	.10
.3	.9661	1.035	.15
.4	.9550	1.047	.20
.5	.9441	1.059	.25
.6	.9333	1.072	.30
.7	.9226	1.084	.35
.8	.9120	1.096	.40
.9	.9016	1.109	.45
1.0	.8913	1.122	.50
.1	.8810	1.135	.55
.2	.8710	1.148	.60
.3	.8610	1.161	.65
.4	.8511	1.175	.70
.5	.8414	1.189	.75
.6	.8318	1.202	.80
.7	.8222	1.216	.85
.8	.8128	1.230	.90
.9	.8035	1.245	.95
2.0	.7943	1.259	1.00
.1	.7852	1.274	.05
.2	.7762	1.288	.10
.3	.7674	1.303	.15
.4	.7586	1.318	.20
.5	.7499	1.334	.25
.6	.7413	1.349	.30
.7	.7328	1.365	.35
.8	.7244	1.380	.40
.9	.7161	1.396	.45
3.0	.7079	1.413	.50
.1	.6998	1.429	.55
.2	.6918	1.445	.60
.3	.6839	1.462	.65
.4	.6761	1.479	.70
.5	.6683	1.496	.75
.6	.6607	1.514	.80
.7	.6531	1.531	.85
.8	.6457	1.549	.90
.9	.6383	1.567	.95
4.0	.6310	1.585	2.00
.1	.6237	1.603	.05
.2	.6166	1.622	.10
.3	.6095	1.641	.15
.4	.6026	1.660	.20
.5	.5957	1.679	.25
.6	.5888	1.698	.30
.7	.5821	1.718	.35
.8	.5754	1.738	.40
.9	.5689	1.758	.45

Decibel (Voltage)	Loss	Gain	Decibel (Power)
5.0	.5623	1.778	.50
.1	.5559	1.799	.55
.2	.5495	1.820	.60
.3	.5433	1.841	.65
.4	.5370	1.862	.70
.5	.5309	1.884	.75
.6	.5248	1.905	.80
.7	.5188	1.928	.85
.8	.5129	1.950	.90
.9	.5070	1.972	.95
6.0	.5012	1.995	3.00
.1	.4955	2.018	.05
.2	.4898	2.042	.10
.3	.4842	2.065	.15
.4	.4786	2.089	.20
.5	.4732	2.113	.25
.6	.4677	2.138	.30
.7	.4624	2.163	.35
.8	.4571	2.188	.40
.9	.4519	2.213	.45
7.0	.4467	2.239	.50
.1	.4416	2.265	.55
.2	.4365	2.291	.60
.3	.4315	2.317	.65
.4	.4266	2.344	.70
.5	.4217	2.371	.75
.6	.4169	2.399	.80
.7	.4121	2.427	.85
.8	.4074	2.455	.90
.9	.4027	2.483	.95
8.0	.3981	2.512	4.00
.1	.3936	2.541	.05
.2	.3890	2.570	.10
.3	.3846	2.600	.15
.4	.3802	2.630	.20
.5	.3758	2.661	.25
.6	.3715	2.692	.30
.7	.3673	2.723	.35
.8	.3631	2.754	.40
.9	.3589	2.786	.45
9.0	.3548	2.818	.50
.1	.3508	2.851	.55
.2	.3467	2.884	.60
.3	.3428	2.917	.65
.4	.3388	2.951	.70
.5	.3350	2.985	.75
.6	.3311	3.020	.80
.7	.3273	3.055	.85
.8	.3236	3.090	.90
.9	.3199	3.126	.95

Decibel (Voltage)	Loss	Gain	Decibel (Power)
10.0	.3162	3.162	5.00
.1	.3126	3.199	.05
.2	.3090	3.236	.10
.3	.3055	3.273	.15
.4	.3020	3.311	.20
.5	.2985	3.350	.25
.6	.2951	3.388	.30
.7	.2917	3.428	.35
.8	.2884	3.467	.40
.9	.2851	3.508	.45
11.0	.2818	3.548	.50
.1	.2786	3.589	.55
.2	.2754	3.631	.60
.3	.2723	3.673	.65
.4	.2692	3.715	.70
.5	.2661	3.758	.75
.6	.2630	3.802	.80
.7	.2600	3.846	.85
.8	.2570	3.890	.90
.9	.2541	3.936	.95
12.0	.2512	3.981	6.00
.1	.2483	4.027	.05
.2	.2455	4.074	.10
.3	.2427	4.121	.15
.4	.2399	4.169	.20
.5	.2371	4.217	.25
.6	.2344	4.266	.30
.7	.2317	4.315	.35
.8	.2291	4.365	.40
.9	.2265	4.416	.45
13.0	.2239	4.467	.50
.1	.2213	4.519	.55
.2	.2188	4.571	.60
.3	.2163	4.624	.65
.4	.2138	4.677	.70
.5	.2113	4.732	.75
.6	.2089	4.786	.80
.7	.2065	4.842	.85
.8	.2042	4.898	.90
.9	.2018	4.955	.95
14.0	.1995	5.012	7.00
.1	.1972	5.070	.05
.2	.1950	5.129	.10
.3	.1928	5.188	.15
.4	.1905	5.248	.20
.5	.1884	5.309	.25
.6	.1862	5.370	.30
.7	.1841	5.433	.35
.8	.1820	5.495	.40
.9	.1799	5.559	.45

Decibel (Voltage)	Loss	Gain	Decibel (Power)
15.0	.1778	5.623	.50
.1	.1758	5.689	.55
.2	.1738	5.754	.60
.3	.1718	5.821	.65
.4	.1698	5.888	.70
.5	.1679	5.957	.75
.6	.1660	6.026	.80
.7	.1641	6.095	.85
.8	.1622	6.166	.90
.9	.1603	6.237	.95
16.0	.1585	6.310	8.00
.1	.1567	6.383	.05
.2	.1549	6.457	.10
.3	.1531	6.531	.15
.4	.1514	6.607	.20
.5	.1496	6.683	.25
.6	.1479	6.761	.30
.7	.1462	6.839	.35
.8	.1445	6.918	.40
.9	.1429	6.998	.45
17.0	.1413	7.079	.50
.1	.1396	7.161	.55
.2	.1380	7.244	.60
.3	.1365	7.328	.65
.4	.1349	7.413	.70
.5	.1334	7.499	.75
.6	.1318	7.586	.80
.7	.1303	7.674	.85
.8	.1288	7.762	.90
.9	.1274	7.852	.95
18.0	.1259	7.943	9.00
.1	.1245	8.035	.05
.2	.1230	8.128	.10
.3	.1216	8.222	.15
.4	.1202	8.318	.20
.5	.1189	8.414	.25
.6	.1175	8.511	.30
.7	.1161	8.610	.35
.8	.1148	8.710	.40
.9	.1135	8.810	.45
19.0	.1122	8.913	.50
.1	.1109	9.016	.55
.2	.1096	9.120	.60
.3	.1084	9.226	.65
.4	.1072	9.333	.70
.5	.1059	9.441	.75
.6	.1047	9.550	.80
.7	.1035	9.661	.85
.8	.1023	9.772	.90
.9	.1012	9.886	.95

Decibel (Voltage)	Loss	Gain	Decibel (Power)
20.0	.1000	10.00	10.00
	Use the same numbers as 0-20 Db., but shift point one step to the left. Thus since 10 Db. = .3162 30 Db. = .03162	Use the same numbers as 0-20 Db., but shift point one step to the right. Thus since 10 Db. = 3.162 30 Db. = 31.62	This column repeats every 10 Db. instead of every 20 Db.
40.0	.01	100	20.00
	Use the same numbers as 0-20 Db., but shift point two steps to the left. Thus since 10 Db. = .3162 50 Db. = .003162	Use the same numbers as 0-20 Db., but shift point two steps to the right. Thus since 10 Db. = 3.162 50 Db. = 316.2	This column repeats every 10 Db. instead of every 20 Db.
60.0	.001	1,000	30.00
	Use the same numbers as 0-20 Db., but shift point three steps to the left. Thus since 10 Db. = .3162 70 Db. = .0003162	Use the same numbers as 0-20 Db., but shift point three steps to the right. Thus since 10 Db. = 3.162 70 Db. = 3162	This column repeats every 10 Db. instead of every 20 Db.
80.0	.0001	10,000	40.00
	Use the same numbers as 0-20 Db., but shift point four steps to the left. Thus since 10 Db. = .3162 90 Db. = .00003162	Use the same numbers as 0-20 Db., but shift point four steps to the right. Thus since 10 Db. = 3.162 90 Db. = 31620	This column repeats every 10 Db. instead of every 20 Db.
100.0	.00001	100,000	50.00

Fig. 25-214. Relationship between decibels, currents, voltage and power ratios.

Fig. 25-215A. James Electronics, Inc. Ultra-Scan ultrasonic camera system.

of materials and shapes. They may also be used for detecting casting core shift and locating lack of bond in laminated parts. The accuracy of the measurement is 1 percent or better.

25.217 ***Describe the basic principles of an ultrasonic cleaning device.***—Ultrasonic cleaning units consist of a tank containing a cleaning solution agitated by a transducer energized from an oscillator operating in the range of 20 to 40 kHz. Small cleaning units generally hold about one gallon of solution and are driven from an oscillator with 35 to 50 watts of output. A 75-gallon tank takes about 3000 watts of power for its operation.

The transducer unit is cemented or bolted to the tank, either internally or externally. The oscillator signal is converted to mechanical vibration by the

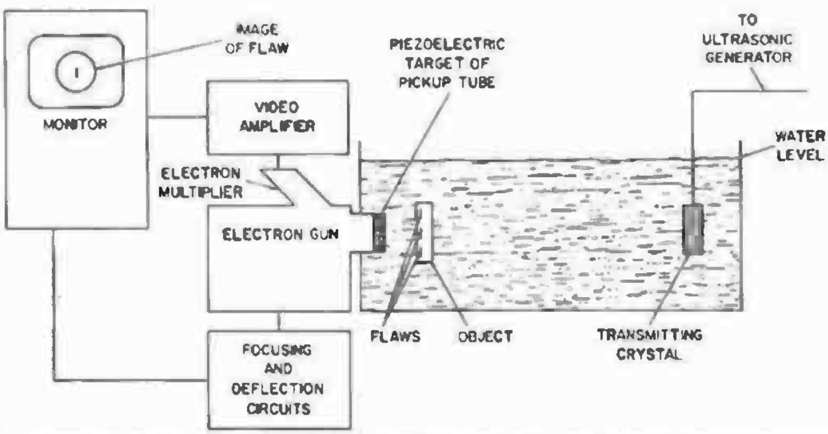


Fig. 25-215B. Basic principle of James Electronics Inc. Ultra-Scan ultrasonic scanning system for detecting flaws in castings and many other materials.

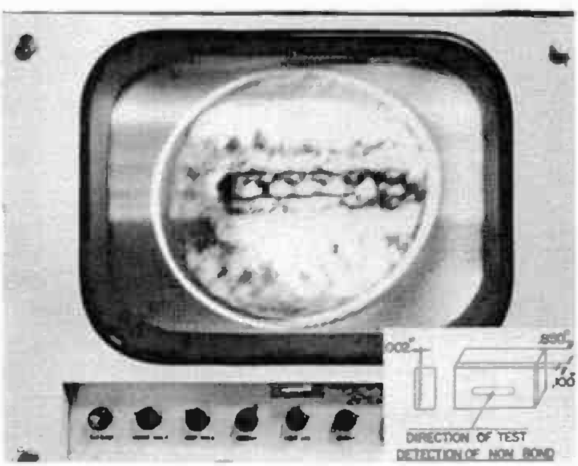


Fig. 25-215C. Ultra-Scan picture of a nonbond. Frequency, 2 MHz. Material, aluminum. The nonbond area is indicated by the dark pattern in the center of the display.

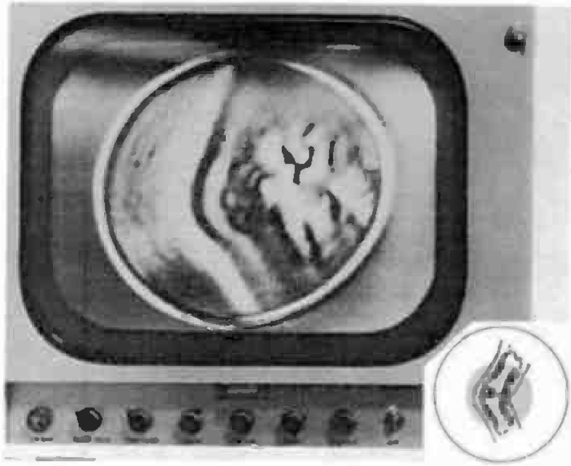


Fig. 25-215D. Ultra-Scan display of a biological examination. Frequency, 2 MHz. Material is bone, blood, and tissue.

Fig. 25-216A. Model-SO-300 ultrasonic thickness-measuring gauge manufactured by Magnaflux Corp.

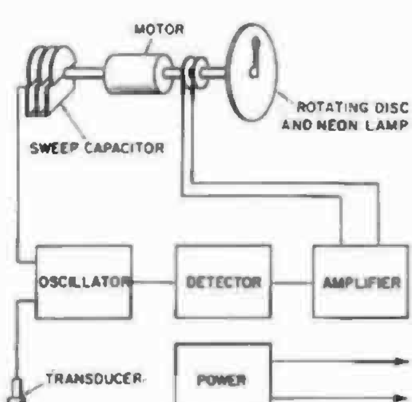


Fig. 25-216B. Basic circuit for Magnaflux ultrasonic thickness-measuring gauge.

transducer and is transmitted to the cleaning solution to cause cavitation, which blasts off the dirt from the surface being cleaned. The use of ultrasonic frequency is not confined to cleaning, but also is made in drilling, in inspection of fractures in metal, in biology, and in many other fields.

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